ECE 202A notes set 8.

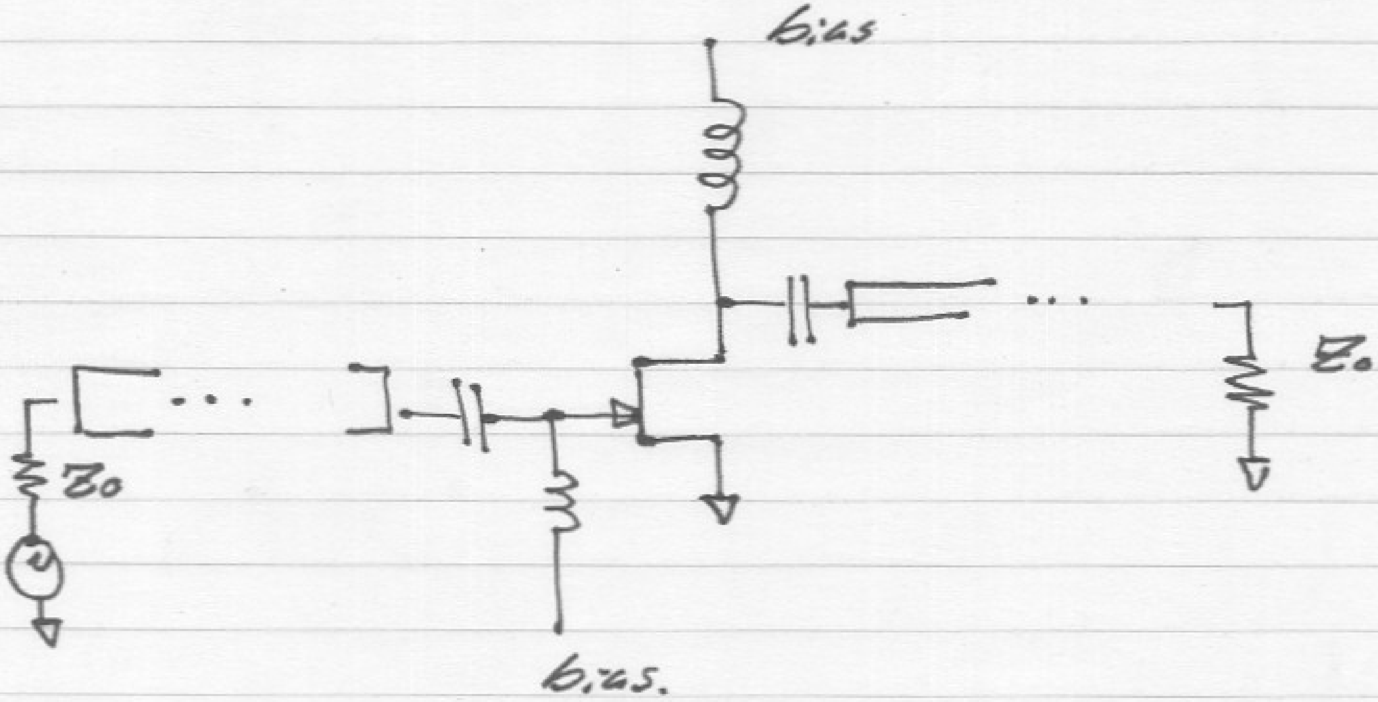
Resistive Amplifier designs

First we must define objectives:

an example will be helpful.

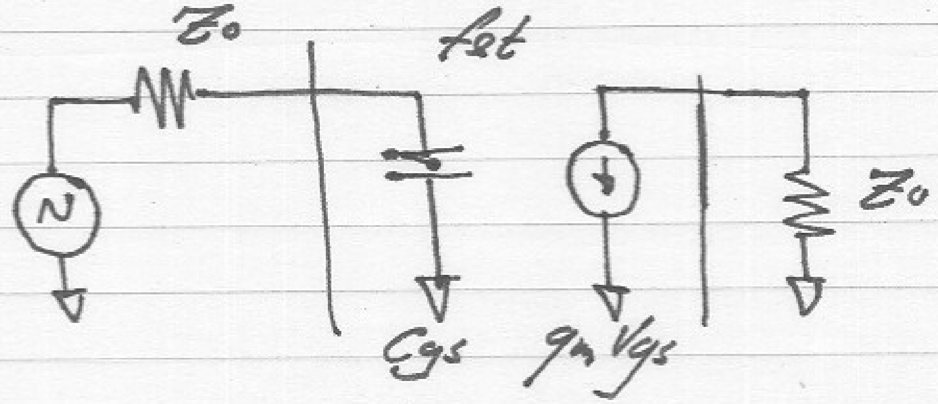
Example:

The worlds' simplest amplifier



If we remove the bias elements

& transmission lines, we obtain



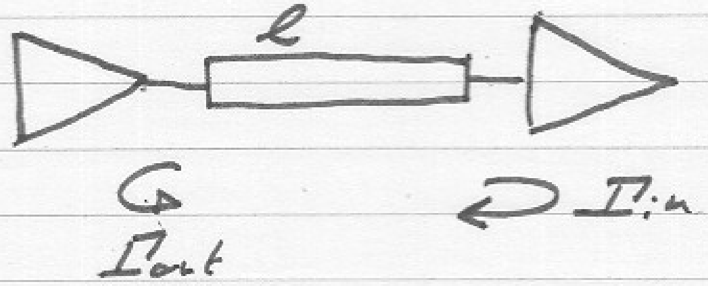
$$\hookrightarrow S_{21} = 2 \cdot \frac{-g_m Z_o}{1 + j\omega Z_o C_{gs}}.$$

So, what is wrong with this amplifier?

Certainly the lack of matching networks reduces the gain, but so what? - If the gain is adequate, I am happy.

More seriously, the input & output reflections will result in standing waves and frequency-response fluctuations.

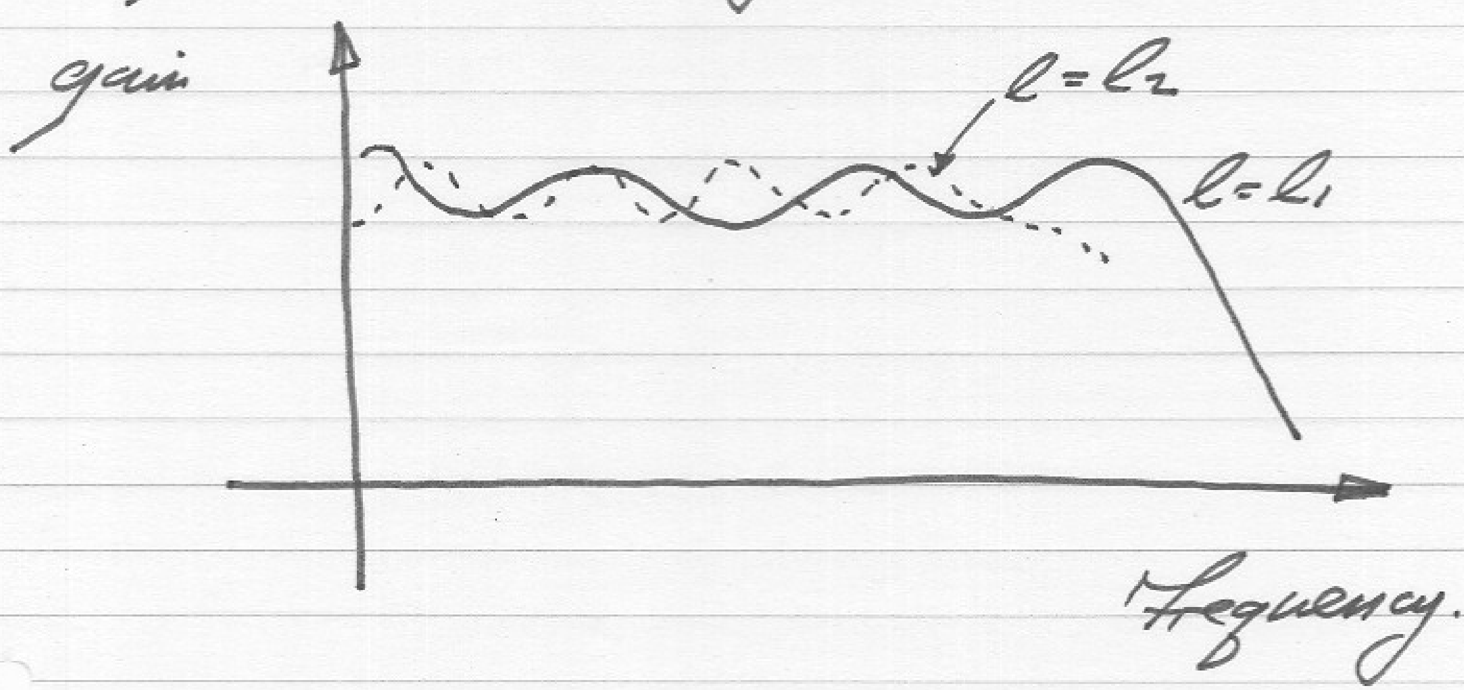
Consider two connected amplifiers:



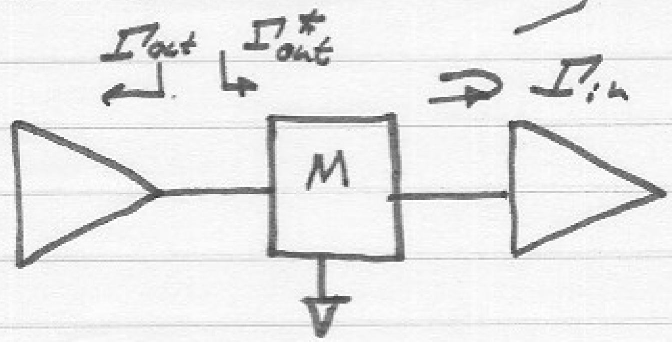
the overall gain will involve a term

$$\frac{1}{1 - \exp(2j\beta l)\,\Gamma_{in}\Gamma_{out}} \qquad \beta = \omega / v$$

which will ~~?~~ vary with both length and frequency:

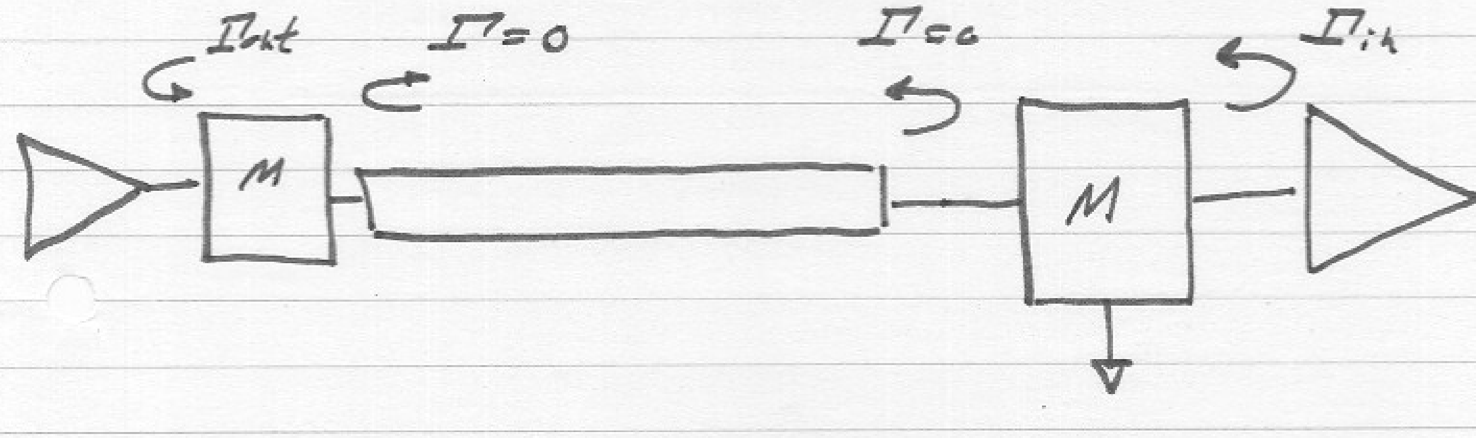


so while we are allowed to directly match 2 circuits together thus:



this interconnect/match strategy is only acceptable if the line lengths are short and the connection permanent.

If the line lengths are long, or if the connections changable (modules) then we must match to 50Ω, thus:



So *any* circuit with an input/output port ~~is~~ must have S_{11} or S_{22} $= 0$ e.g. be impedance-matched.

Two kinds of matches:

Reactive Impedance Matching (lossless)

Reflection coefficients $\rightarrow 0$

Gain $\rightarrow$ maximum available (maybe)

Resistive Impedance Matching (lossy)

Reflection coefficients $\rightarrow 0$

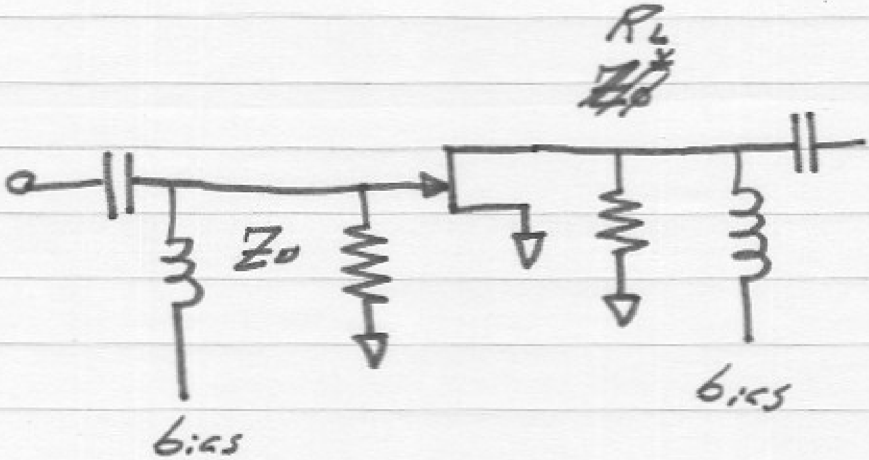
Power wasted in termination elements

Gain less than MAG.

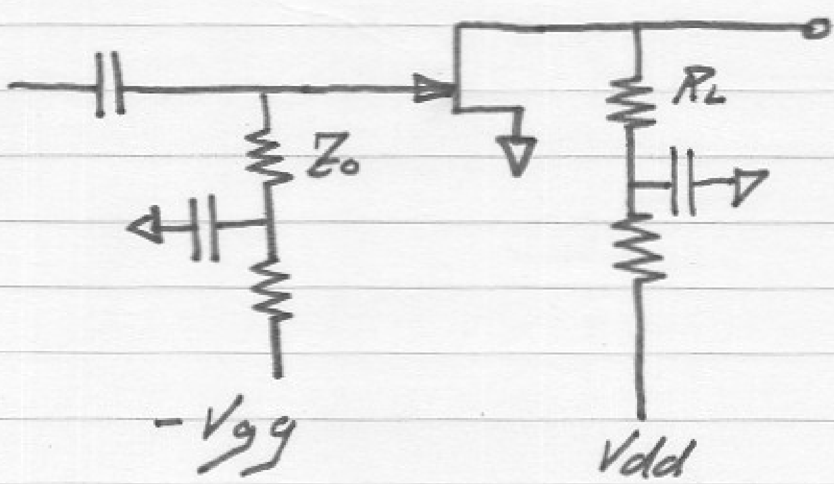
This leads us to:

<u>The worlds second-simplest amplifier:</u>

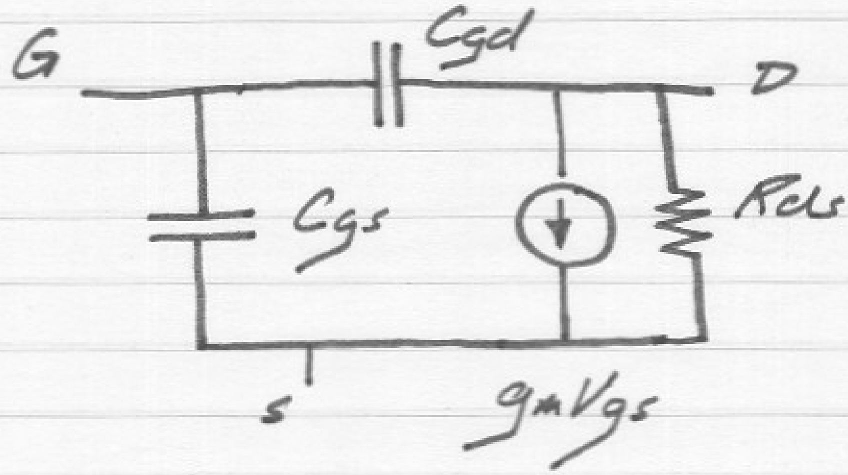
Common-source stage with 50-Ω terminations.



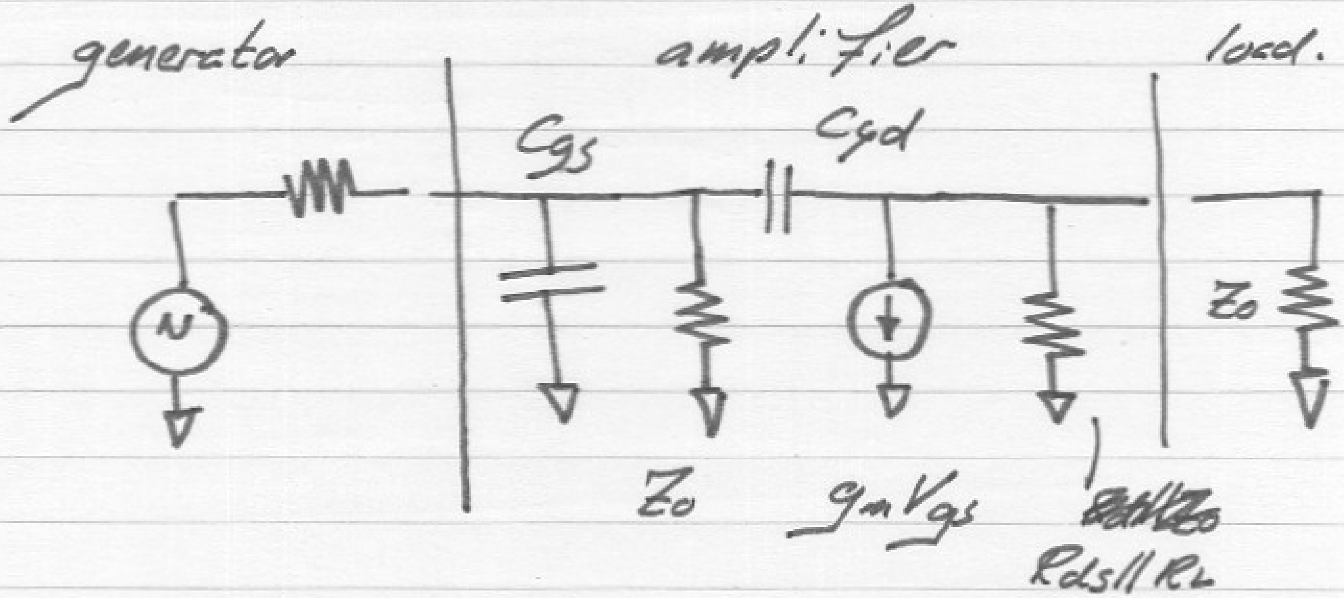
or



Lets use this FET Model



which gives an a.c. model thus:



we have chosen R_L so that $R_{ds} || R_L = Z_o$

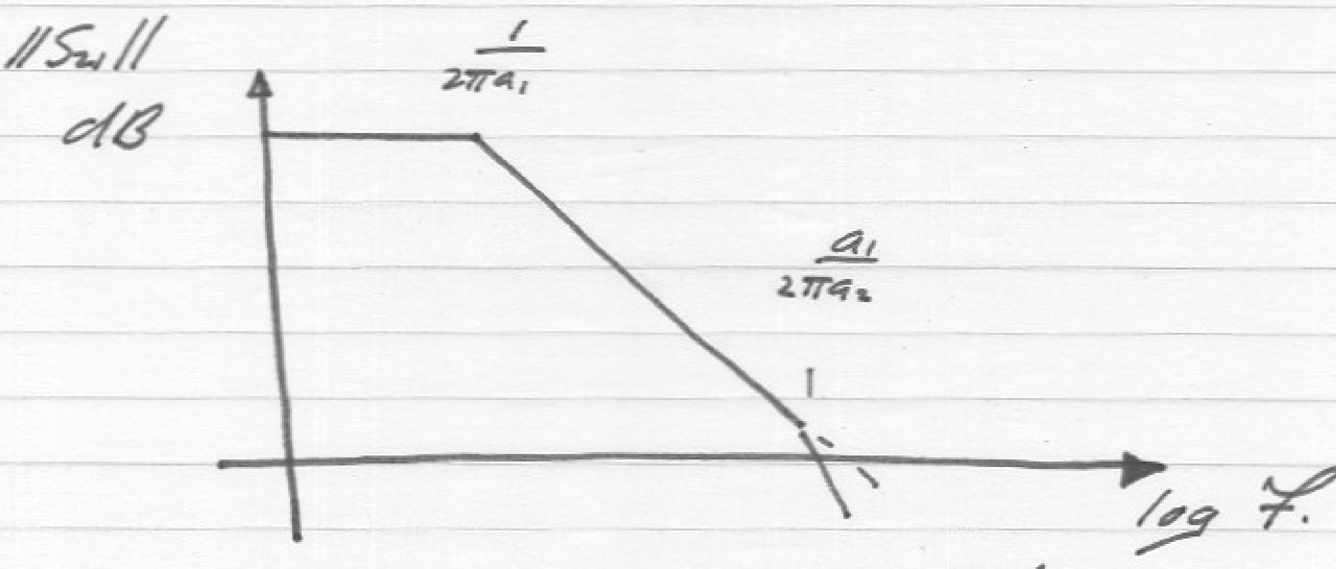
recall that $V_{out}/V_{gen} = S_{21}/2$ if $R_{gen} = R_L = Z_0$

$$\Rightarrow \quad S_{21} \cong -g_m (Z_0/2) \frac{1}{1 + j\omega a_1}$$

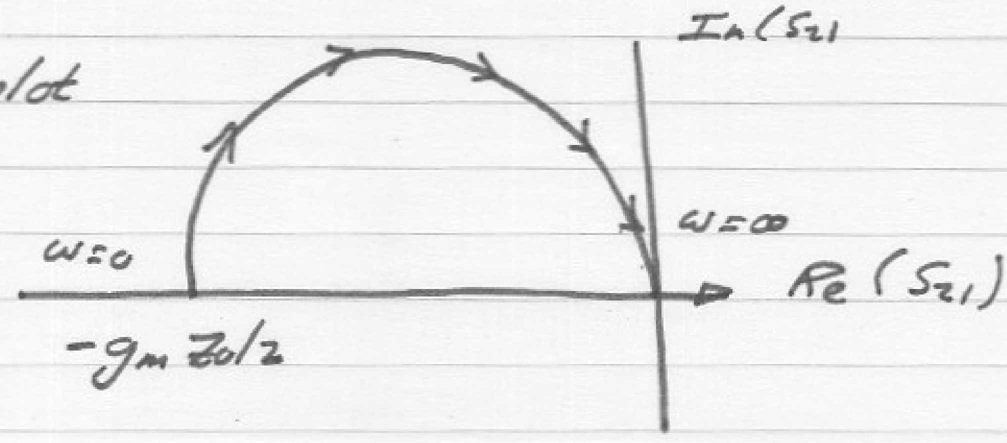
I have used the MOTC & dropped the 2nd pole

$$a_1 \doteq C_{gs}(Z_0/2) + C_{gd}\left[\frac{Z_0}{2}(1 + g_m Z_0/2) + Z_0/2 \right]$$

the Miller effect should be obvious ...

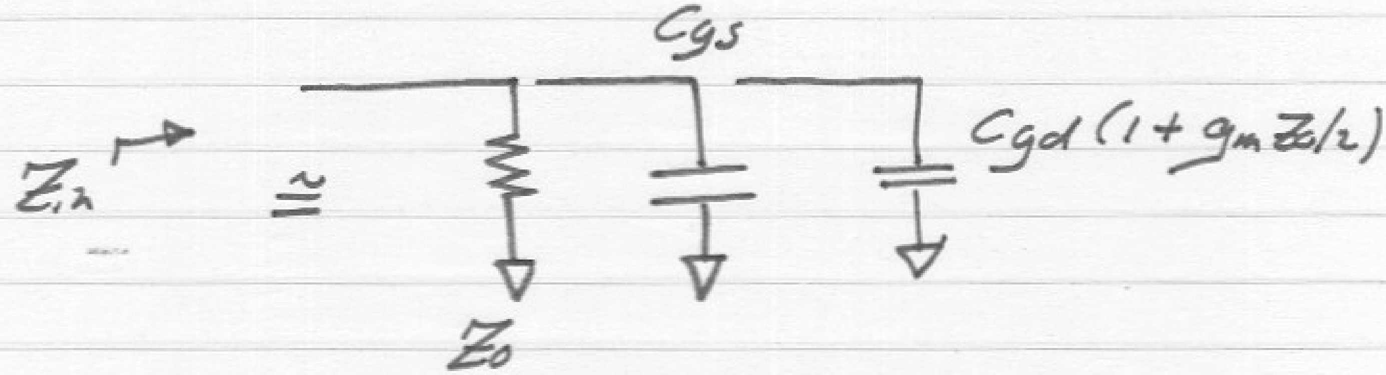


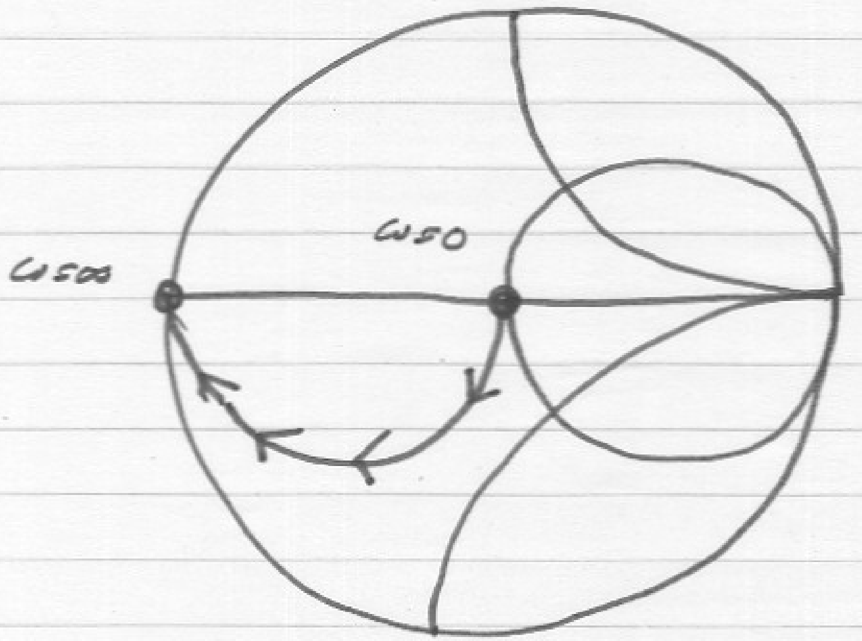
polar plot



Input impedance:

By Miller approximation:

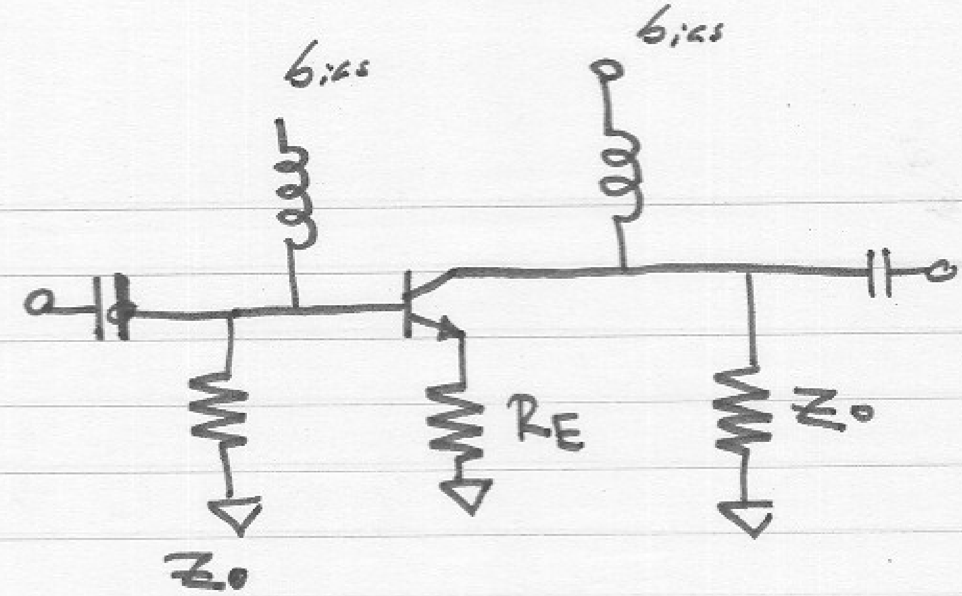


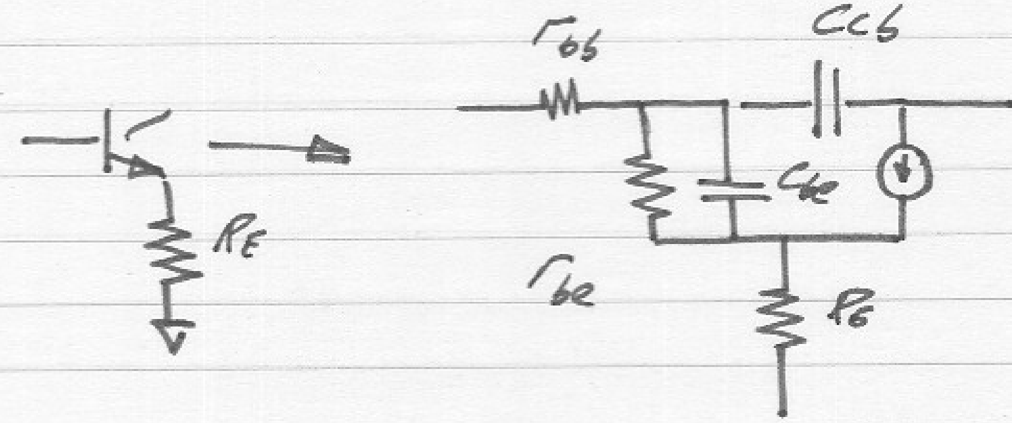


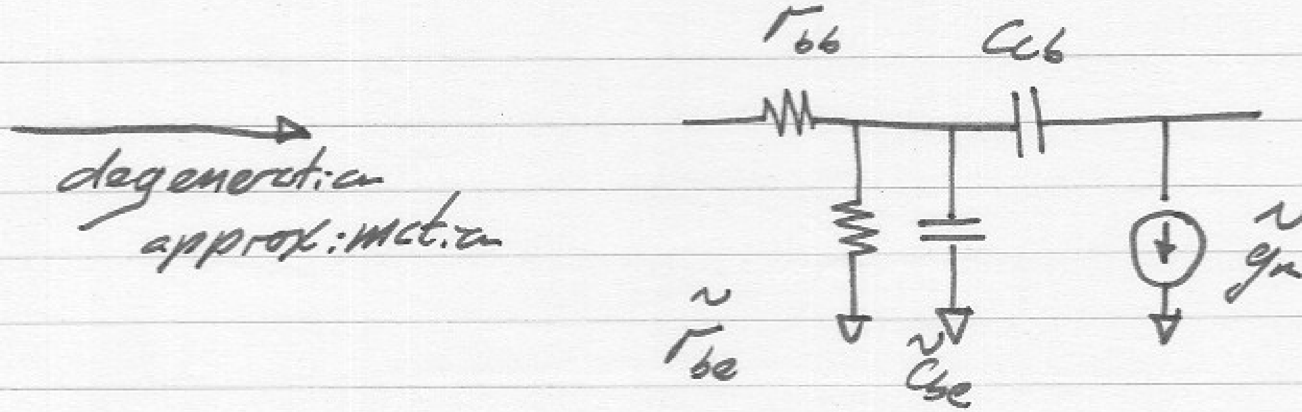
Comments

1) output impedance really strongly involves the secondary-pole effects → analysis harder.

2) It is fairly easy to carry this analysis ~~of~~ over to a bipolar circuit of the same type.



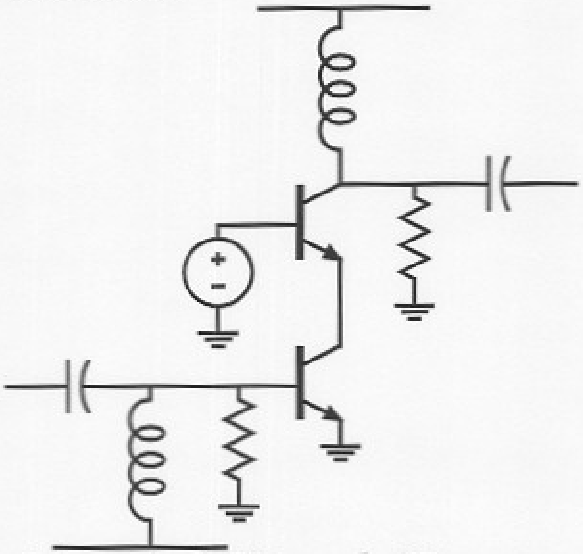




... and the circuit can now be analyzed...

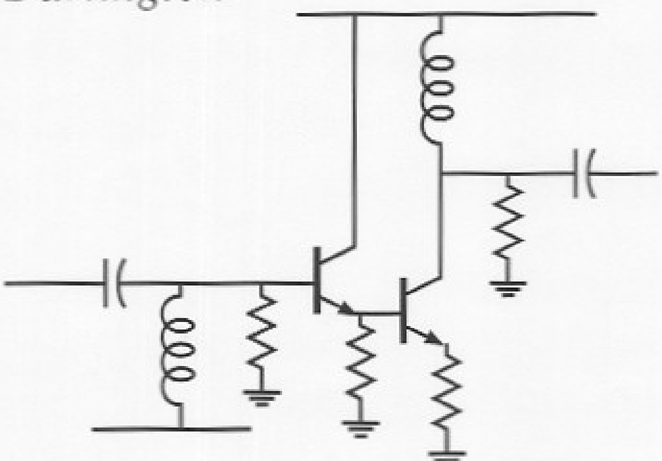
Other Broadband Stages:

Cascode:



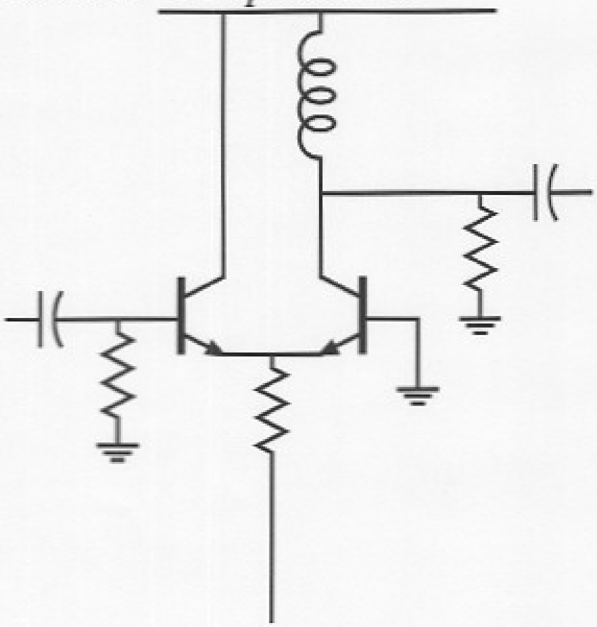
Cascaded CE and CB stages, Eliminates Miller Multiplication effect relative to CE stage
But watch for miller-effect-like time-constant multiplication with the CB stage

Darlington



If done properly, the Emitter follower stage drives the CE stage with a lower impedance than that of the generator. The time constants associated with the CE transistor capacitances are therefore reduced. But the Emitter follower introduces its own time constants...

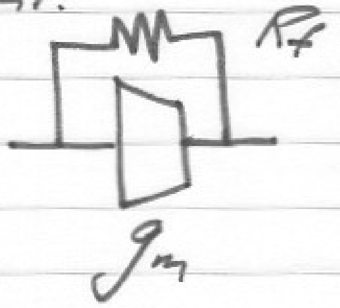
Emitter-Coupled Pair



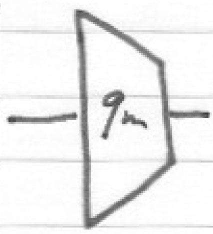
A cascaded CC and CB stage. Analyze by MOTC. No major miller effect, similar performance to cascode...

The broadband Feedback Amplifier

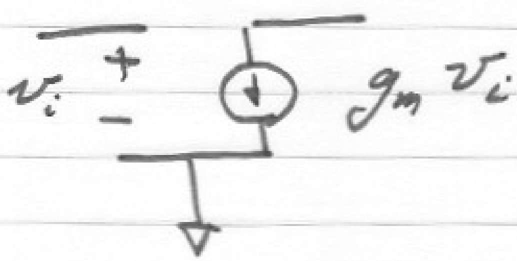
General:



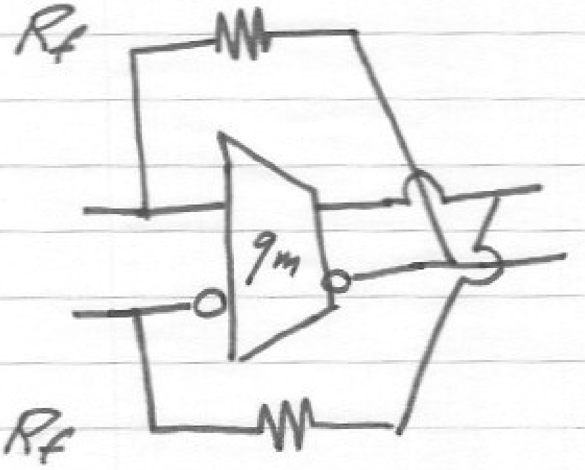
the block



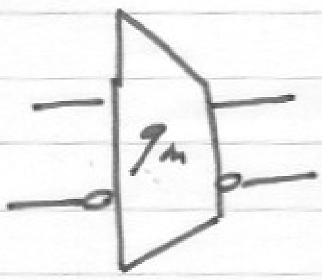
is a short hand for this:



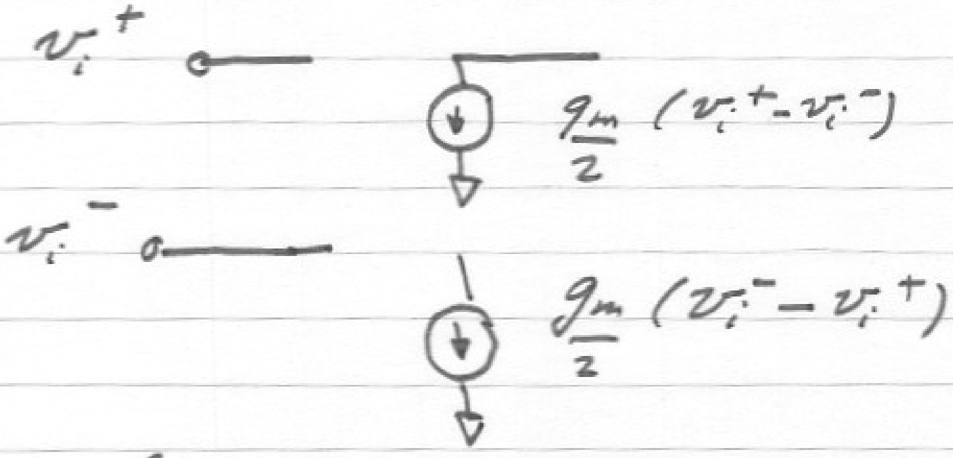
A differential form is like so:



where the block

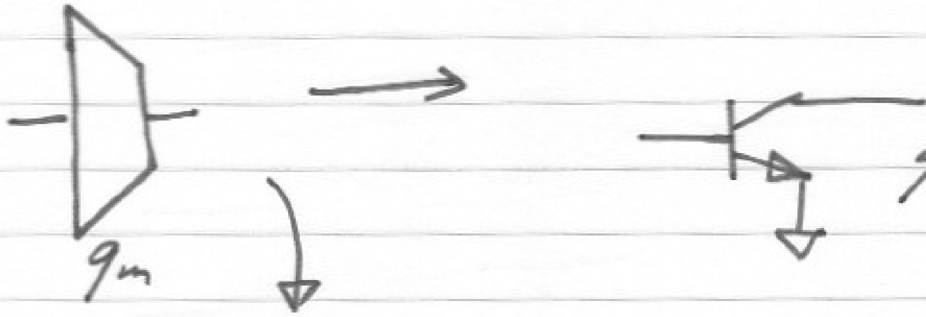


represents this:

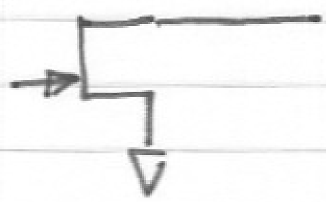


the factor of 1/2 is <u>not</u> standard; but we use <u>here</u>

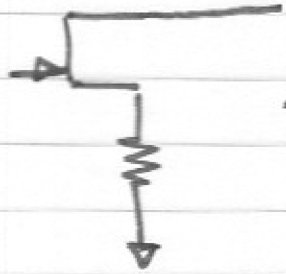
Physically, the g_m block is:



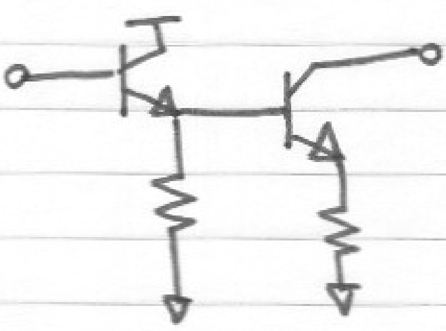
$$g_m = \frac{q I_E}{kT},$$



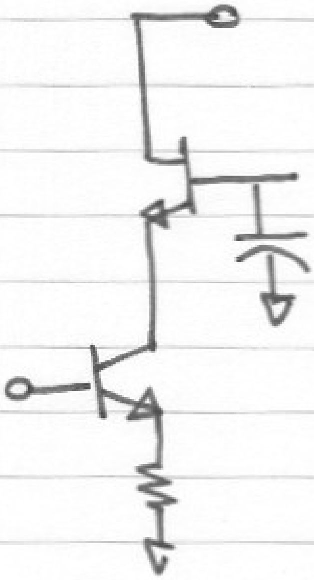
$$g_m = \frac{1}{R_E + \frac{kT}{q I_E}}$$



$$g_m = \frac{1}{R_S + 1/g_{m,FET}}$$

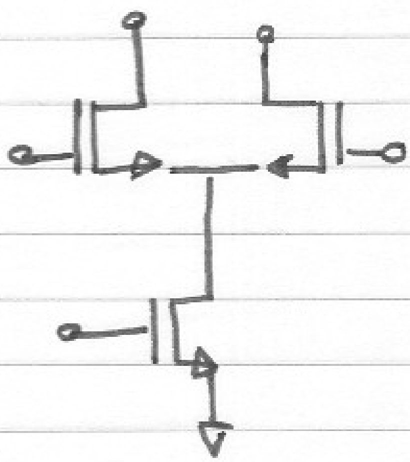


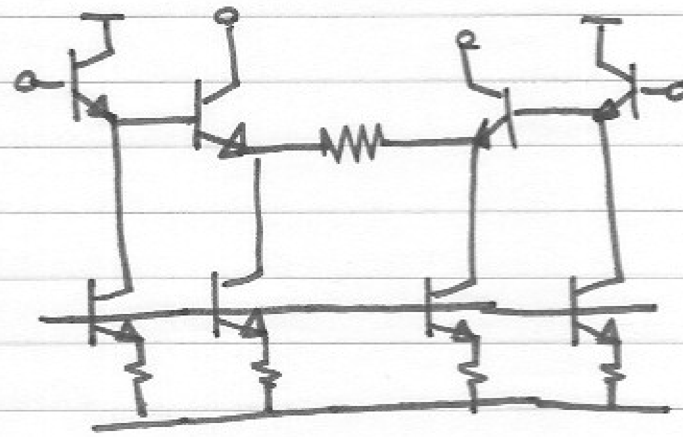
or

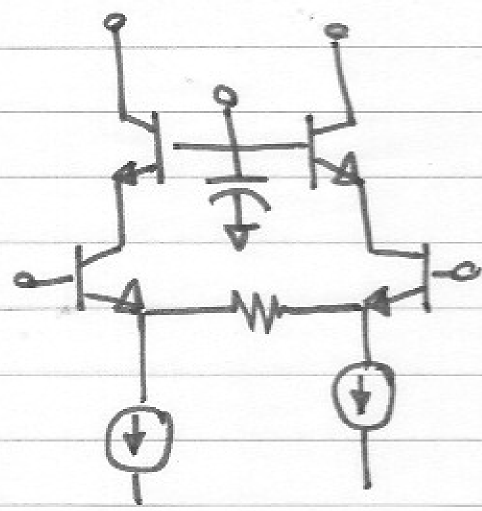


or ...

differential gm block examples:

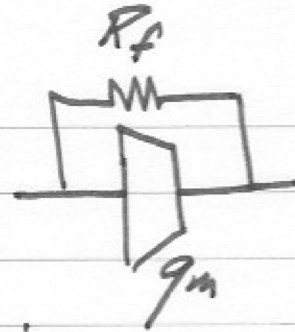


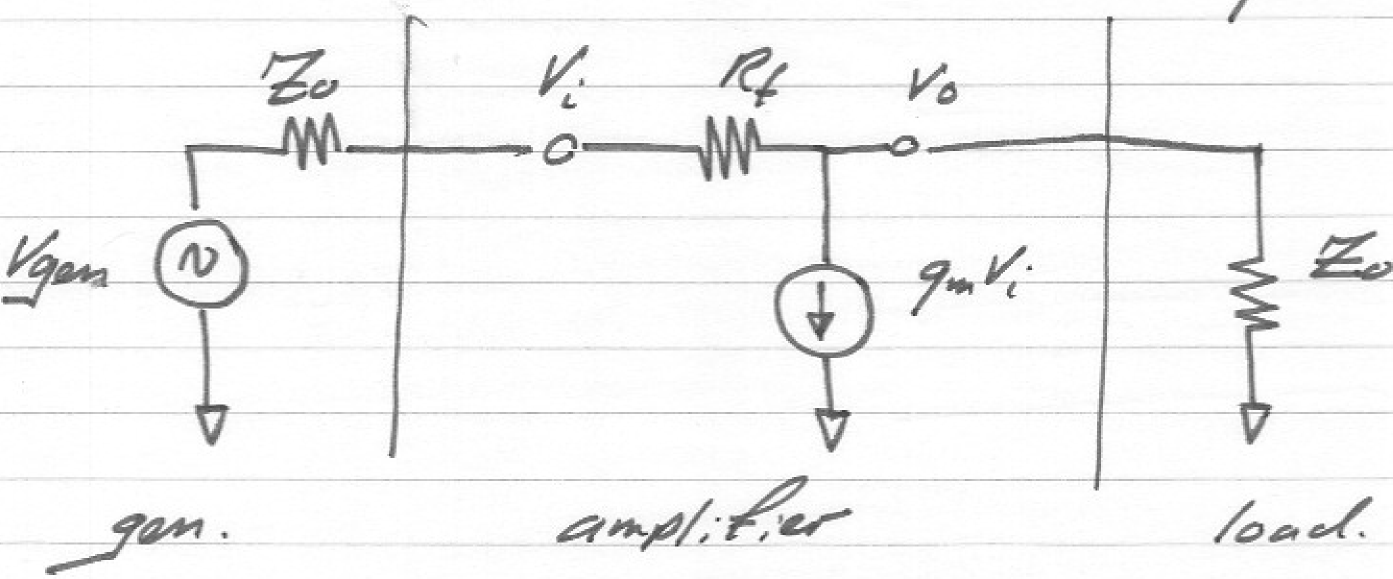




etc ...

Basic feedback amplifier:





Make the math easy:

Suppose we desire a gain $\underline{A}$
A is negative.

And suppose we desire

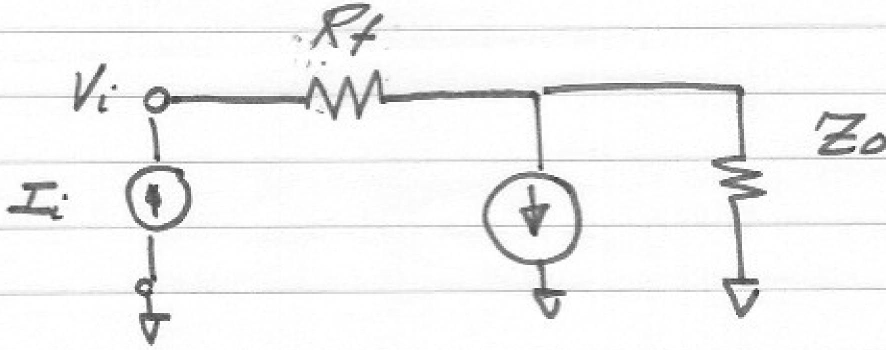
$$Z_{in} = Z_{out} = Z_o$$

this is achieved by setting

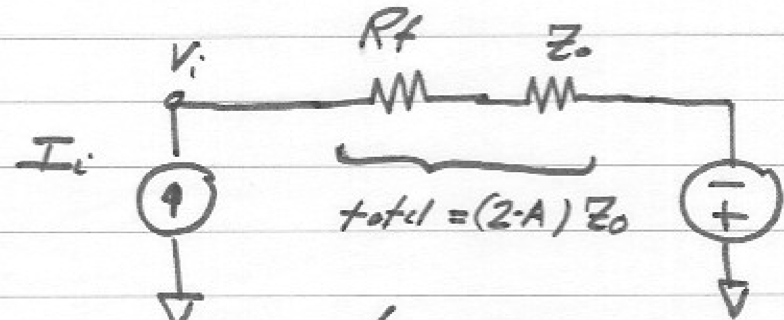
$$g_m = \frac{1-A}{Z_o} \quad \& \quad R_f = (1-A) Z_o$$

lets prove this...

Is $R_{in} = Z_0$???

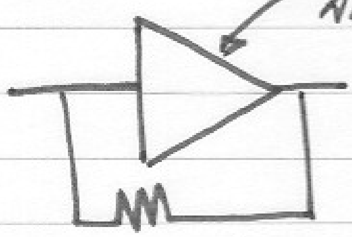


$g_m V_i$



$g_m V_i Z_o = (1-A) V_i$

$A_v = -(1-A)$

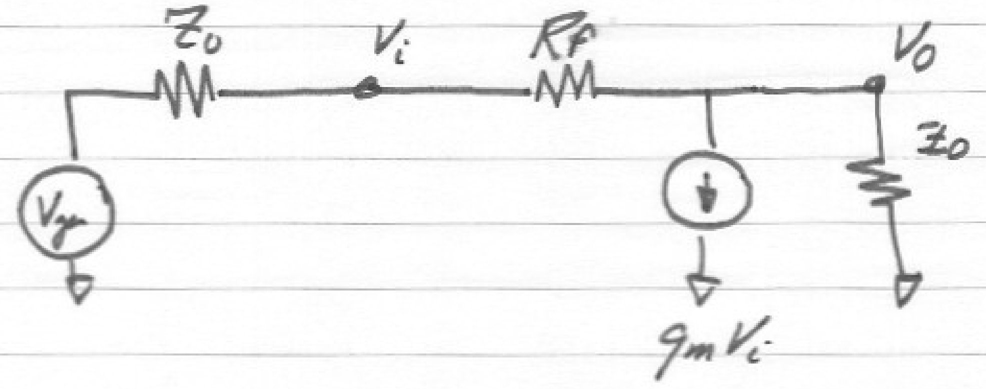


$R = (2-A) Z_o$

by Miller's theorem, $Z_{in} = \dfrac{(2-A) Z_o}{1+(1-A)} = Z_o$

$\dfrac{R}{1-A_v}$

what is the gain?



we have found $R_{in} = Z_0$, so $V_{in} = V_{gen}/2$.

$\Sigma I = 0$ at V_{out}:

$$\frac{V_0}{R_f} + \frac{V_0}{Z_0} + g_m V_i - \frac{V_i}{R_f} = 0$$

$$\frac{V_0}{Z_0(1-A)} + \frac{V_0}{Z_0} + \frac{(1-A)\,V_i}{Z_0} - \frac{V_i}{Z_0(1-A)} = 0$$

$$V_0\left[\frac{1}{1-A} + 1\right] + V_i(-1)\left[\frac{1}{1-A} - (1-A)\right] = 0$$

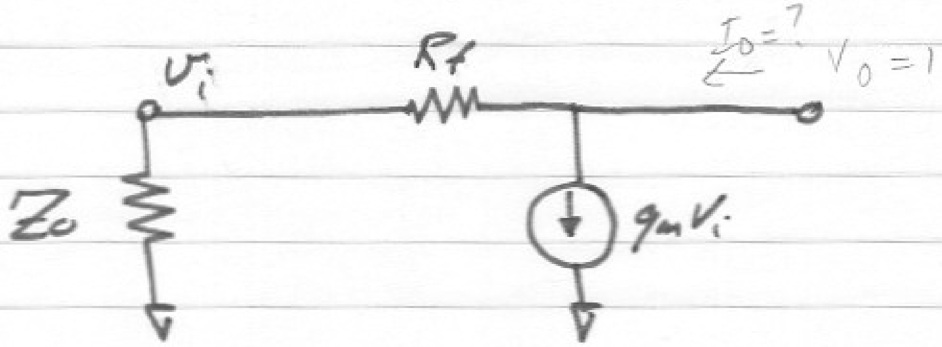
$$V_0\left[2-A\right] + V_i(-1)\left[1-(1-2A+A^2)\right] = 0$$

$$V_0\left[2-A\right] + V_i(-1)\left[-A^2+2A\right] = 0$$

$$\rightarrow V_0/V_i = A$$

☺

what is the output impedance ?



$$G_{out} = \frac{1}{R_{out}} = \frac{1}{Z_0 + R_f} + \frac{Z_0}{Z_0 + R_f} \cdot g_m$$

$$= \frac{1}{Z_0 + Z_0(1-A)} + \frac{Z_0}{Z_0 + Z_0(1-A)} \cdot \frac{1-A}{Z_0}$$

So

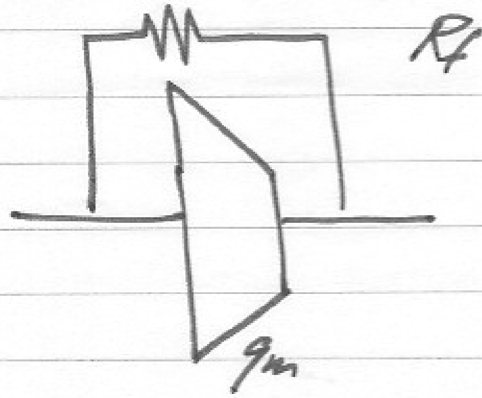
$$\frac{Z_0}{R_{out}} = \frac{1}{2-A} + \frac{1}{2-A}(1-A)$$

$$= \frac{2-A}{2-A}$$

$$= 1$$

= ☺

ok, to wrap it up:



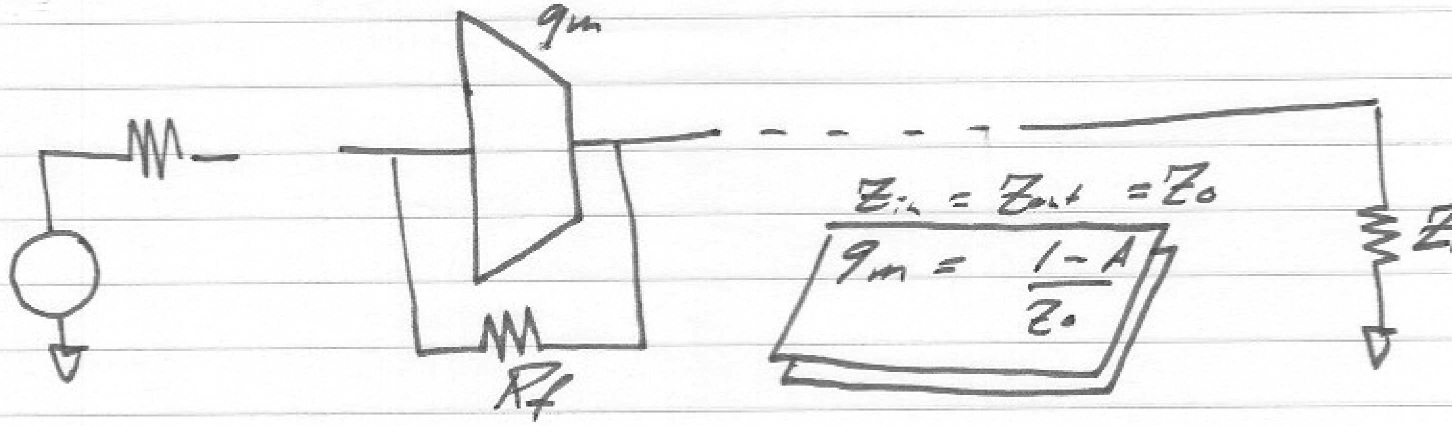
has $Z_{in} = Z_{out} = Z_0$

and gain A (negative)

If $R_f = Z_0(1-A)$

$$g_m = \frac{1-A}{Z_0}.$$

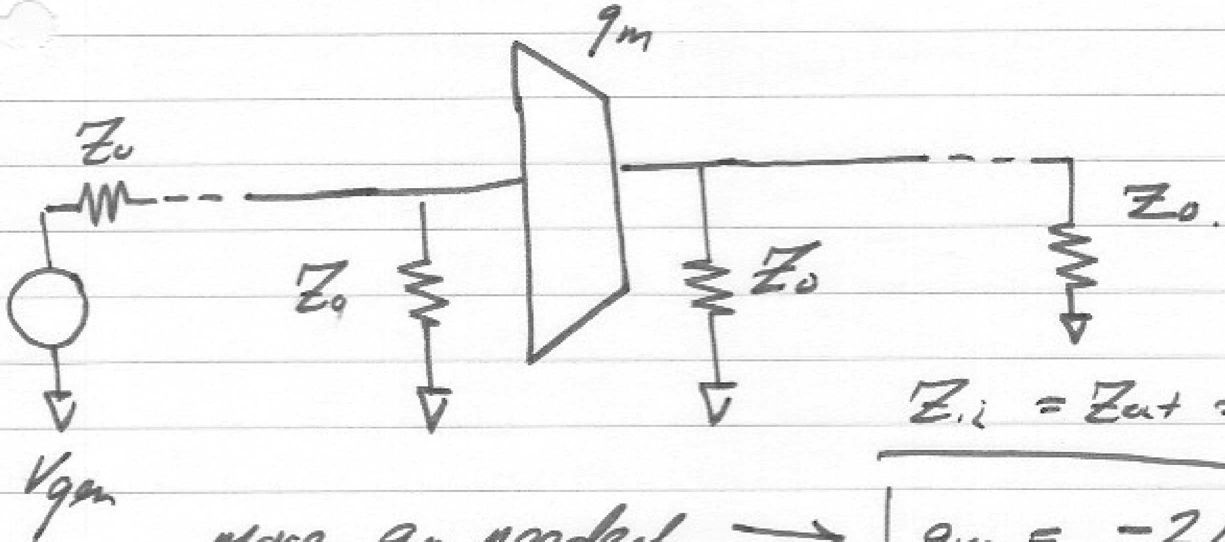
So, why do this?



$Z_{in} = Z_{out} = Z_o$

$\boxed{g_m = \frac{1-A}{Z_o}}$

... when we could do this ...

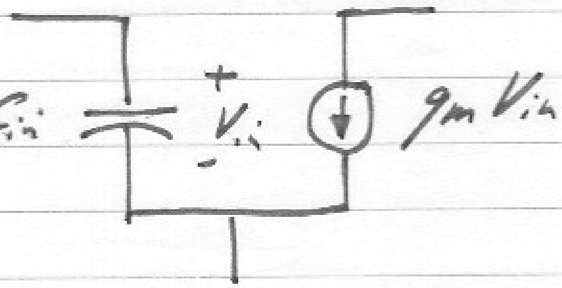


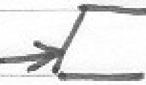
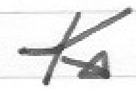
$Z_{in} = Z_{out} = Z_o$

more gm needed for a given gain. → $\boxed{g_m = -2A/Z_o.}$

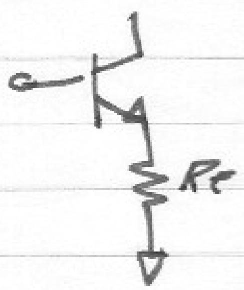
the answer is superior bandwidth

If we assume a device model like so



for  or this 

then $f_T = \frac{g_m}{2\pi C_{in}} \longrightarrow \boxed{C_{in} = \frac{g_m}{2\pi f_T}}$

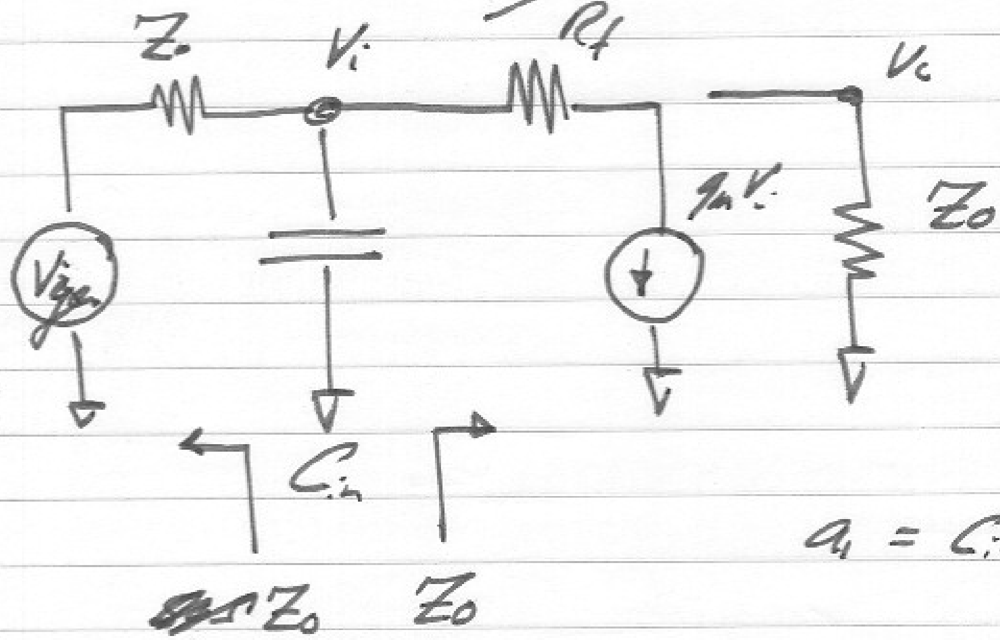
even for 

$$\frac{\tilde{g}_m}{g_m} = \frac{\tilde{C}_{in}}{C_{in}}$$

so $\tilde{C}_{in} = \frac{\tilde{g}_m}{2\pi f_T}$

the input capacitance is always proportional to g_m, and the ~~resist~~ resulting bandwidth is thus larger for the feedback amplifier.

Bandwidth analyses:



$$a_1 = C_{in} \cdot Z_0/2$$

but $C_{in} = \dfrac{g_m}{2\pi f_T} = \dfrac{1}{2\pi f_T} \dfrac{1-A}{Z_0}$

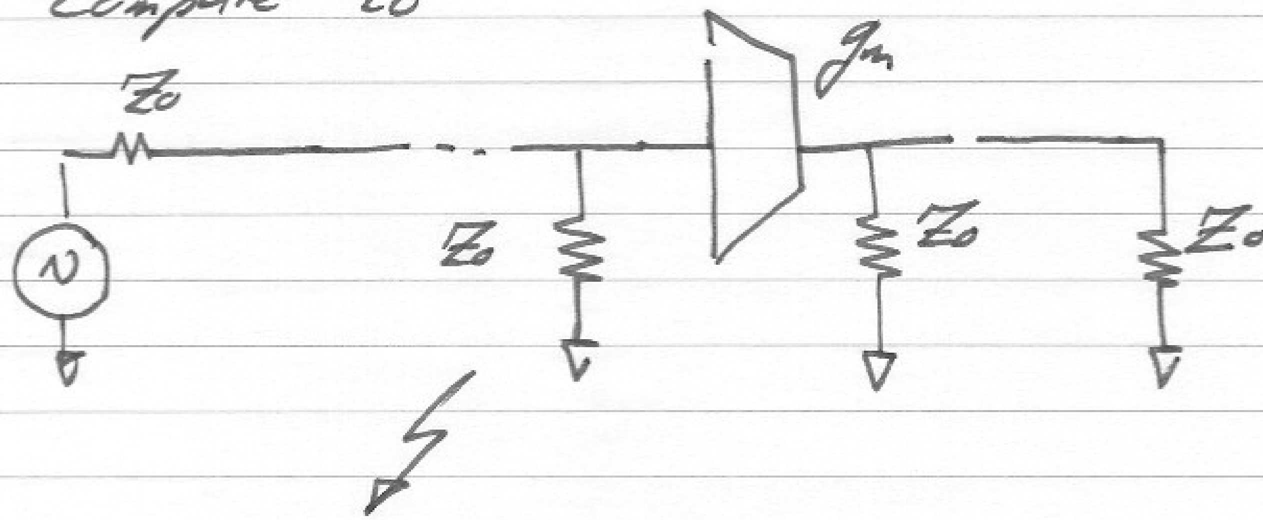
so $a_1 = \dfrac{1}{2\pi f_T} \dfrac{1-A}{2}$

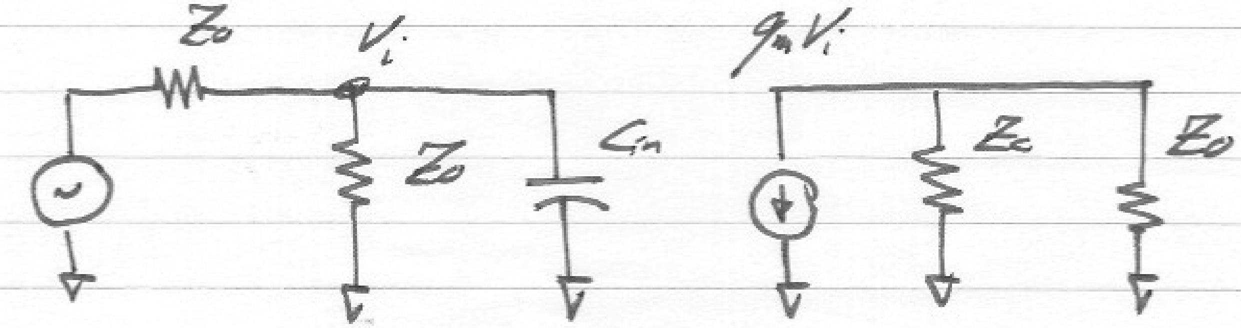
and $f_{3dB} \cong \dfrac{1}{2\pi a_1} = f_T \dfrac{2}{1-A}$

$$\boxed{f_{3dB} = f_T \frac{2}{1-A}}$$

(given this highly simplified device model...)

Compare to





$a_1 = (Z_0/2)\,C_{in}$ but $C_{in} = \dfrac{g_m}{2\pi f_T} =$

$$C_{in} = \cancel{g_m}\,\frac{1}{2\pi f_T}\cdot\frac{-A}{(Z_0/2)}$$

so $a_1 = (Z_0/2)\,\dfrac{1}{2\pi f_T}\;\dfrac{-A}{Z_0/2}$

$$a_1 = \frac{-A}{2\pi f_T}$$

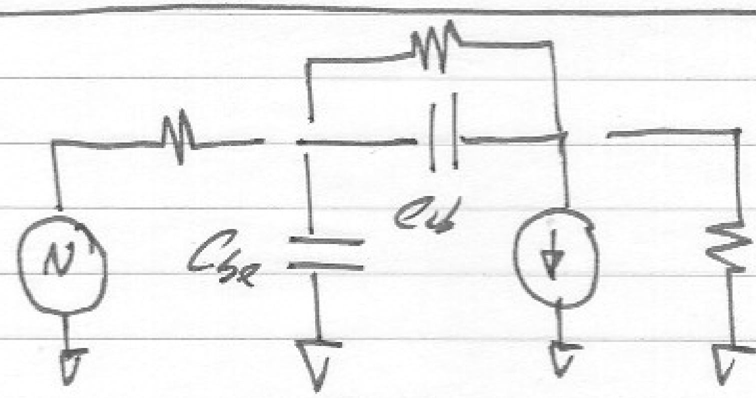
$$\boxed{f_{3dB} = \frac{1}{2\pi a_1} = \frac{f_T}{-A}}$$

The feedback amplifier has superior bandwidth!

Feedback $$f_{3dB} = f_T \cdot \left(\frac{2}{1-A}\right)$$

Resistive termination $$f_{3dB} = f_T \left(\frac{1}{-A}\right)$$

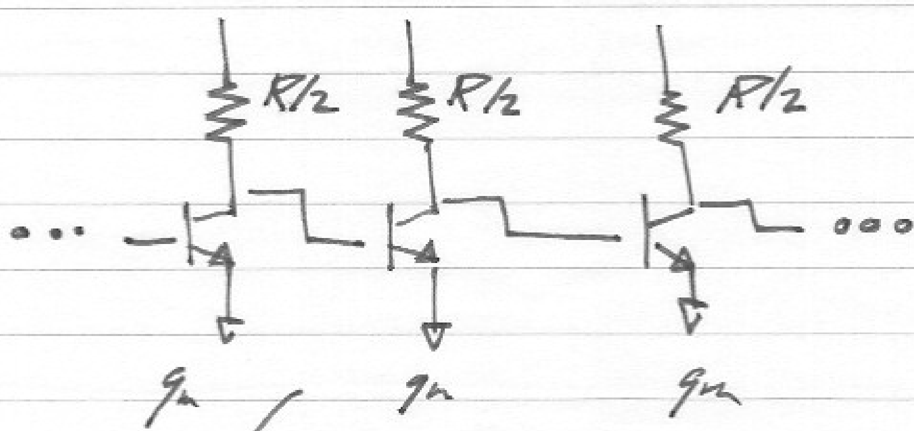
For moderately high gain, this is almost 2:1 superior!



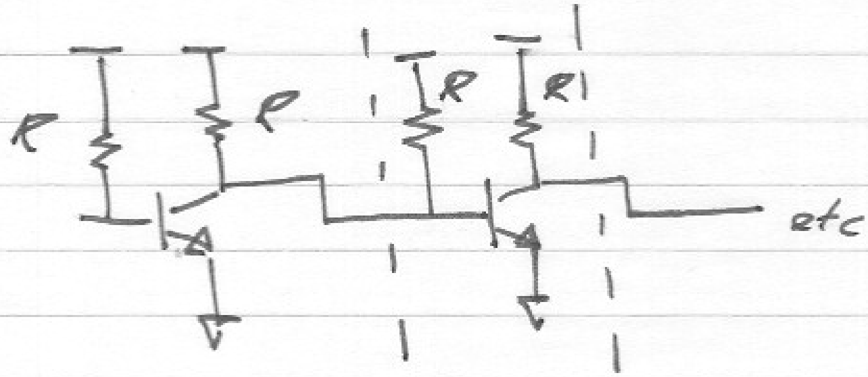
In the presence of C_{cb}, it is easy to show that

$$a_1 \cong Z_0/2 \cdot C_{be} + C_{cb}(1-A)(Z_0/2)$$

the superior bandwidth holds for "non"-microwave" amplifiers, e.g. those not having to meet a $Z_{in} = Z_{out} = Z_0$ criterion.

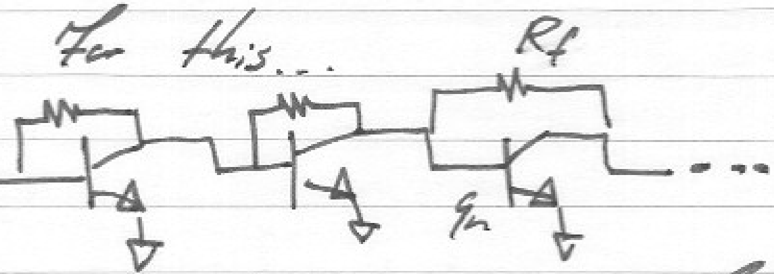


equals



each stage obeys $f_{3dB} = \dfrac{f_T}{-A}$

while for this...



each stage obey $f_{3dB} = f_T \cdot \dfrac{2}{1-A}$