Mixed Signal IC Design Notes set 3: More High Frequency Amplifier Design

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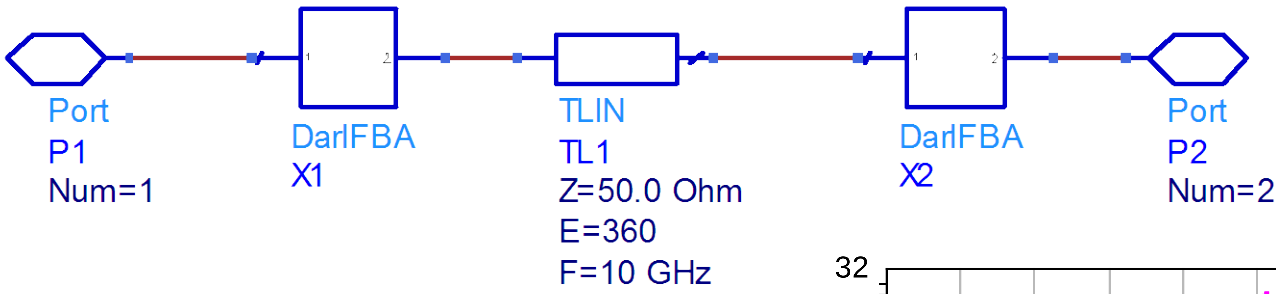
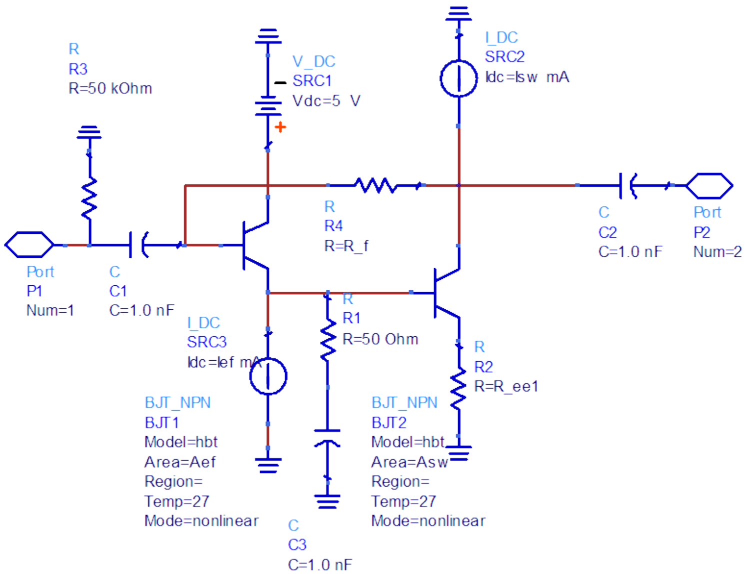
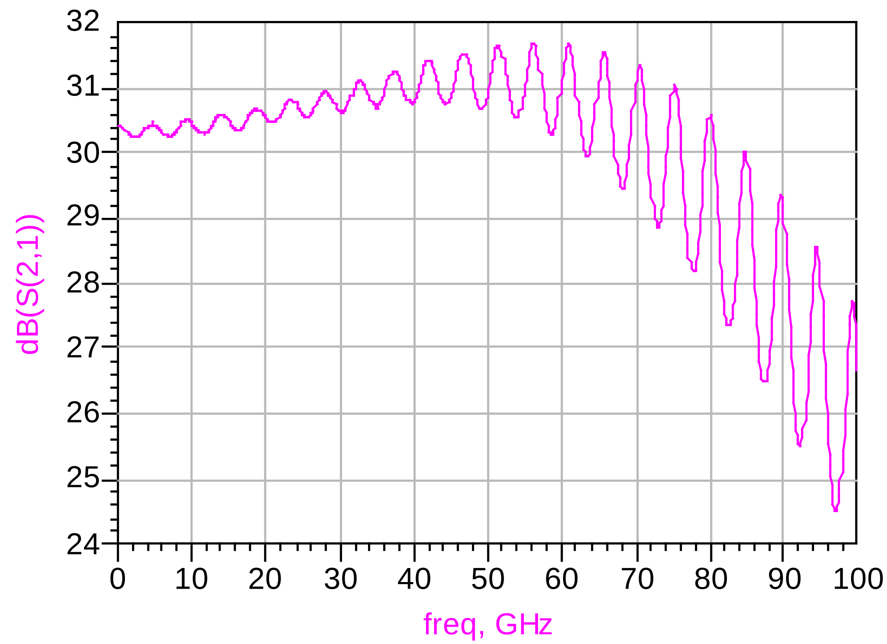
Why do we care about impedances matched to 50 Ohms?

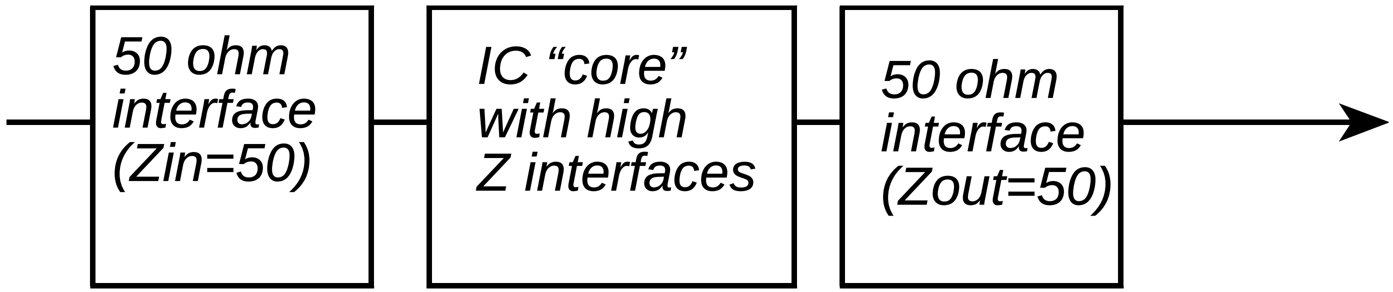






Standing waves on transmission lines cause gain/phase ripples of the form $\left(1-\Gamma_S \Gamma_L \exp(-j2f\tau)\right)^{-1}$. Either we must have short transmission lines, or the lines must be well-terminated

Why do we care about impedances matched to 50 Ohms?



If IC has small dimensions, we can use high-z internal interfaces, and just interface to e.g. 50 Ohms at chip boundaries or (in some cases) between major circuit blocks on wafer

Why do we care about transmission lines ?

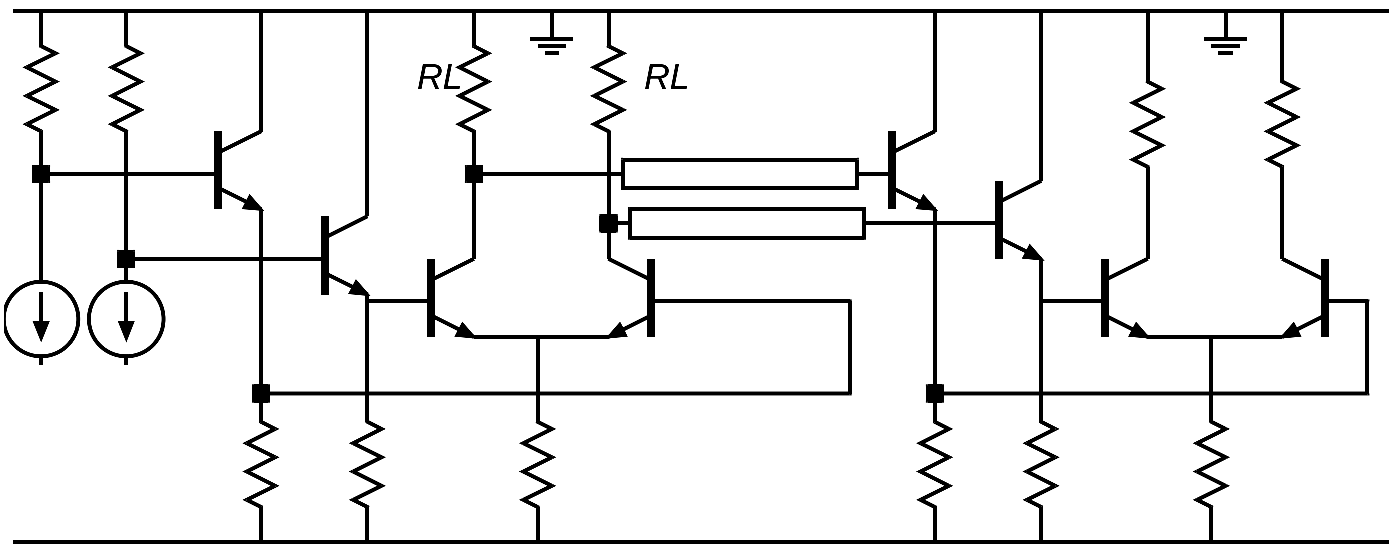
There are times when we would like to ignore transmission lines and S - parameters.

If the IC has interconnects which are short, then we can treat all interconnects as **lumped**.

This means we can ignore standing waves on the lines.

But, we must be aware that $C_{line} = \tau_{line} / Z_o$, $L_{line} = \tau_{line} * Z_o$,
hence a short line can be modeled as lumped
but its parasitics **cannot be ignored**.

Effect of lumped wiring parasitics on IC performance



Here we can model these short lines as lumped. Ignoring BJT parasitics,

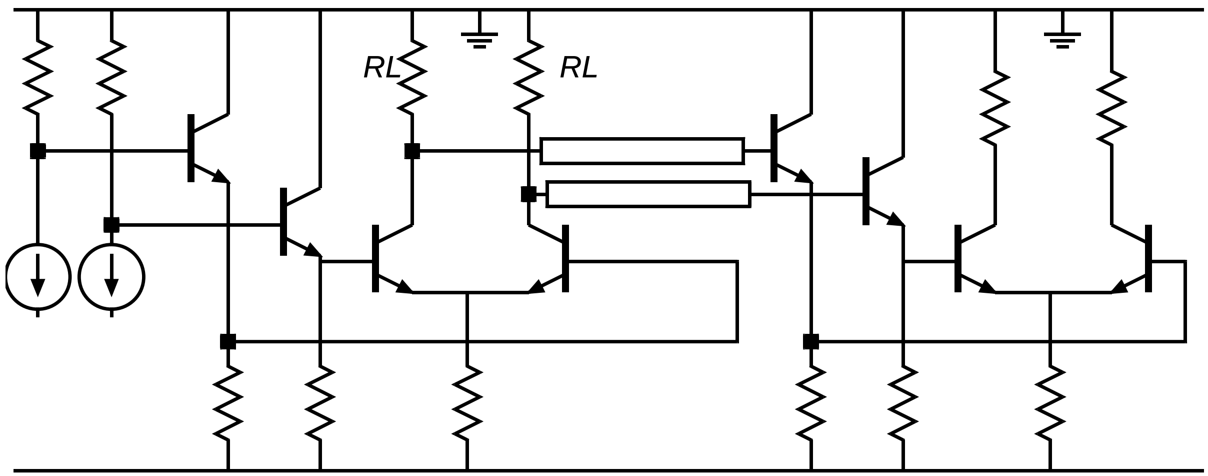
$C_{line} = \tau_{line} / Z_o, L_{line} = \tau_{line} * Z_o,$

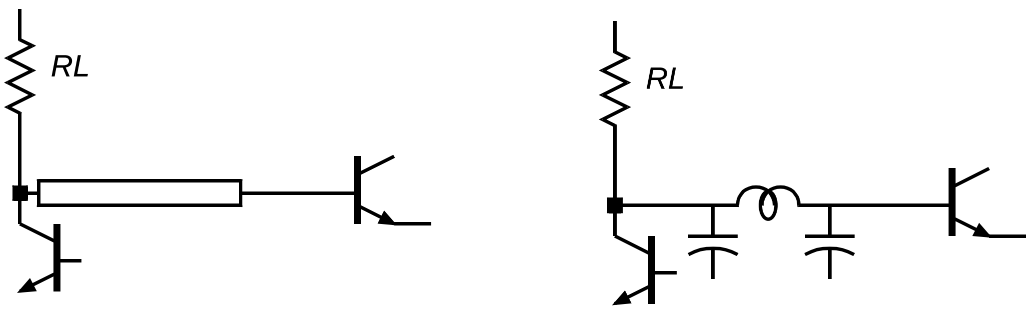
And we get an RC charging time of

$\tau_{\text{wiring}} = C_{line} R_L$

Since the gain is $g_m R_L = (kT / qI) R_L$, designing for low power (low I) means designing for high R_L and hence interconnect capacitance becomes more important

Effect of lumped wiring parasitics on IC performance





Once we have studied EF frequency response in detail, we will understand that

$$L_{line} = \tau_{line} * Z_o,$$

...can have substantial effect upon the EF stability and pulse response

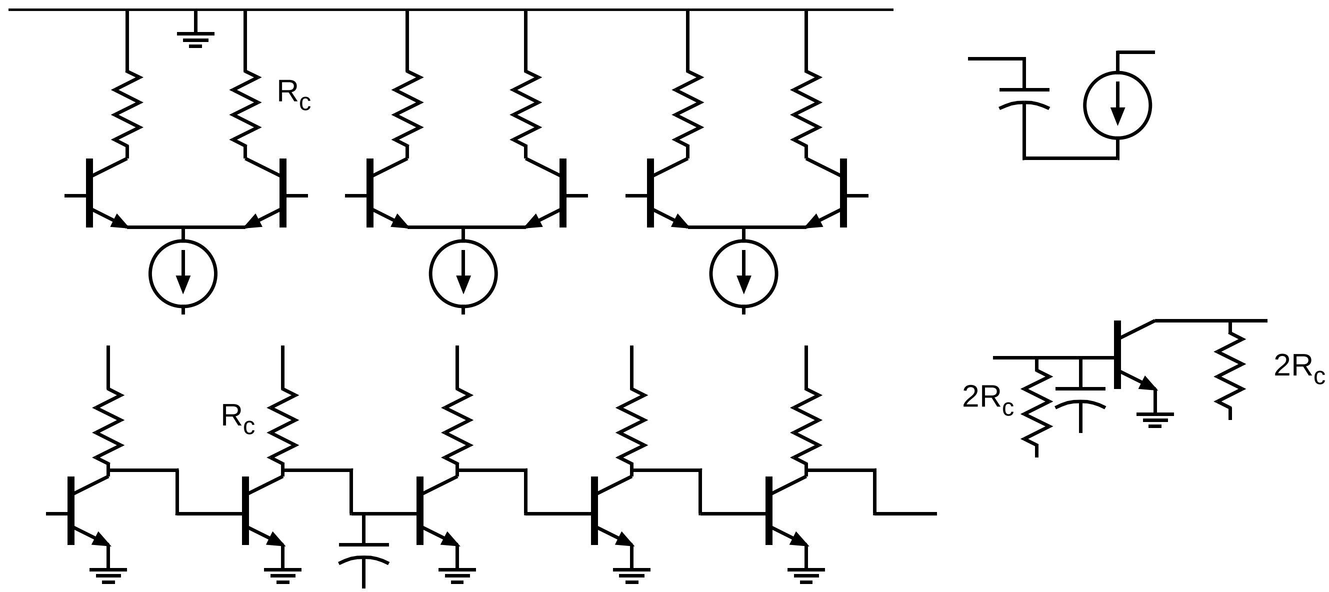
Getting more bandwidth

At this point we have learned basics (MOTC etc)

How can we get more bandwidth…. ?

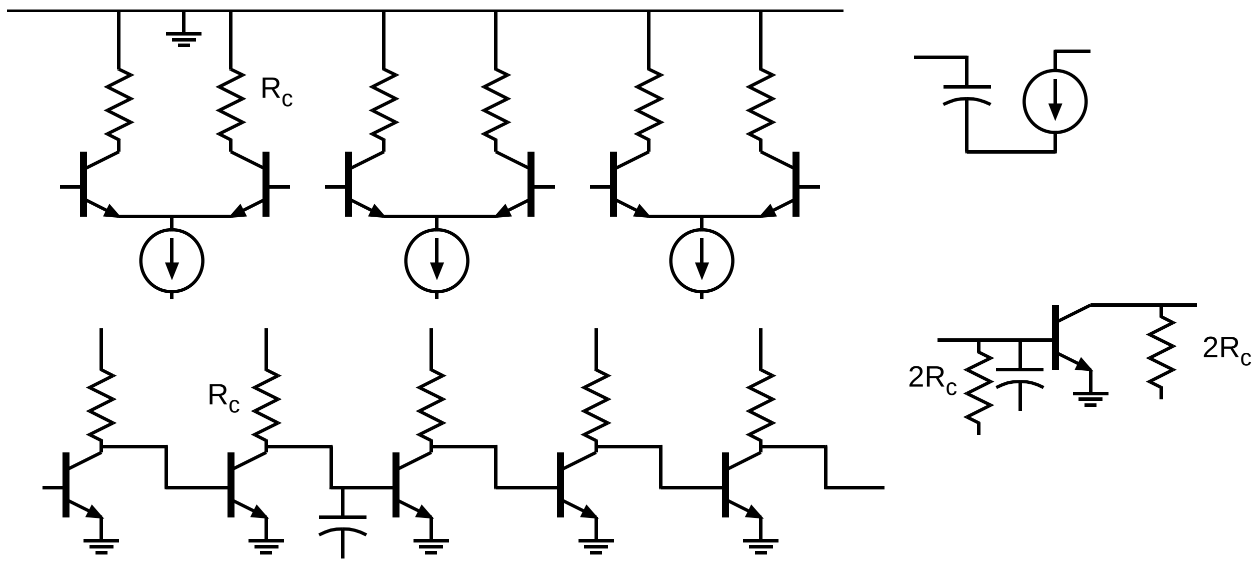
- Resistive feedback
- transconductance-transimpedance
- emitter-follower buffers, benefits and headaches
- emitter degeneration…basics
- emitter degeneration and area scaling
- ft-doubler stages..
- distributed amplifiers
- ft-doubler stages…simple, RL, power, with Darlington
- broadbanding / peaking with LC networks….
- distributed amplifiers
- examples….

Broadband gain blocks: "worlds simplest amplifier"



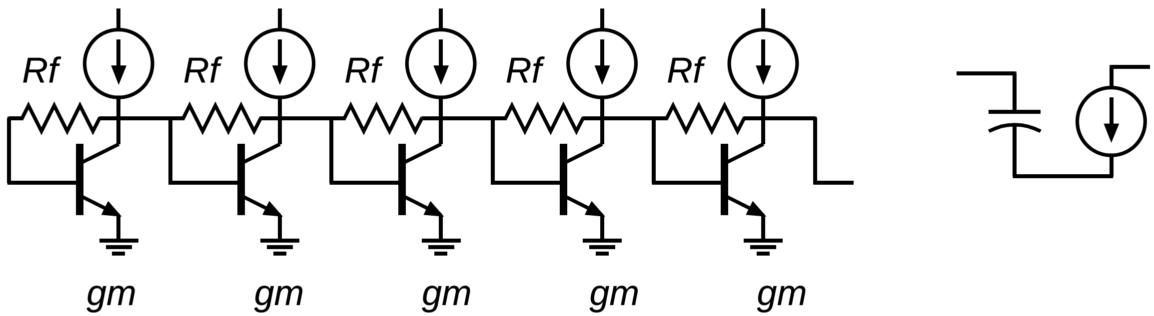
Examine a cascade of simple differential pairs. For the differential signal path, this becomes a CE stage cascade. If we model each BJT as a simple g_m & C_{in} combination, the gain per stage is $-g_m R_C$, the bandwidth is $1/2\pi R_C C_{in}$, and the gain - bandwidth product is $g_m / 2\pi C_{in} = f_\tau$. This is the origin of the " f_τ limit". We can do much better.

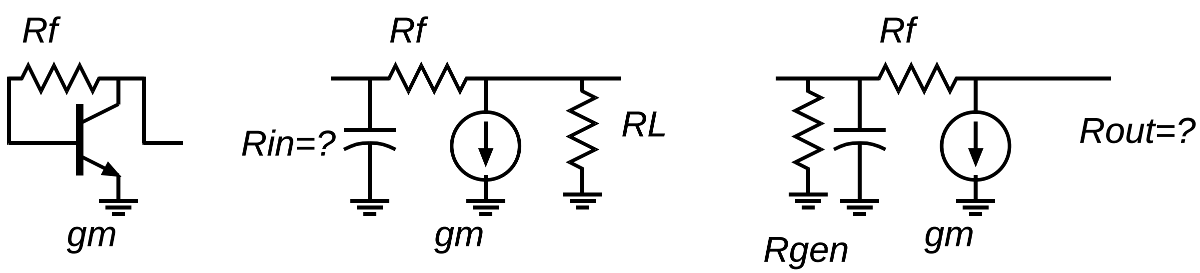
Broadband gain blocks: "worlds simplest amplifier"



Let us think this through. Transistors are made to provide transconductance g_m. An input voltage is converted into an output current. Our simplest design strategy is to convert this output current back into a voltage using a resistor...hence the gain relationship. Transconductance *always* comes at the price of input capacitance, by the ratio $C_{in} = g_m / 2\pi f_\tau$

Resistive feedback stages, I: gain and impedance



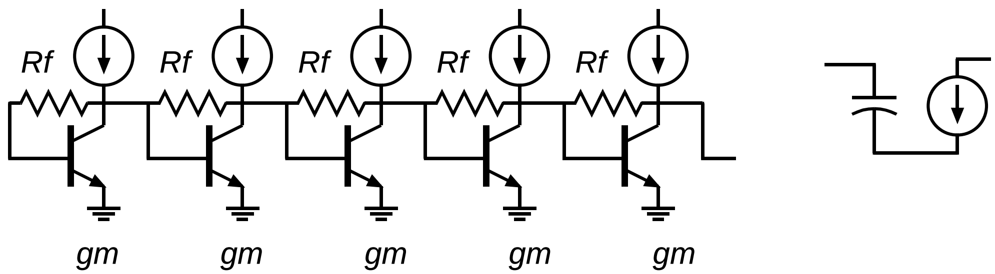


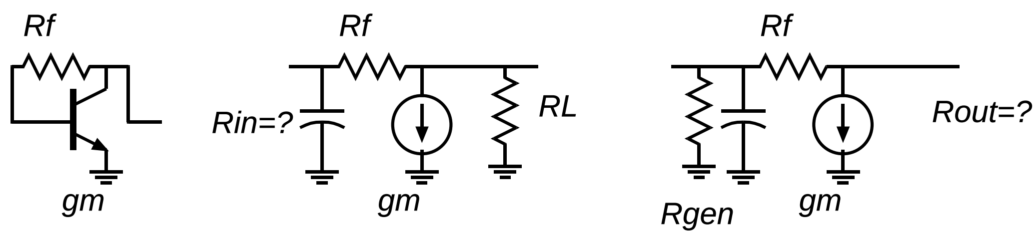
This is a more sophisticated way of converting g_m into voltage gain....nodal analysis to be done in lecture....

Per-stage gain of A (negative) with input and output resistances of $R_{in/out}$ are obtained if we set

$g_m = (1-A)/R_{in/out}$ and $R_f = (1-A)*R_{in/out}$.

Resistive feedback stages, I: bandwidth and gain-bandwidth





Since $g_m = (1-A)/R_{in/out}$ and $C = g_m / 2\pi f_\tau$,

$$f_{3dB} = 1/\big(2\pi (R_{in/out}/2)C\big) = g_m / \big(\pi C (1-A)\big)$$

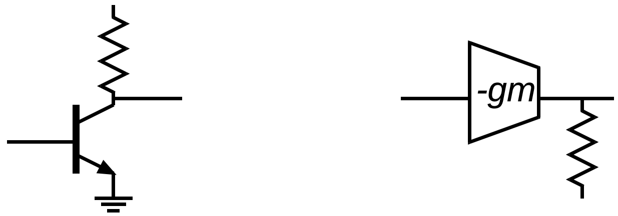
$$f_{3dB} = 2 f_\tau / (1-A)....$$

Note that for $A >> 1$, this is *twice* the bandwidth of a simple CE stage!

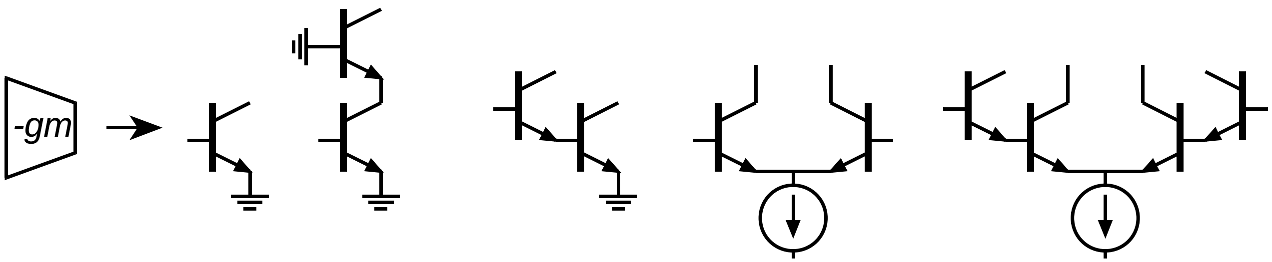
Why have we nearly doubled the bandwidth ?

The resistive feedback has provided *output* impedance of $R_{in/out}$ without losing significant output current in a physical loading resistance. We need roughly 1/2 the g_m to obtain the desired gain at a given impedance....so we get 1/2 the C, and hence twice the bandwidth...

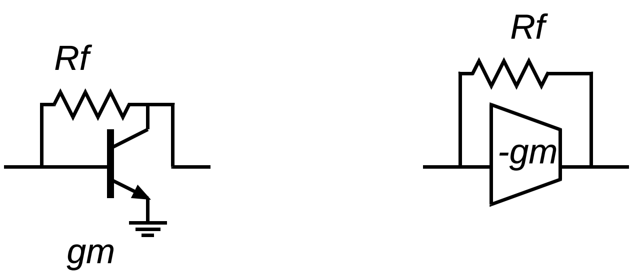
Variations on resistive feedback.



resistive loading...



gm blocks

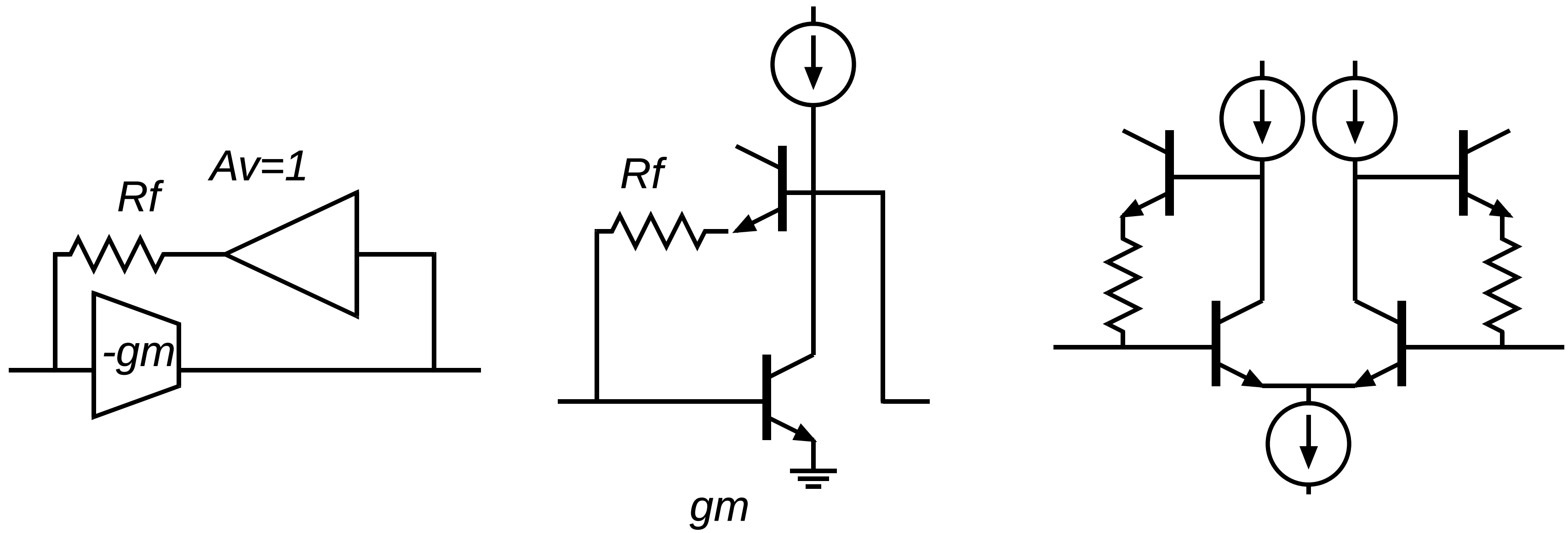


resistive feedback

We can use sets of transistors to build the g_m block.

This block can then be used with either resistive loading or resistive feedback to build an amplifier

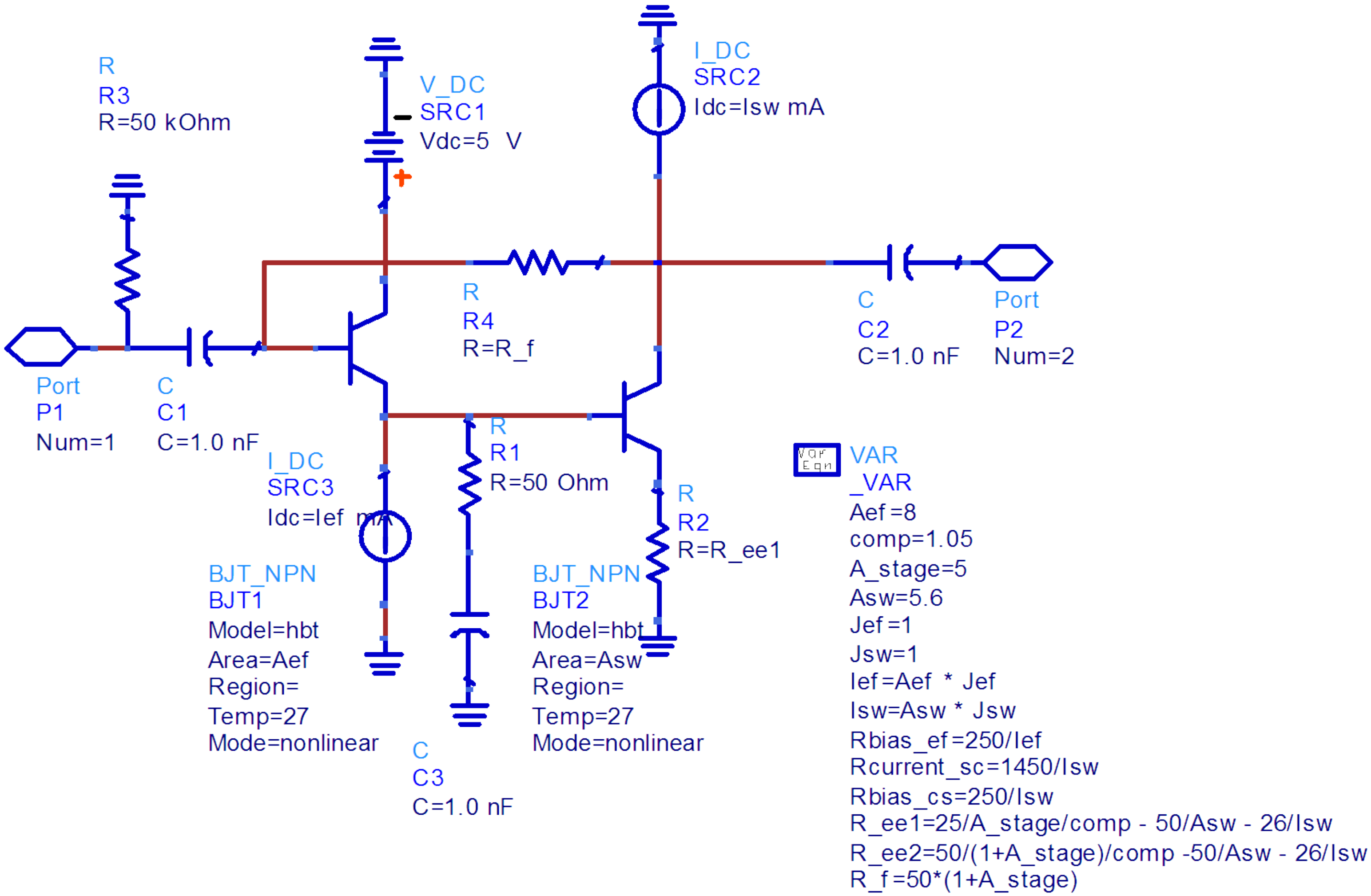
Feedback with buffer



This is often used for ease in biasing. The required g_m is also reduced to $g_m = -A / R_{in/out}$.

This improves bandwidth for low-gain stages, though the improvement is partially offset by the additional capacitances of the EF stage.

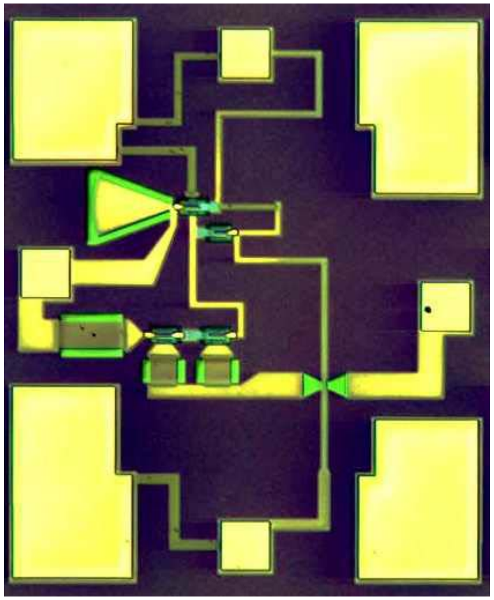
Resistive feedback as 50 Ohm gain blocks



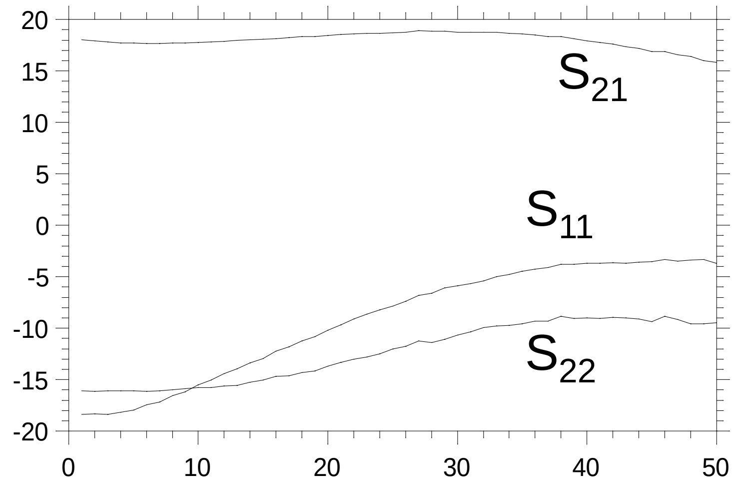
Note the gain relationships...

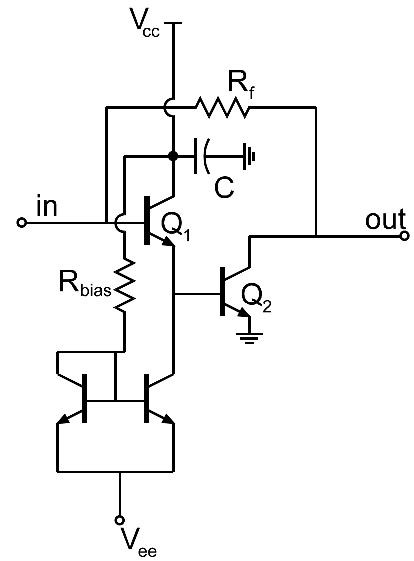
High Speed Amplifiers

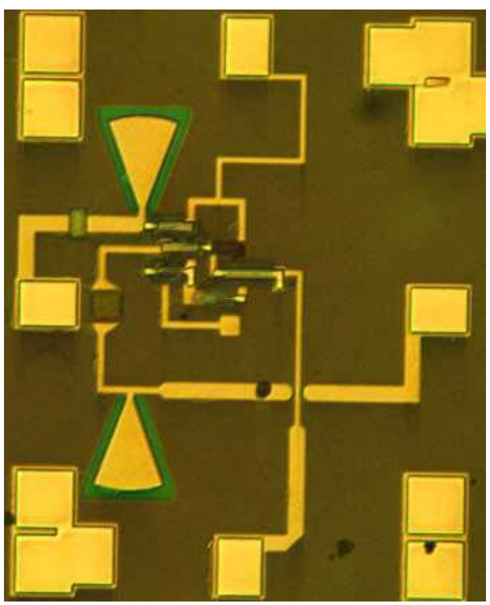
Dino Mensa
PK Sundararajan



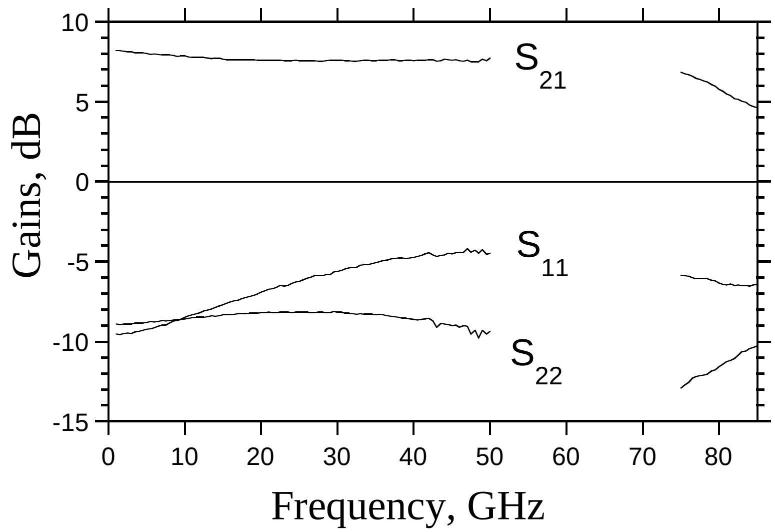
18 dB, DC-50 GHz

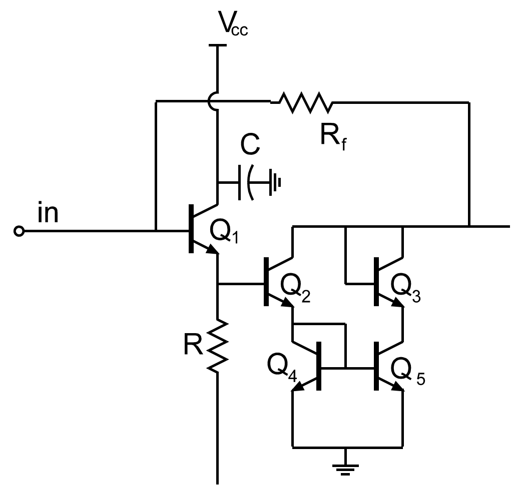




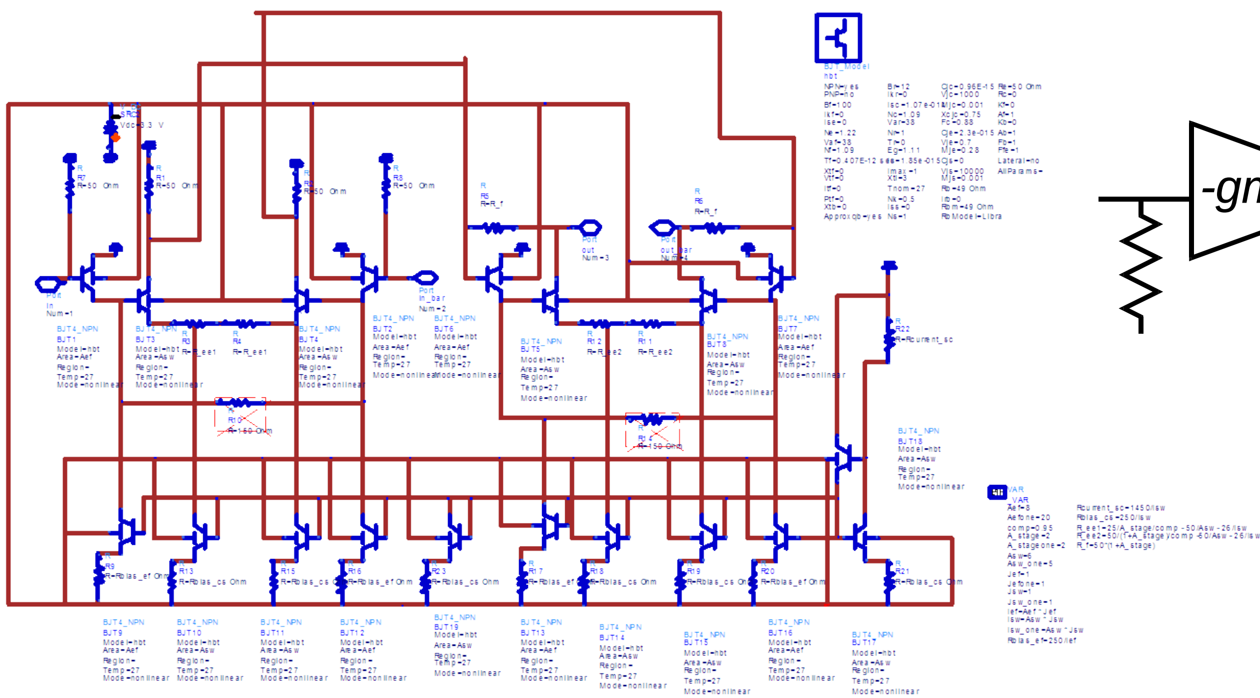


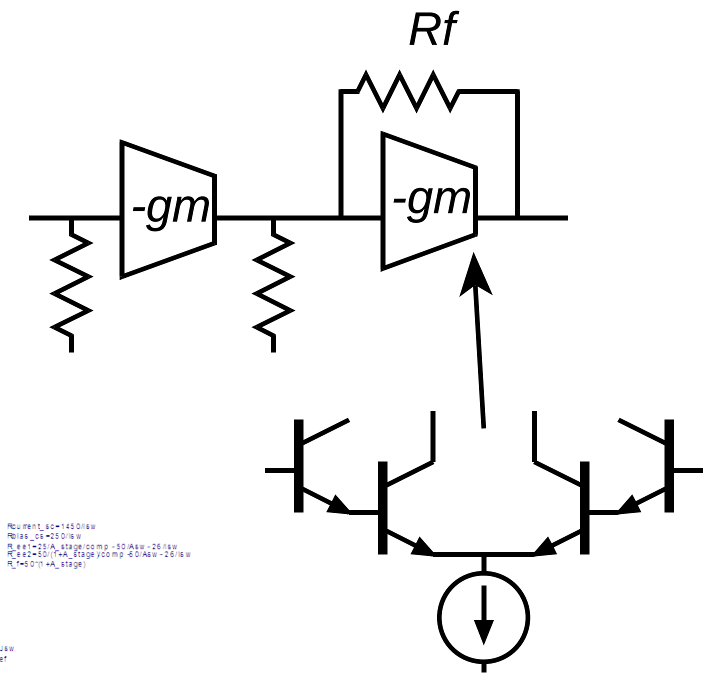
8.2 dB, DC-80 GHz





2-stage differential amplifier





First stage is resistively terminated, second stage has resistive feedback…this is because first stage is allowed to limit, or can have AGC applied, either of which would violate the feedback gain/impedance relationships...

Cherry-Hooper Gm-Zt amplifier

Very popular feedback varaint. Here input/output impedances are not consistent. Input impedance of 2nd stage is

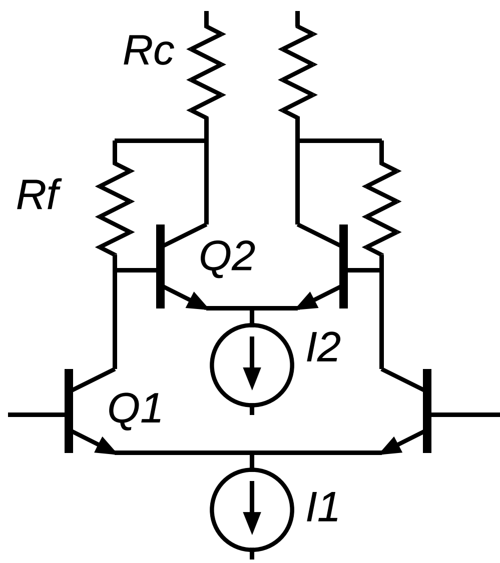
$$R_{in} = (R_c + R_f)/(1 + g_{m2}R_c) \rightarrow 1/g_{m2}.$$

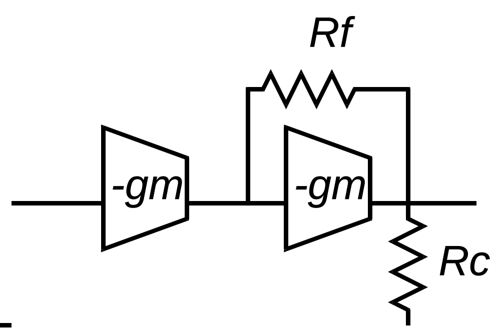
Transimpedance of second stage is (derive these)

$$Z_{T2} = R_f\left(g_m - G_f\right)/\left(g_m + G_C\right) \rightarrow R_f.$$

Overall gain is

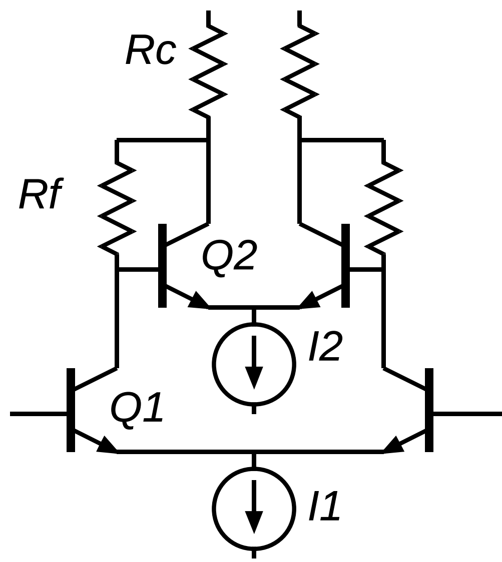
$$A_v = g_{m1}Z_{T2}$$

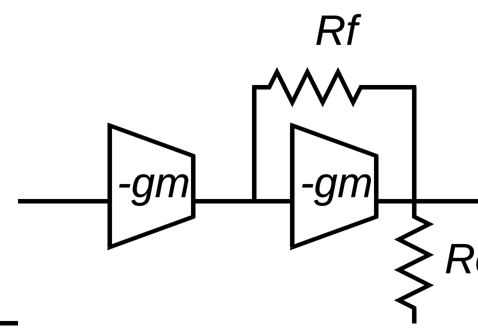




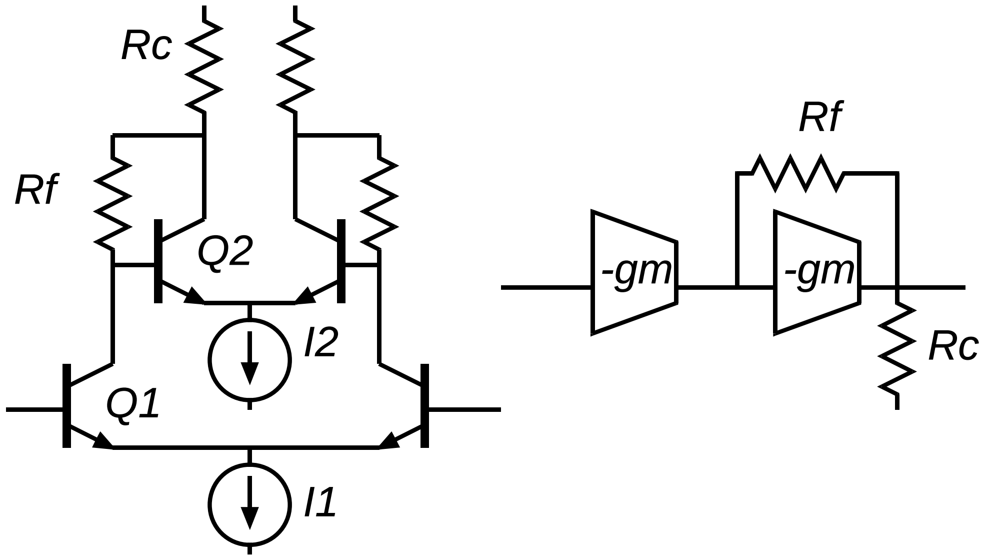
Cherry-Hooper Gm-Zt amplifier

Let us work through impedances and gains both carefully and by back-of-envelope…
…why is first order time constant so small ?



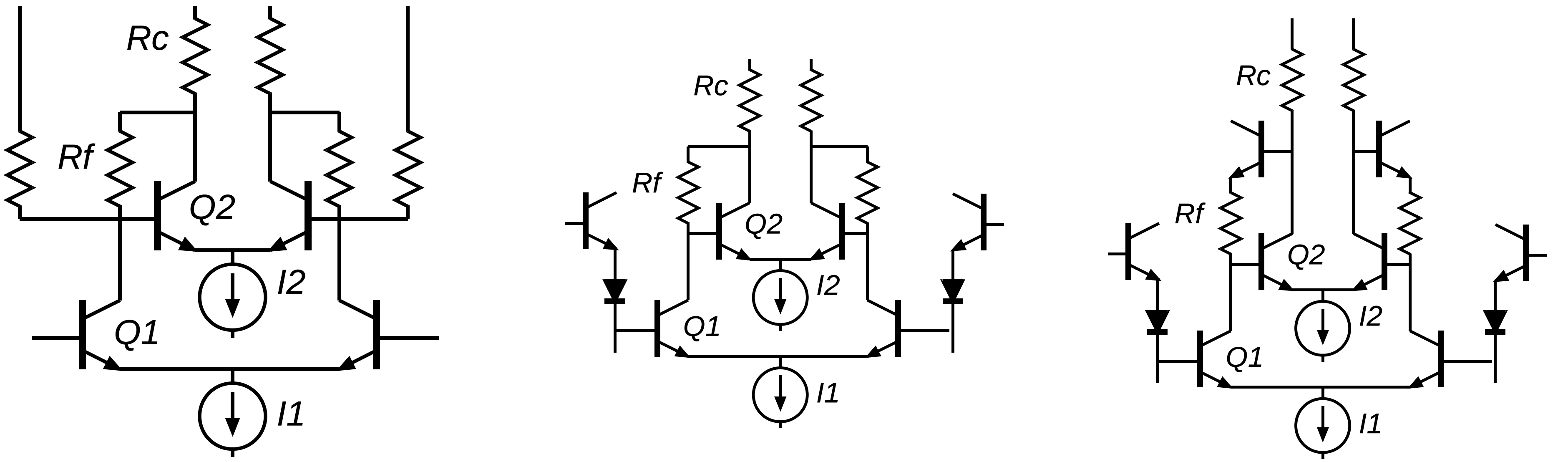


Cherry-Hooper Gm-Zt amplifier--limiting behavior

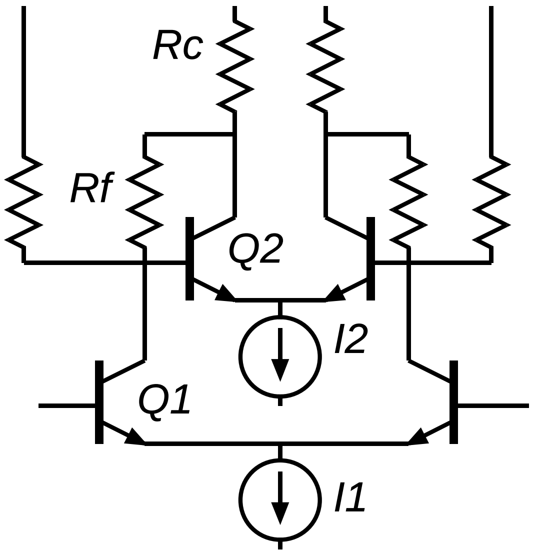


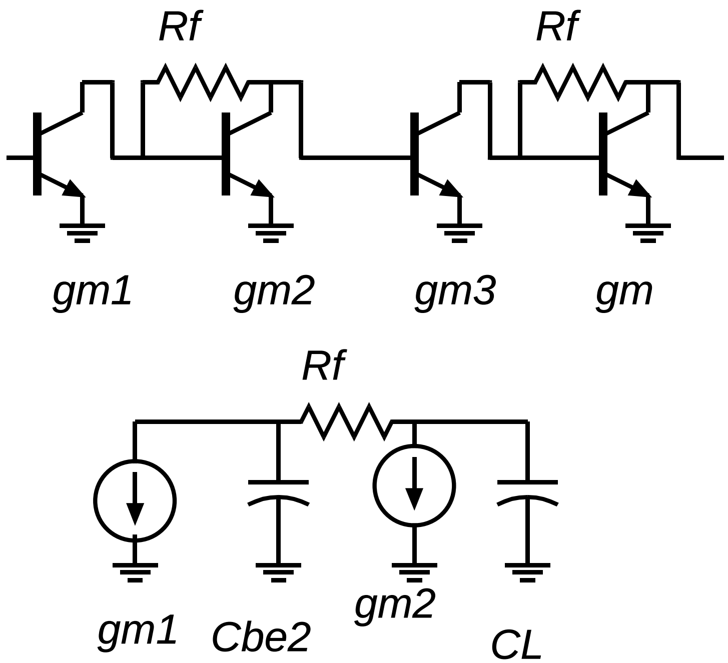
If $I_1 < I_2$ (derive this) amplifier will always limit on the gm stage, not the Zt stage....hence large-signal operation maintains low impedances (hence short time constants) provided by feedback.

Cherry-Hoopers: DC bias variants



Cherry-Hooper bandwidth---zero'th order





By inspection,

$$a_1 = \left(C_{be2} + C_L\right)/g_{m2}$$

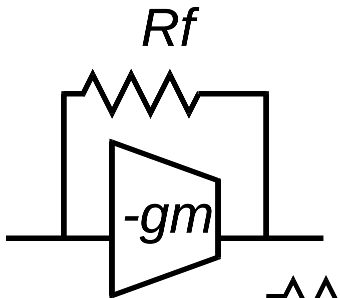
$$a_2 = R_L C_{be2} C_L / g_{m2}$$ (work more detailed example in lecture)

Note immediately that first - order time constant is small,

second - order time constant is NOT.

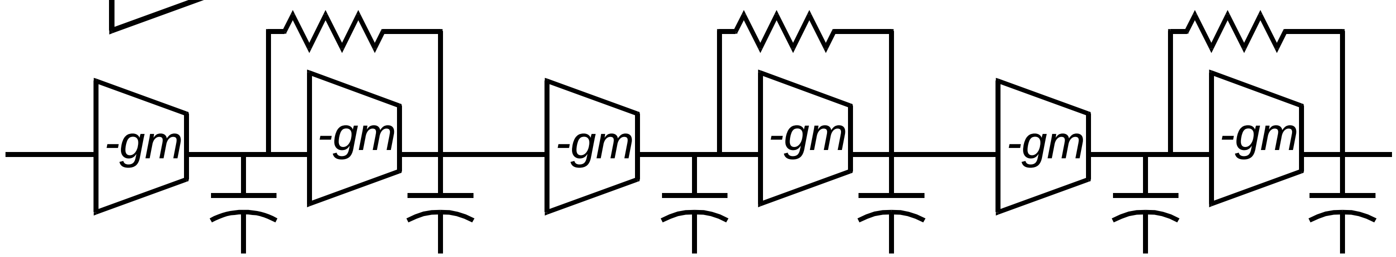
Much more detailed analysis in Hitachi paper....

Gm-Zt amplifier: understanding time constant behavior

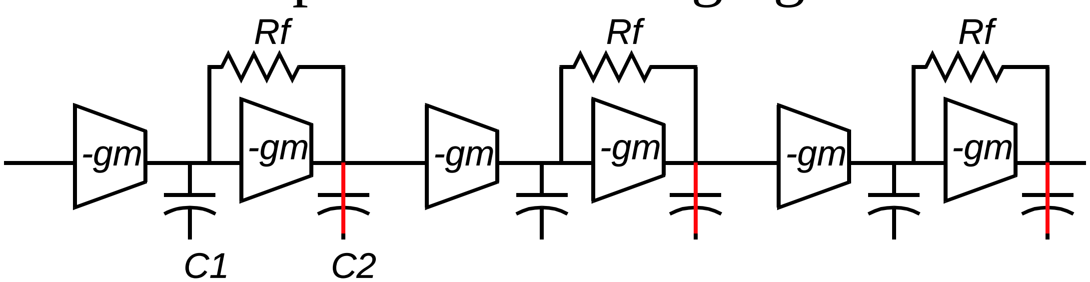


This has input and output impedance of

$$R_{in} = R_{out} = 1/g_{m1}$$



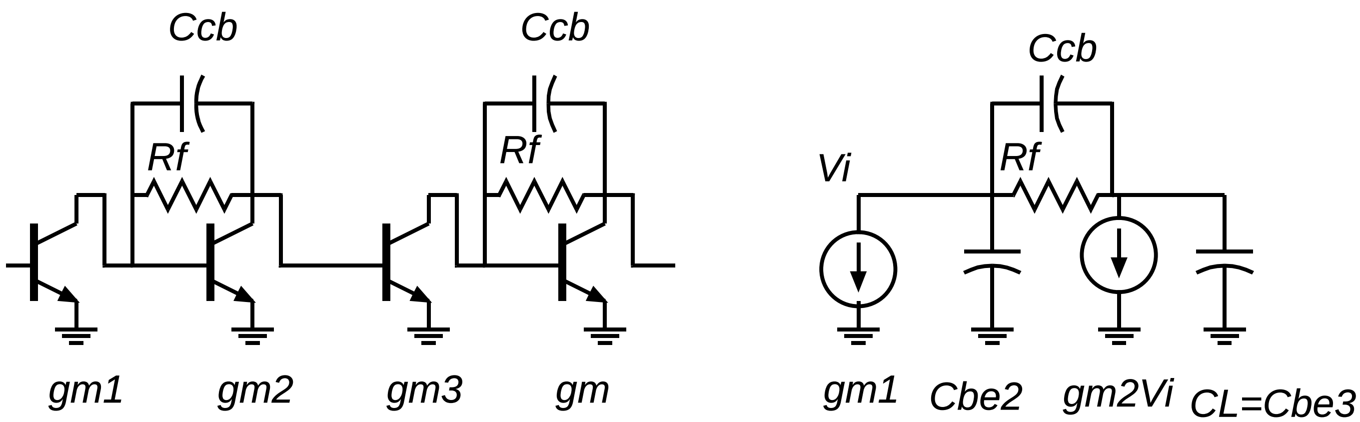
So each capacitor has charging time constant of C/g_{m1}



Shorting C2 makes the C1 charging time constant $= C_1 R_f$

so terms in a2 are of the form $(C_2/g_m)C_1 R_f$.

Cherry-Hooper bandwidth---with Ccb

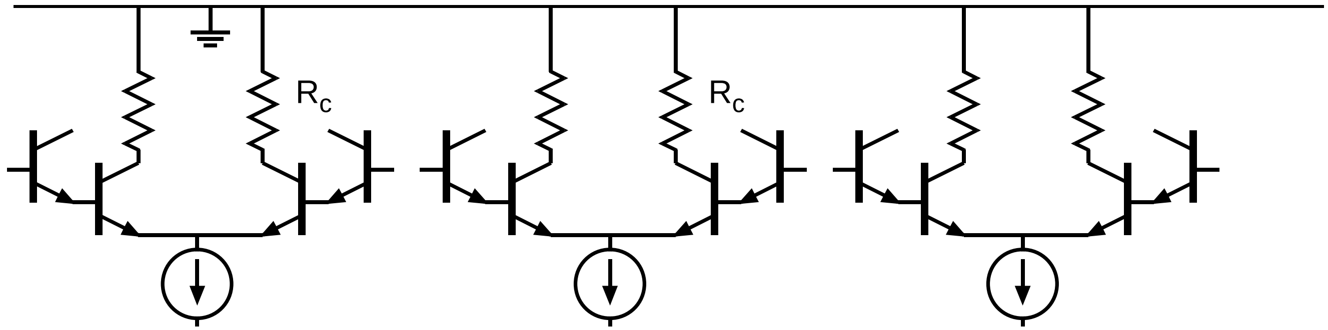


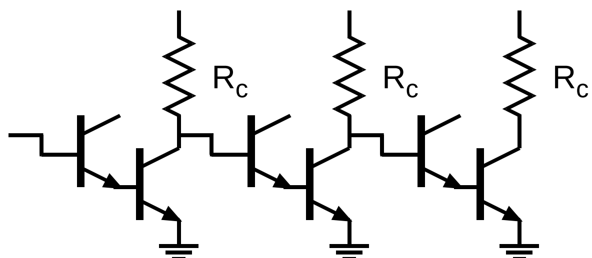
$$a_1 = C_{be2} / g_{m2} + C_{be3} / g_{m2} + C_{cb} R_f$$

$$a_2 = (C_{be2} / g_{m2})(C_{be3} R_f) + (C_{be2} / g_{m2})(C_{cb} R_f) + (C_{be3} / g_{m2})(C_{cb} R_f)$$

note that Ccb has reduced bandwidth and has improved damping

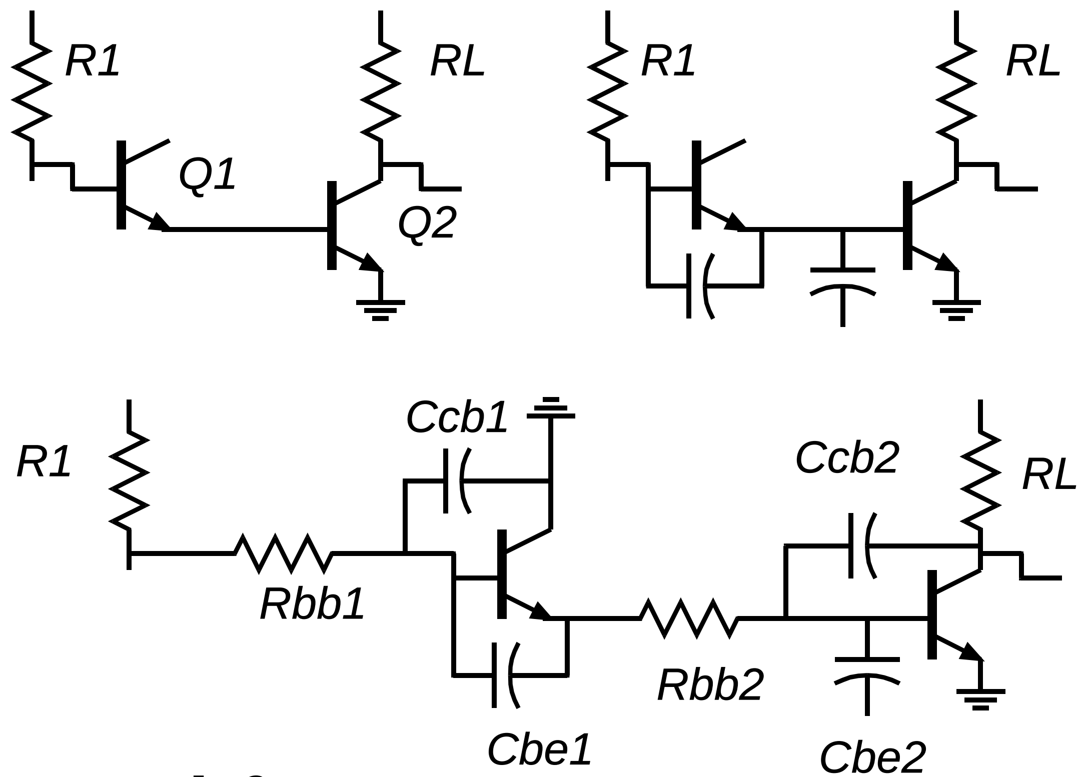
Darlingtons 1





Why not just use emitter followers to buffer the stage input capacitances ????

Darlingtons 2



Taking Rex = 0 and $\beta = \infty$...

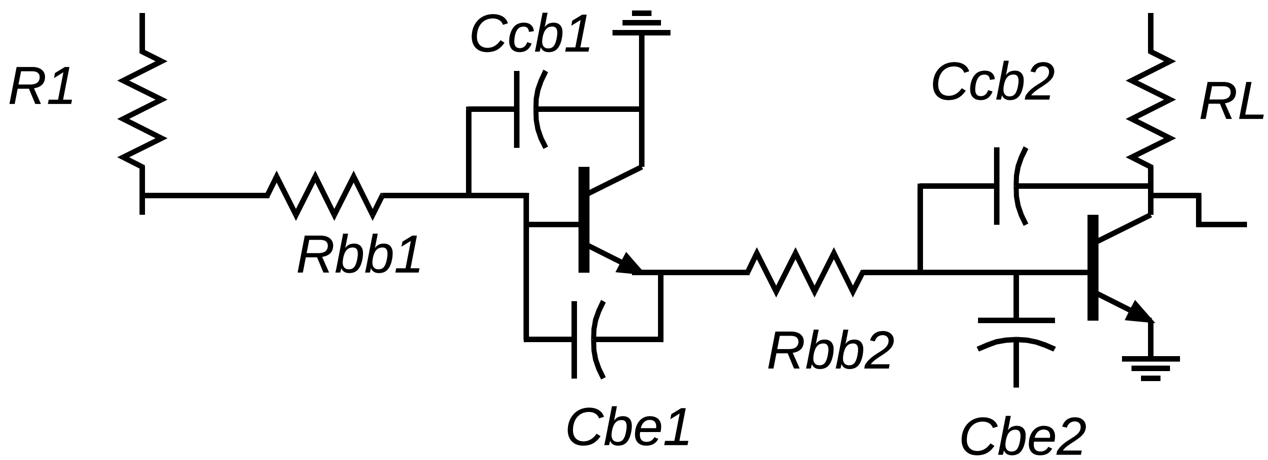
$$a_1 = C_{be2}(R_{bb2} + r_{e1})$$

$$+ C_{cb2}\left((R_{bb2} + r_{e1})(1 + g_{m2}R_L) + R_L\right)$$

$$+ C_{cb1}(R_{bb1} + R_1) + C_{be1}(0!)$$

$$a_2 = C_{be2}C_{be1}(R_{bb2} + r_{e1})(R_1 + R_{bb1})(r_{e1}/(r_{e1} + R_{bb2}) + \ldots$$

Darlingtons 3

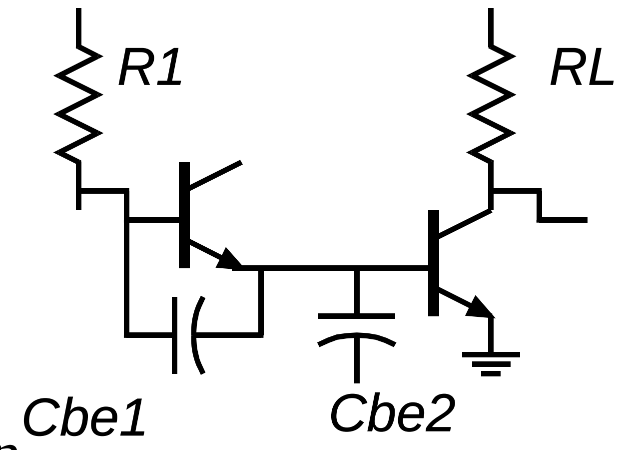


Note the greatly reduced charging time for C_{be2} :

$a_1 = C_{be2}(R_{bb2} + r_{e1})$ instead of $a_1 = C_{be2}(R_{bb2} + R_1)$

This is the motivation for the Darlington commonly given in introductory texts. But a more detailed consideration reveals that (1) the charging time for C_{be2} is at a minimum $C_{be2}R_{bb2}$ (more on this later!) and (2) that $C_{be2}(R_{bb2} + R_1)$ **must reappear** in the second-order time constant, with consequent reduction in circuit damping

Darlingtons



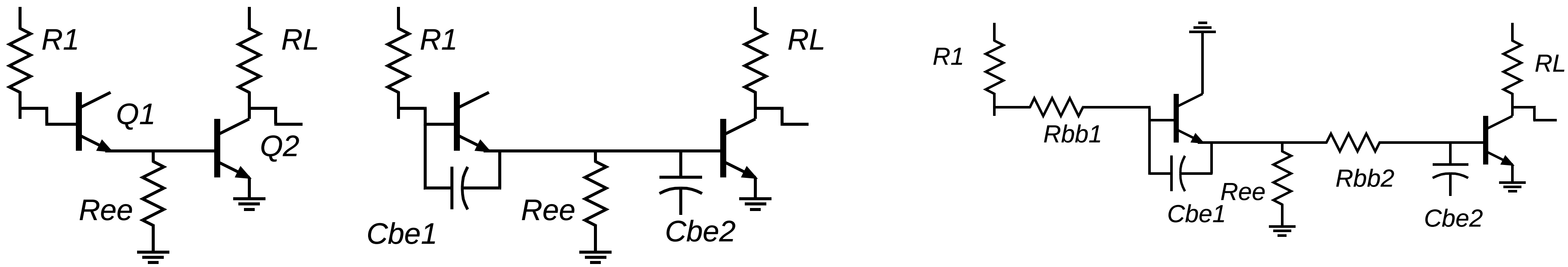
Clearly, exact problem is hard....lets look at effect of C_{be} alone (all other BJT parasitics are zero). Then

$a_1 = C_{be2} r_{e1}$

$a_2 = C_{be2} C_{be1} r_{e1} R_1$

If we have a cascade of stages, then note that $R_L = R_1$ and that gain = A = R_1 / r_{e2}. The damping factor is proportional to the ratio $a_1 / \sqrt{a_2}$. Note we can make the first-order time constant very small, but not the second-order time constant ..(work through...)...attempts to broadband response result in loss of damping

Darlingtons: improving the damping



Addition of Ree has 2 effects. The Q1 EF gain is reduced

$$a_1 = C_{be1}\left(\left(R_1 + R_{bb1}\right)\left(1 - A_{v,ef}\right) + ...\right) + C_{be2}\left(R_{bb2} + R_{ee} \| R_{out,ef}\right)$$

$$a_2 = C_{be1}\left(\left(R_1 + R_{bb1}\right)\left(1 - A_{v,ef}\right) + ...\right) C_{be2}\left(R_{ee} \| \left(R_1 + R_{bb1}\right)\right)$$

So, the terms in a1 associated with Cbe1 are increased. More importantly, terms in a2 associated with Cbe2, (e.g. R_{22}^{1}) are reduced due to the parallel loading of R_{ee}

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