Mixed Signal IC Design
Notes set 4:
Broadband Design Techniques

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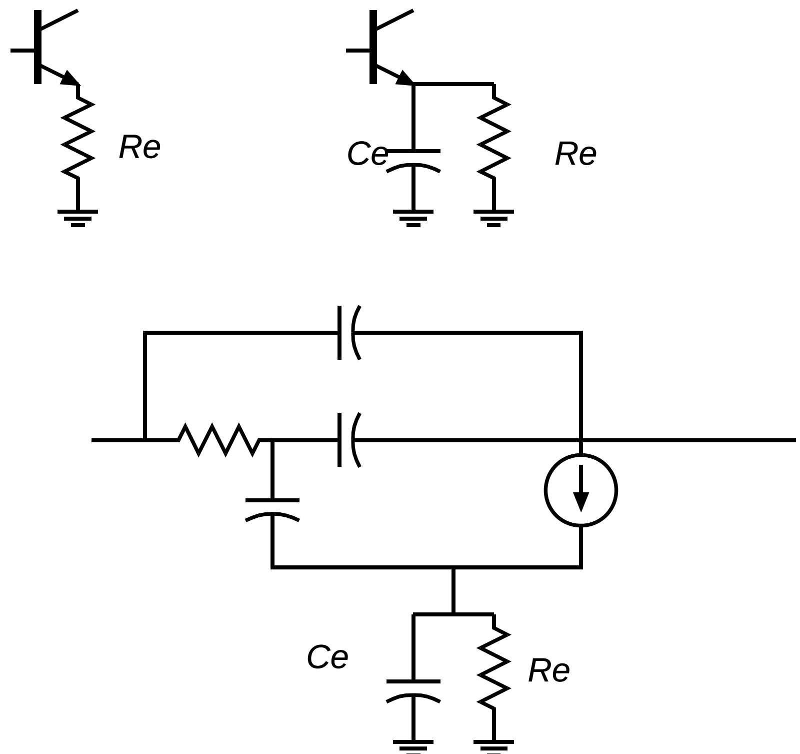
Getting more bandwidth

At this point we have learned basics (MOTC etc)

How can we get more bandwidth…. ?

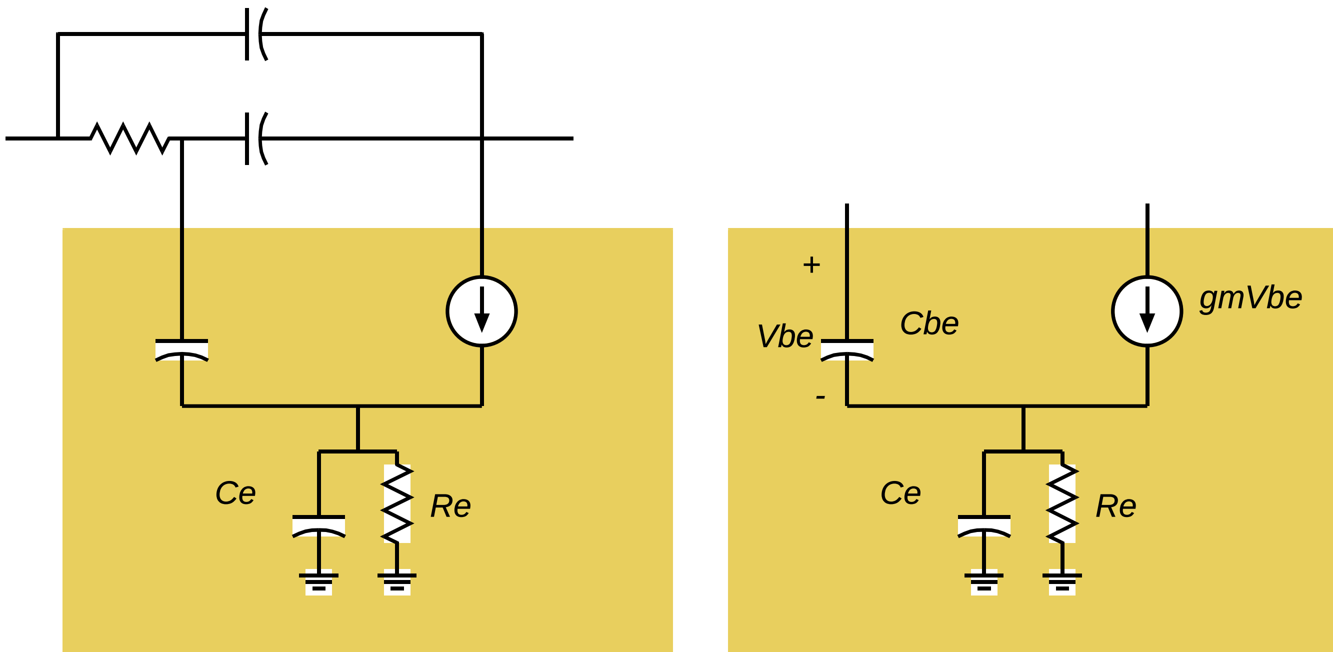
- Resistive feedback
- transconductance-transimpedance
- emitter-follower buffers, benefits and headaches
- emitter degeneration…basics
- emitter degeneration and area scaling
- ft-doubler stages..
- distributed amplifiers
- broadbanding / peaking with LC networks….
- examples….

Emitter Degeneration, I



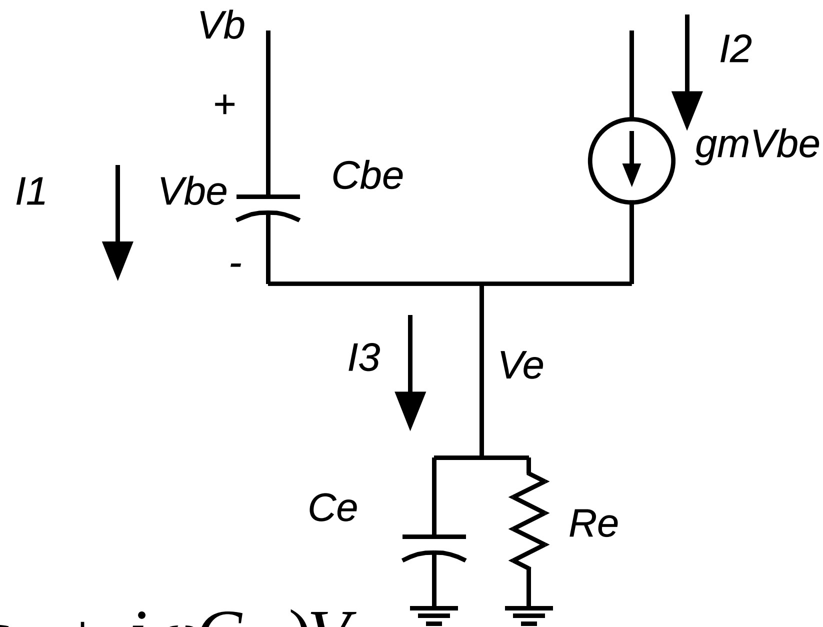
Emitter degeneration was introduced using only an emitter resistance. We will find we instead want an RC combination such that $R_e C_e = C_{be} / g_m$.

Emitter Degeneration, 2



Next note that the topology is such that we can work with R_{bb}, C_{cb} etc removed, and add them back later....

Emitter Degeneration, 3



Working from Vbe:

$I_1 = j\omega C_{be} V_{be} \quad I_2 = g_m V_{be}$ so $I_3 = (g_m + j\omega C_{be}) V_{be}$

But $V_E = I_3 \left(1/R_E + j\omega C_E\right)^{-1} = V_{be} * g_m R_E$

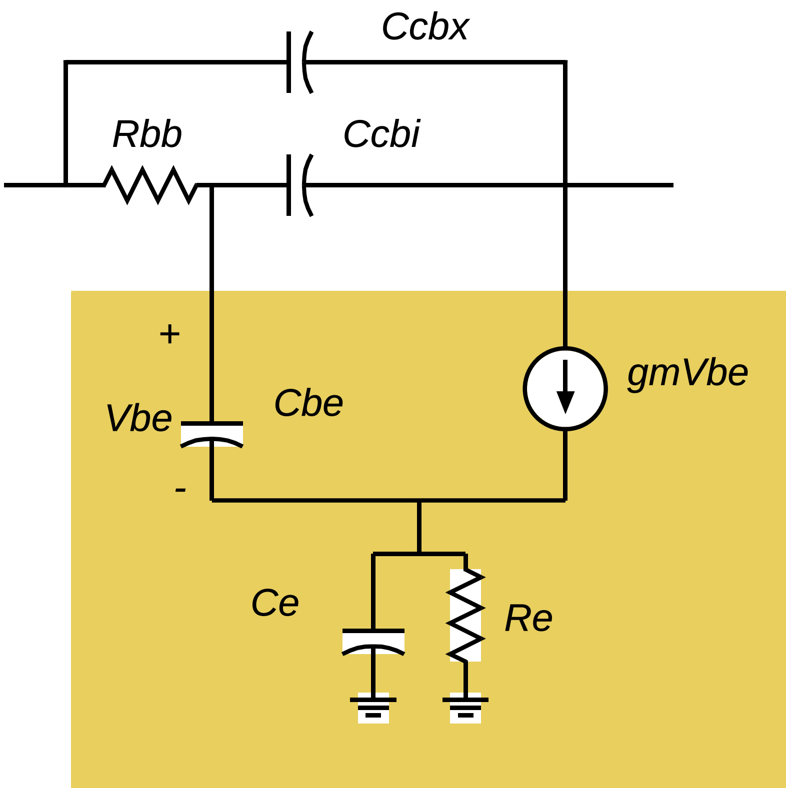
(note that the frequency dependent terms cancel)

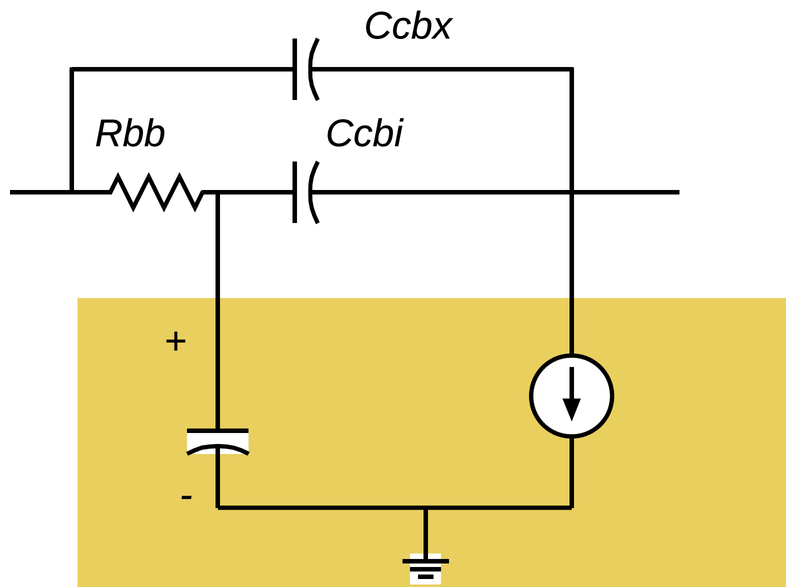
So, $I_2 = V_{be}/(R_E + 1/g_m)$ and

$Z_{in} = j\omega C_{in} = j\omega C_{be}(1 + g_m R_E)^{-1}$

R_{bb} and C_{cb} etc can now be added back...they dont change

Emitter Degeneration, Summary



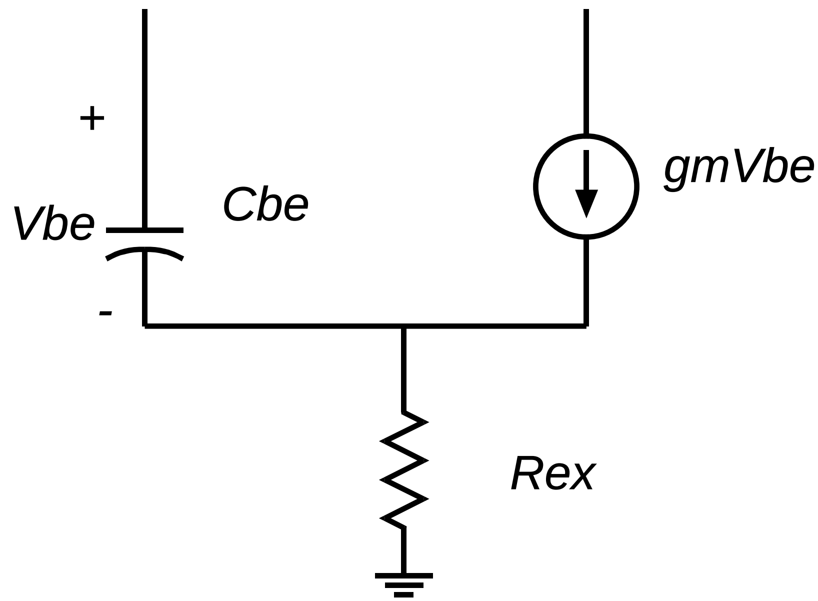


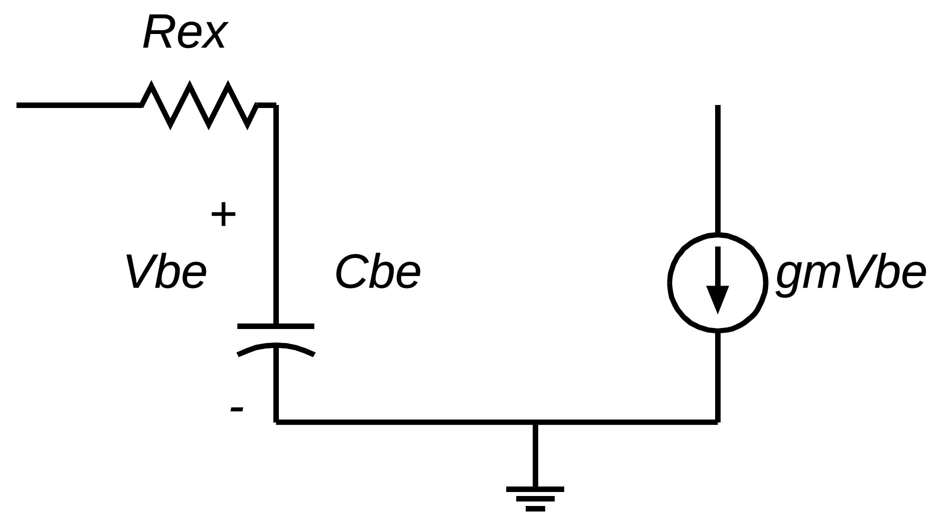
$$C_{in} = C_{be}(1 + g_m R_E)^{-1}$$

$$g_{m,extrinsic} = g_m (1 + g_m R_E)^{-1}$$

Other element values do not change...

Emitter Degeneration, Unbypassed emitter resistance

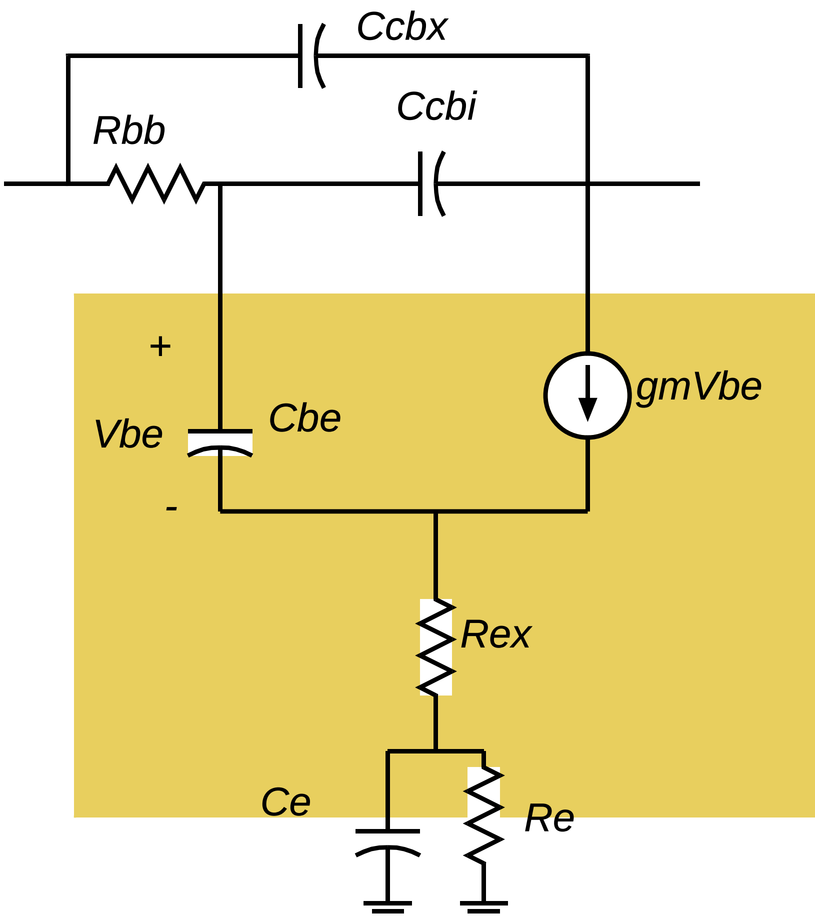


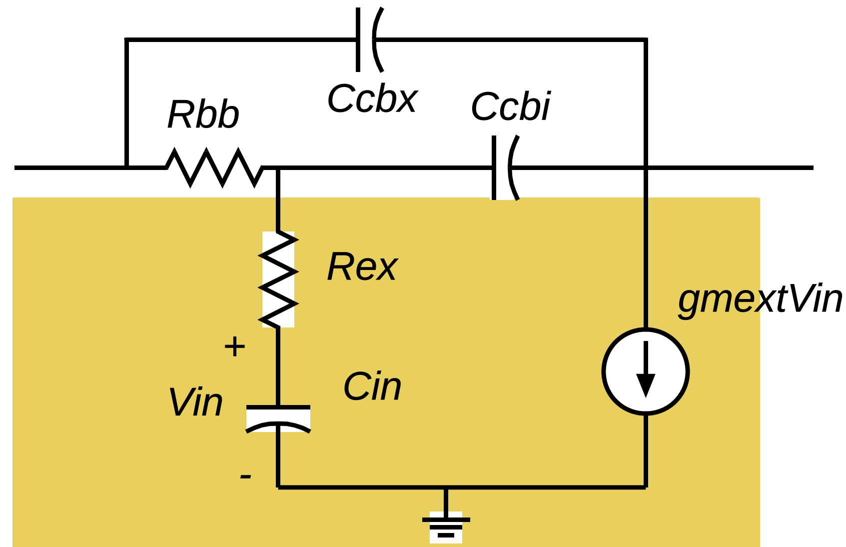


$$g_{m,extrinsic} = g_m (1 + g_m R_E)^{-1}$$

can be shown by elementary nodal analysis, work through…
Key observation: degeneration by this method results in a series input resistance, which can increase Cbe charging time.

Emitter Degeneration, combination case

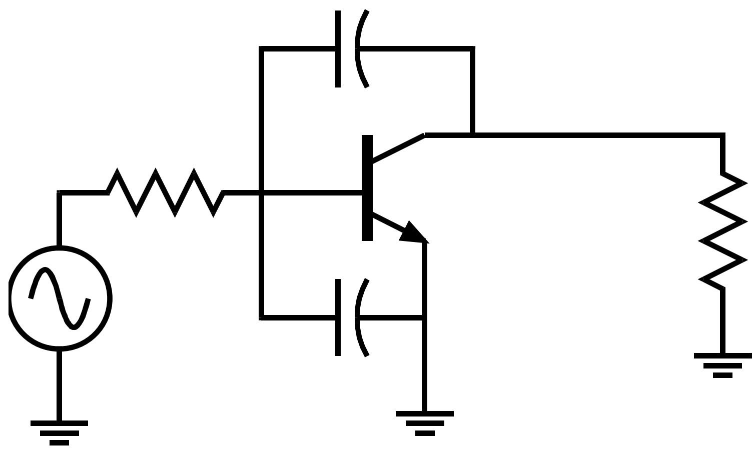




$$C_{in} = C_{be}(1+g_m R_E)^{-1}$$

$$g_{m,extrinsic} = g_m(1+g_m R_E)^{-1}$$

Emitter Degeneration and Area Scaling 1



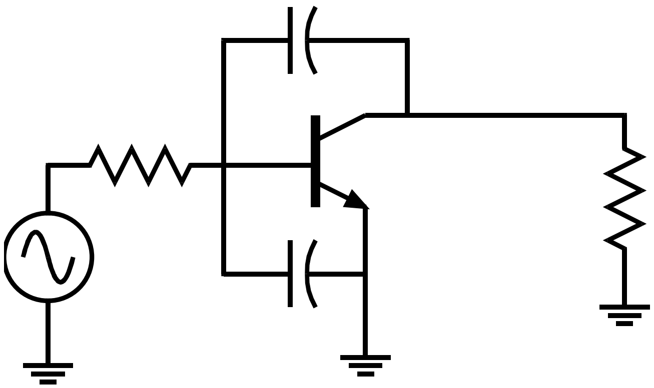
emitter area =Ae
current density =Je
*Current =Ae*Je=Ie*
transconductance=gm
input capacitance=Cbe
collector capacitance=Ccb
base resistance=Rbb

Consider the CS stage. If we desire a gain $A_V = -g_{m,ext} R_L$,

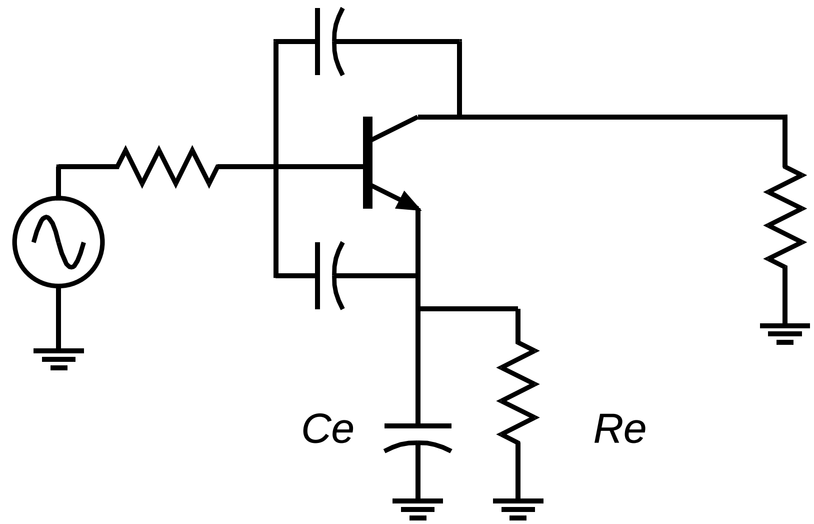
then we have

$$a_1 = C_{be}(R_{gen} + R_{bb}) + C_{cb}\left[(R_{gen} + R_{bb})(1 + g_{m,ext} R_L) + R_L\right]$$

Emitter Degeneration and Area Scaling 2



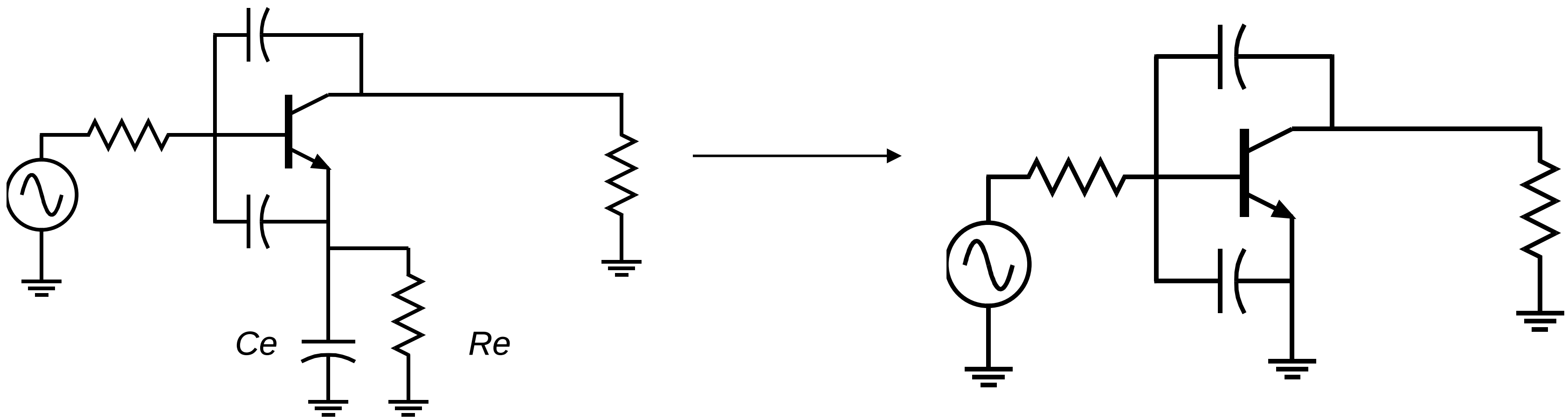
emitter area =Ae
current density =Je
Current =Ae*Je=Ie
transconductance=gm
input capacitance=Cbe
collector capacitance=Ccb
base resistance=Rbb



Bigger transistor, before degeneration
emitter area =kAe
current density =Je
Current =kAe*Je=Ike
transconductance=kgm
input capacitance=kCbe
collector capacitance=kCcb
base resistance=Rbb/k

now degenerate:
Re picked such that (1+gmRe)=k
intrinsic transconductance=kgm
extrinsic transconductance=kgm/(1+gmRe)=gm
input capacitance=kCbe/(1+gmRe)=Cbe
collector capacitance=kCcb
base resistance=Rbb/k

Emitter Degeneration and Area Scaling 3



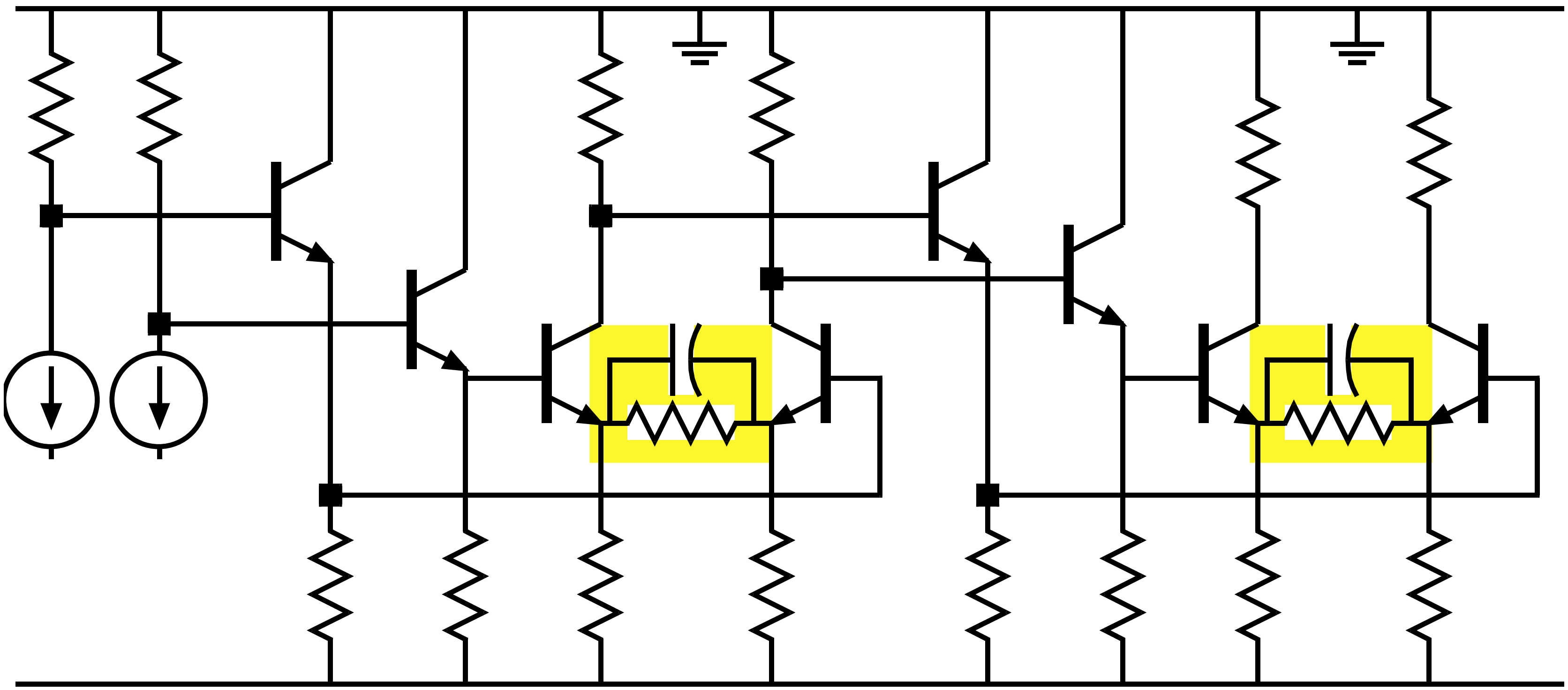
We have made the transistor K times bigger, then degenerated by K:1. We are left with the original gm and Cin, but have increased Ccb by K:1 and decreased Rbb by K:1

$$a_1 = C_{be}\left(R_{gen} + \frac{R_{bb}}{k}\right) + kC_{cb}\left[\left(R_{gen} + \frac{R_{bb}}{k}\right)(1 + g_{m,ext}R_L) + R_L\right]$$

$$a_1 = R_{gen}C_{be} + C_{cb}\left[R_{bb}(1 + g_{m,ext}R_L)\right] + \frac{C_{be}R_{bb}}{k} + kC_{cb}\left[R_{gen}(1 + g_{m,ext}R_L) + R_L\right]$$

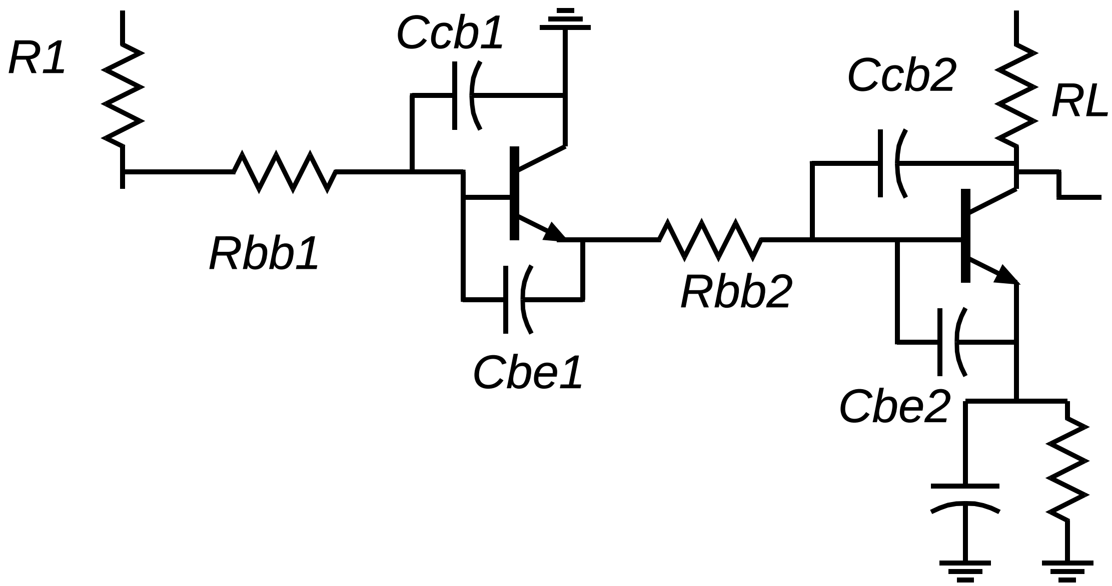
So, appropriate choice of k exchanges delay terms associated with Rbb vs Ccb, and gives the highest bandwidth

Emitter Degeneration and Area Scaling 4: Emitter Follower Case



Why might we want to do this ?

Emitter Degeneration and Area Scaling 5: Emitter Follower Case



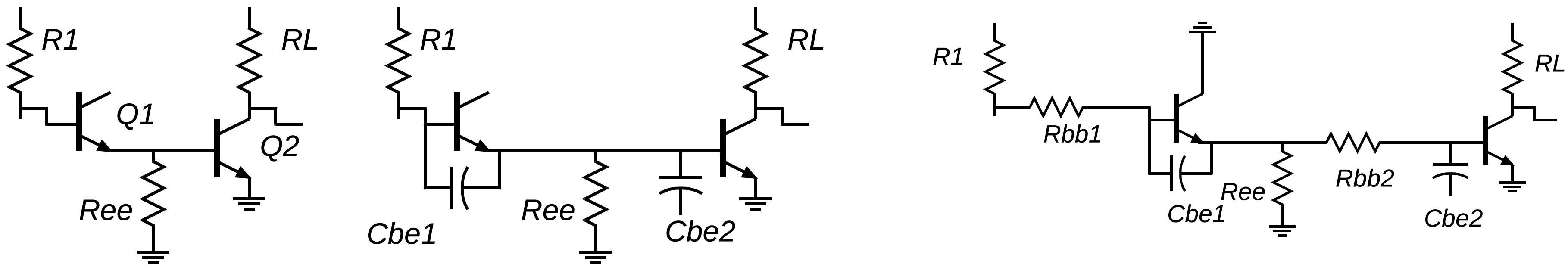
Considering only terms in a1 associated with the CE stage,

$$a_1 = ... + kC_{cb2}\left(\left(\frac{R_{bb2}}{k} + R_{out,1}\right)\left(1 + g_{m,ext}R_L\right) + R_L\right) + C_{be2}\left(\frac{R_{bb2}}{k} + R_{out,1}\right)$$

Note that the undegenerated $R_{bb}C_{be}$ time constant of an HBT is several times $1/2\pi f_\tau$, and can be a major bandwidth limit.

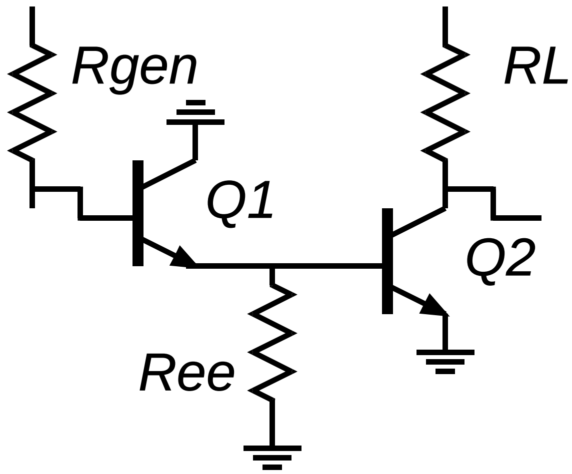
Area scaling/degeneration as above is one method to address this.

Ft-doubler stages....the Darlington Revisited

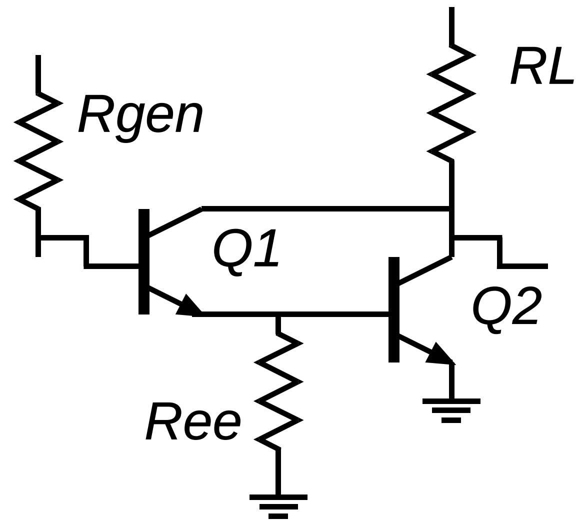


Recall the improvement in Damping provided by Ree….
…is there a larger lesson here ?

2 Kinds of Darlington:



EF collector grounded



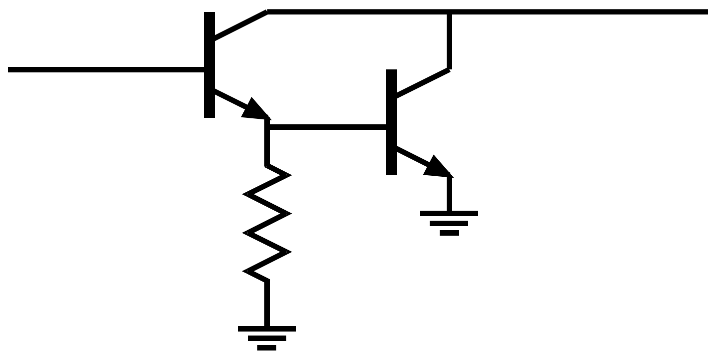
Collectors tied together

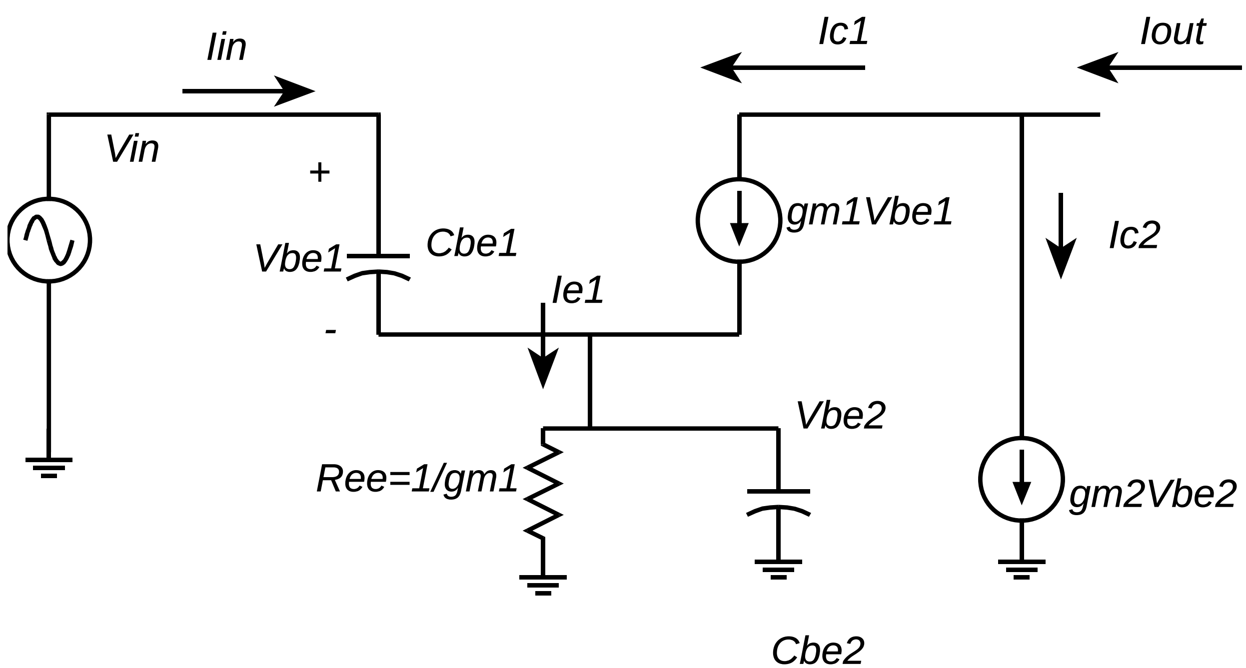
If collectors are tied together:

AC collector currents of both transistors contribute to output :)

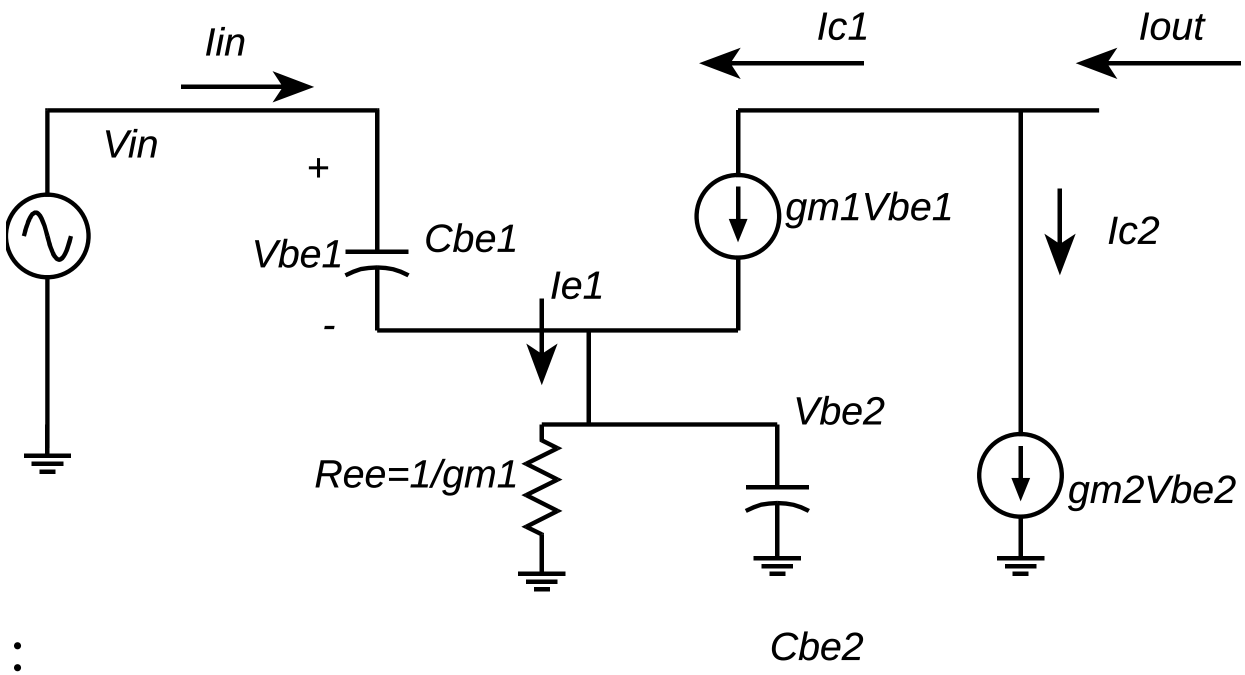
Both transistor Ccb's undergo Miller multiplication :(

Ft-doubler stages 2





Ft-doubler stages 3



Work from V_{be1} :

$$I_{in1} = j\omega C_{be} V_{be1} \;,\; I_{c1} = g_m V_{be1} \,,\; I_{e1} = g_m V_{be1}(1 + j\omega C_{be}/g_m)$$

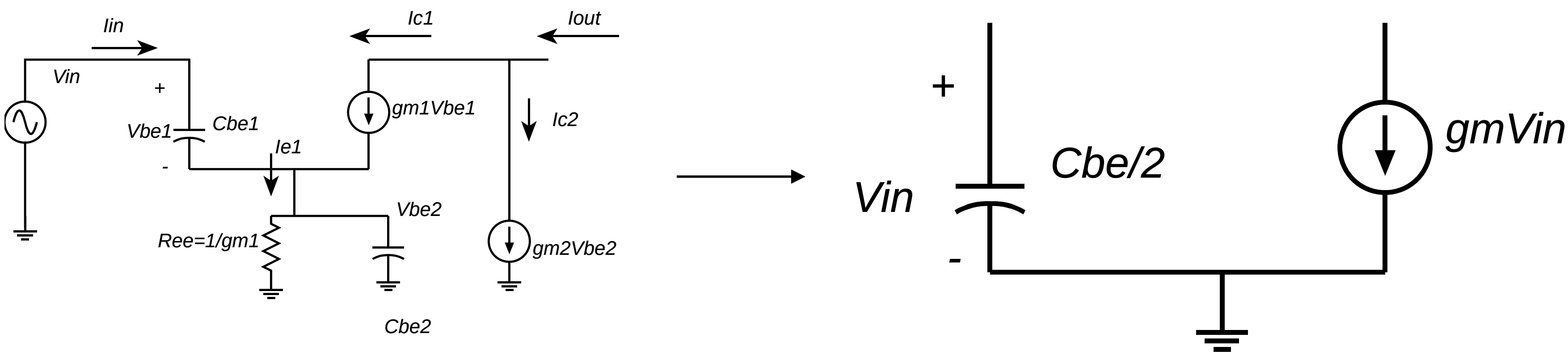
$$V_{be2} = I_{e1} R_{ee} /(1 + j\omega C_{be} R_{ee}) = g_m V_{be1} R_{ee} \frac{(1 + j\omega C_{be}/g_m)}{(1 + j\omega C_{be} R_{ee})}$$

Choose R_{ee} such that $C_{be} R_{ee} = C_{be}/g_{m1}$...hence $R_{ee} = 1/g_m$

$$V_{be2} = (g_m R_{ee}) V_{be1} = V_{be1} \text{...so } V_{be1} = V_{be2} = V_{in}/2$$

$$I_{in} = j\omega C_{be}/2 \text{ and } I_{\text{out}} = g_m V_{be} \text{ hence } I_{out}/I_{in} = 2g_m / j\omega C_{be}$$

Ft-doubler stages 4



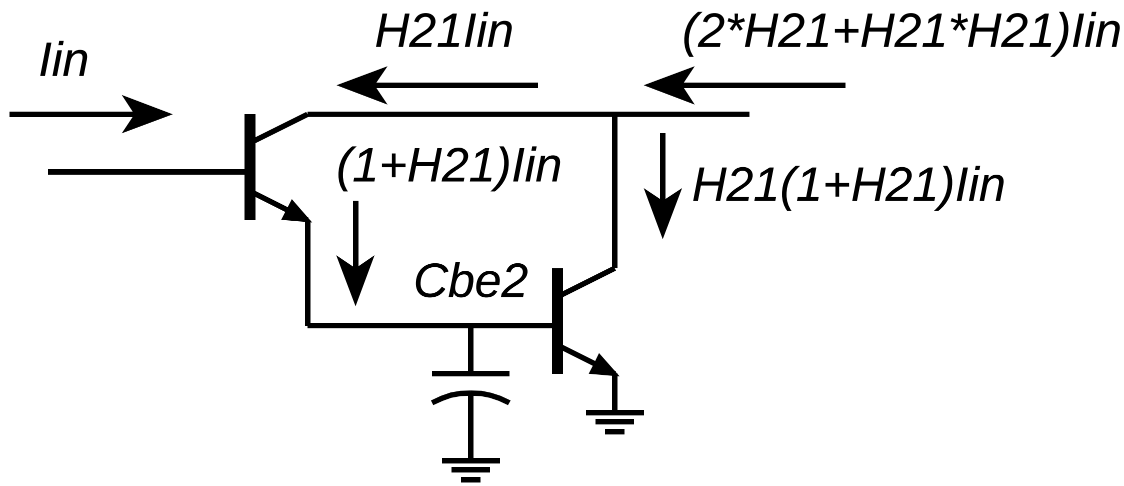
$$I_{in} = j\omega C_{be}/2 \text{ and } I_{\text{out}} = g_m V_{be}$$

$$I_{out}/I_{in} = 2g_m / j\omega C_{be} = 2f_\tau / jf$$

..we have doubled f_τ !

Note that this analysis has ignored Rbb and Ccb...we have certainly not doubled fmax, nor will we have doubled overall circuit bandwidth, given the presence of these other parasitics.

ft-doubler vs Darlington...thinking in terms of current gains:



Current gain is $I_{out} / I_{in} = 2H_{21} + (H_{21})^2$ where $H_{21} = f_\tau / jf$

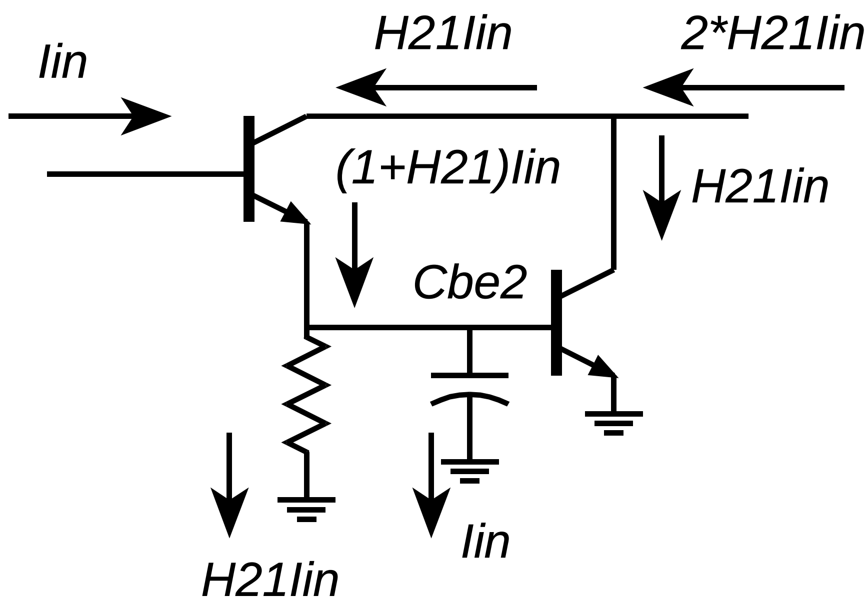
Since $I_{out} = I_{c1} + I_{c2} = g_m V_{be1} + g_m V_{be2} = g_m (V_{be1} + V_{be2}) = g_m V_{in}$

$$I_{in} / V_{in} = Y_{in} = \frac{g_m}{2H_{21} + (H_{21})^2} = \frac{g_m}{2(f_\tau / jf) + (f_\tau / jf)^2} = \ldots$$

Current gain varies at lower frequencies as $(f_\tau / jf)^2$

which gives high current gain but negative input conductance

ft-doubler vs Darlington...thinking in terms of current gains:



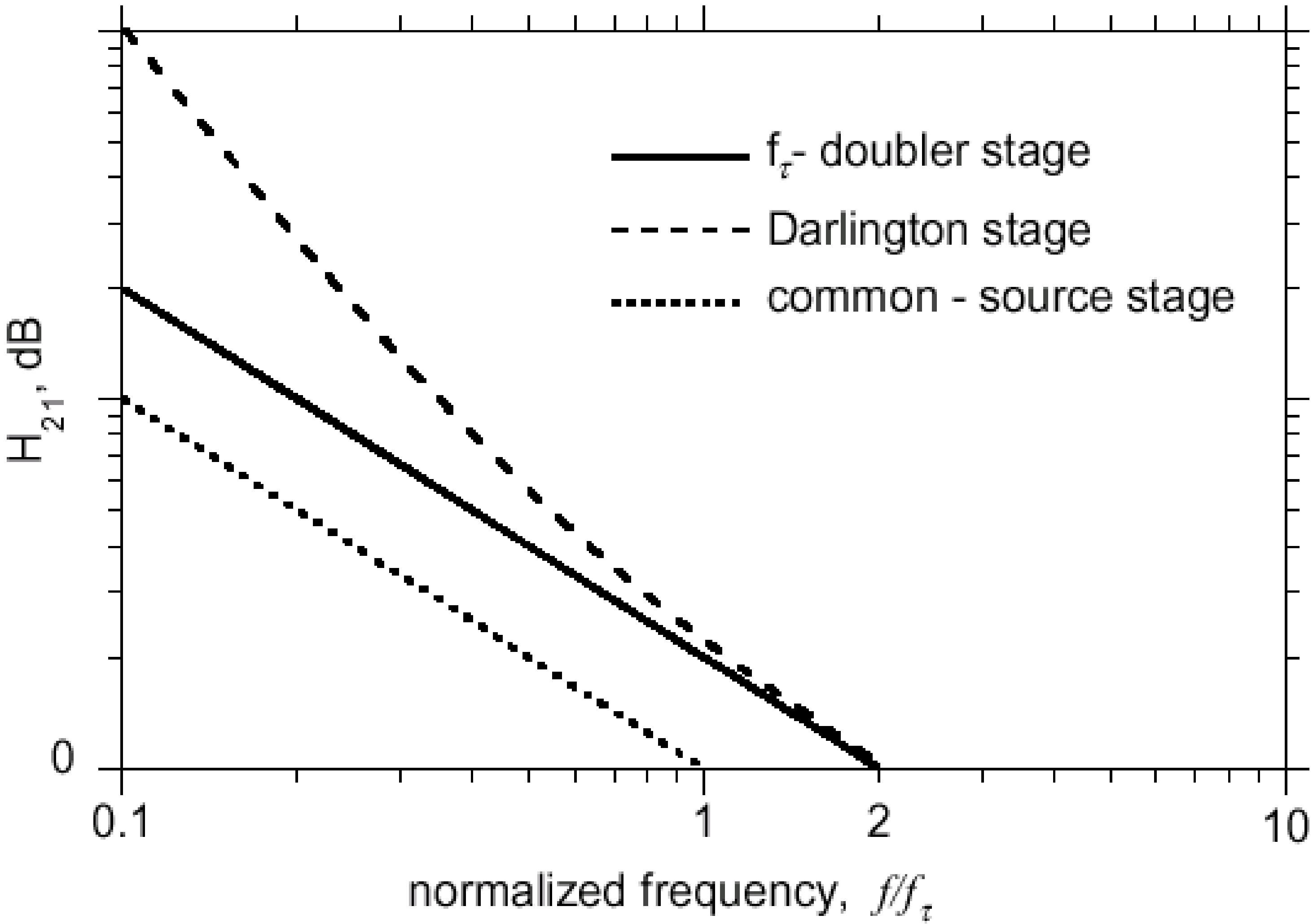
Current gain is $I_{out}/I_{in}=2H_{21}$ where $H_{21}=f_\tau/jf$

Since $I_{out}=I_{c1}+I_{c2}=g_mV_{be1}+g_mV_{be2}=g_m(V_{be1}+V_{be2})=g_mV_{in}$

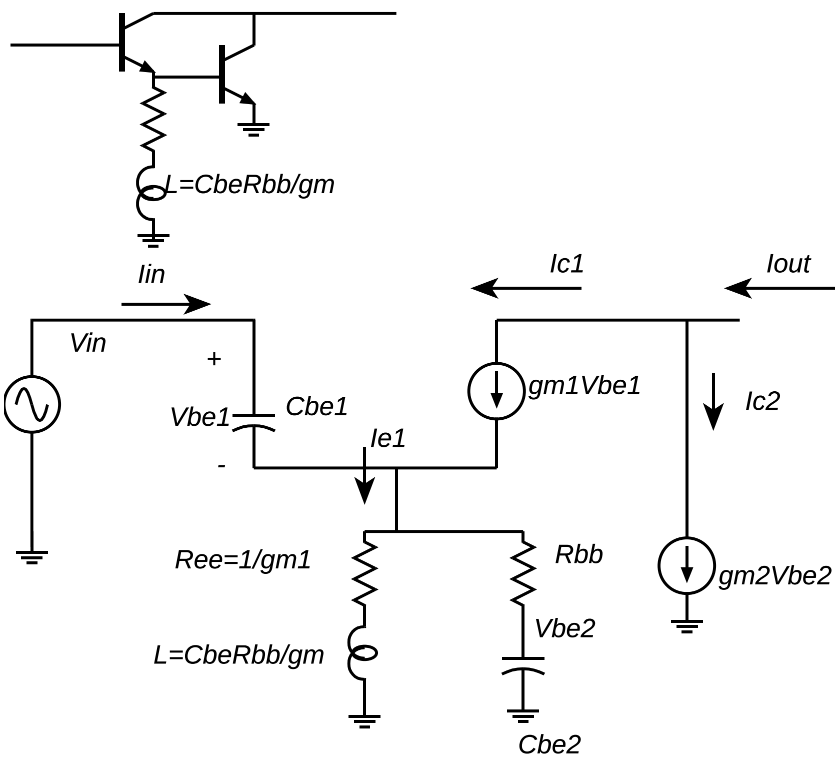
$$I_{in}/V_{in}=Y_{in}=\frac{g_m}{2H_{21}}=\frac{g_m}{2(f_\tau/jf)}=\frac{jfg_m}{2f_\tau}=jf(2\pi C_{be}/2)$$

Current gain varies at all frequencies as $(2f_\tau/jf)$

ft-doubler vs Darlington...current gains:



ft-doubler ...correction for Base resistance

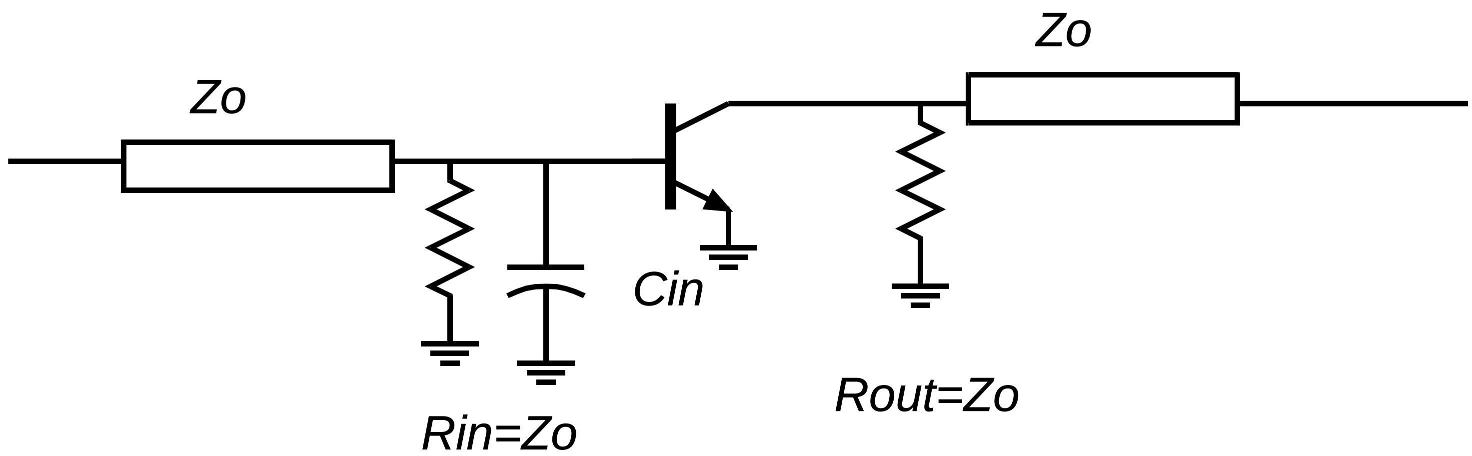


The indicated inductance is necessary for equal current splitting between the 2 transistors in the presence of Rbb. In this case the input impedance is Cbe/2 in series with Rbb/2…again note that this correction is significant because RbbCbe>>Cbe/gm

ft-doubler vs Darlington

What is the conclusion of all this ? We can view the ft-doubler as a special case of the connected-collector Darlington with heavy emitter-follower loading: in this case the second-order (resonant) response is entirely suppressed. More generally we might vary the EF loading to vary the damping. (The particular case of power amplifiers favors the ft-doubler due to efficiency reasons…)

Tuning: to prevent reflections, to peak gain



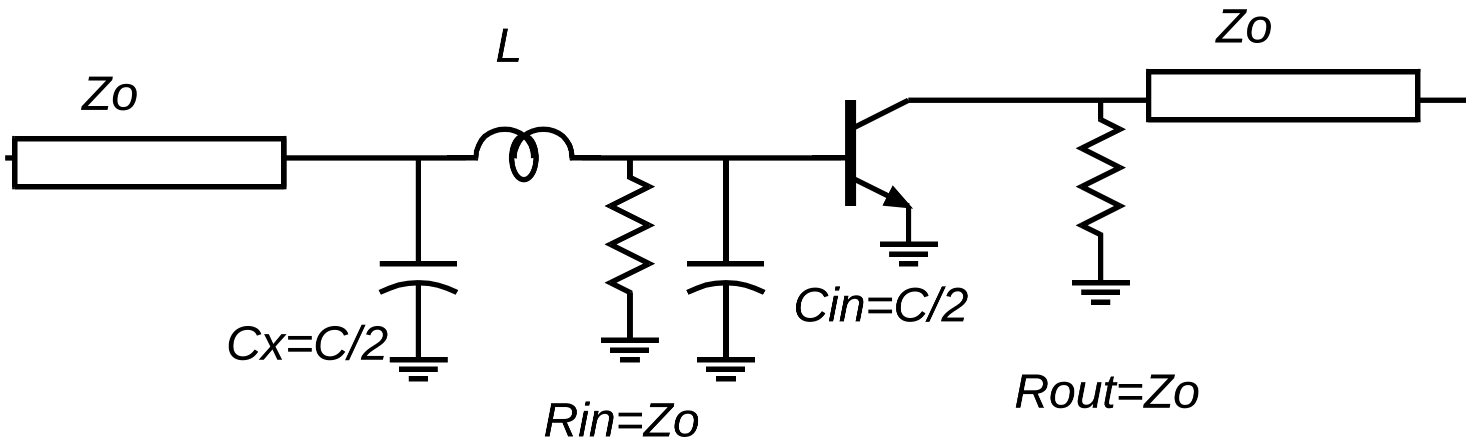
$$S_{21} = \frac{-g_m Z_o / 2}{1 + j\omega C Z_0 / 2}$$

$$S_{11} = \frac{j\omega C Z_0 / 2}{1 + j\omega C Z_0 / 2}, \text{ where } Z_{in} = \left(Z_o \left\| \frac{1}{j\omega C} \right. \right)$$

The input capacitance has hurt both S11 and S21,

input match and gain

Pi-section input tuning....



Absorb the transistor input capacitance into a pi-section

$C_{in} = C_{be} = C/2 = \tau / Z_o 2$

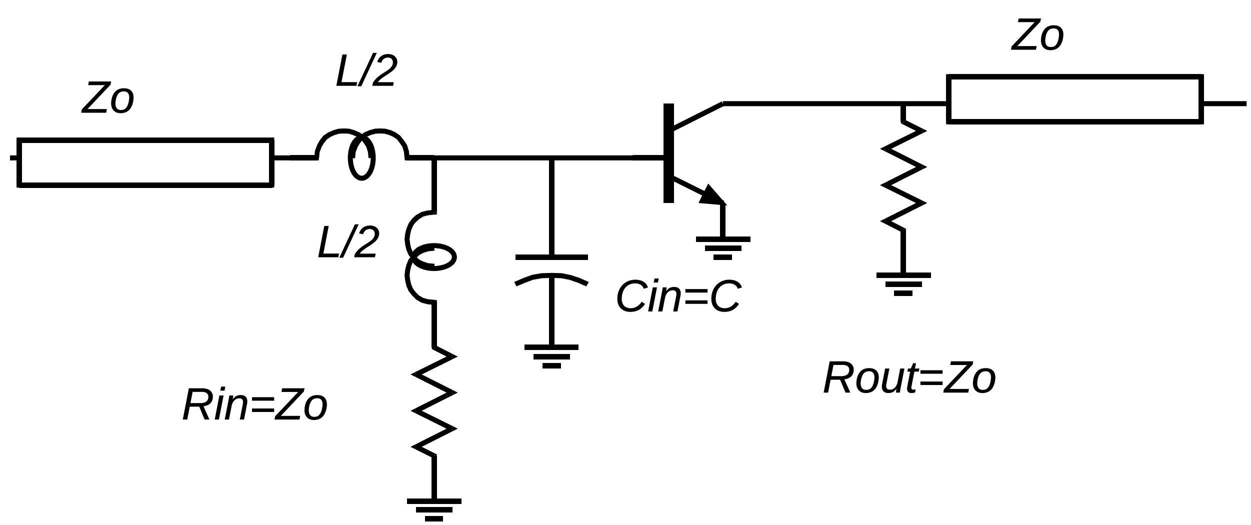
$L = \tau Z_o = (2C_{in} Z_o) Z_o$

Effective for freqeucies such that the pi-section is short

(hence $f_0 < 1/\pi Z_o C$)

Gain flatness and match improved for f<fo, but made worse for f>f0

T-section input tuning....



Absorb the transistor input capacitance into a T-section

$C_{in} = C_{be} = C = \tau / Z_o$

$L = \tau Z_o = (2C_{in}Z_o)Z_o$

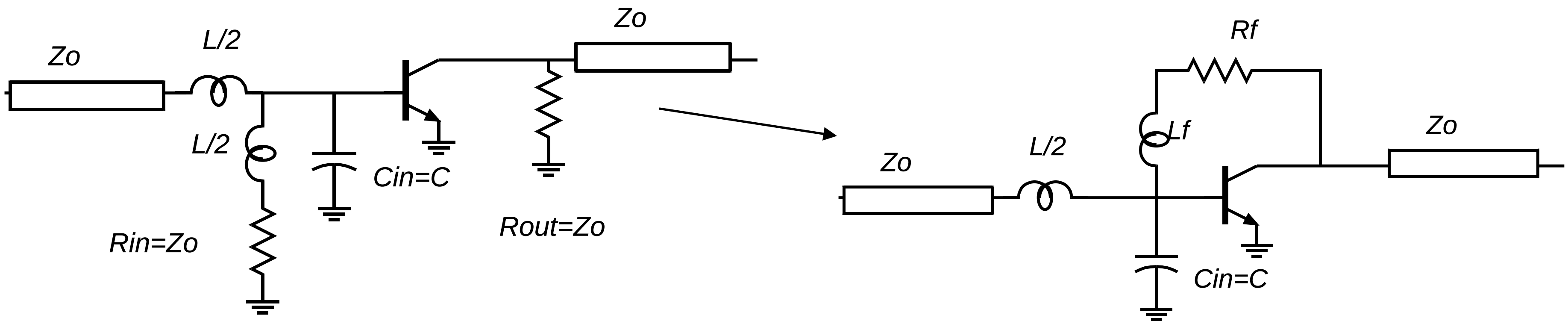
Effective for freqeucies such that the pi-section is short

(hence $f_0 < 1/\pi Z_o C$)

Note that for a given Cin, we have doubled fo relative to

the pi-section case...

T-section tuning in feedback amplifiers



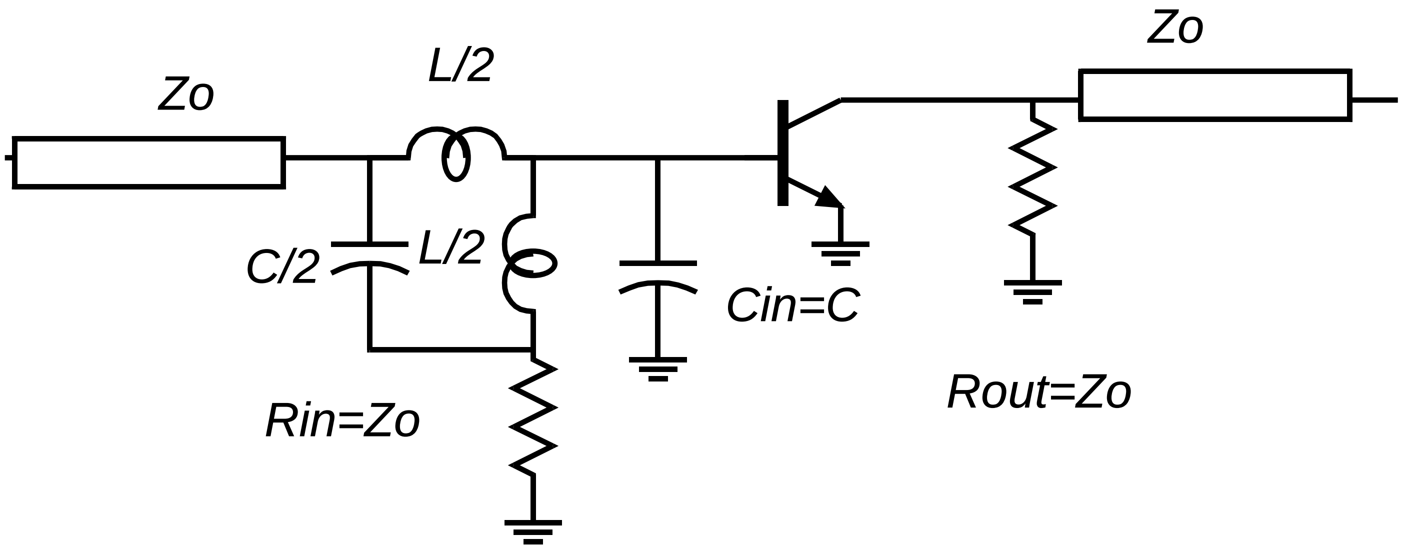
By Miller Approximation :

$$R_f = Z_o(1 - A)$$

$$L_f = (L/2)(1 - A)$$

This is frequently used in 50 Ohm FBAs...danger is reduced stability.

Bridged-T-section input tuning....

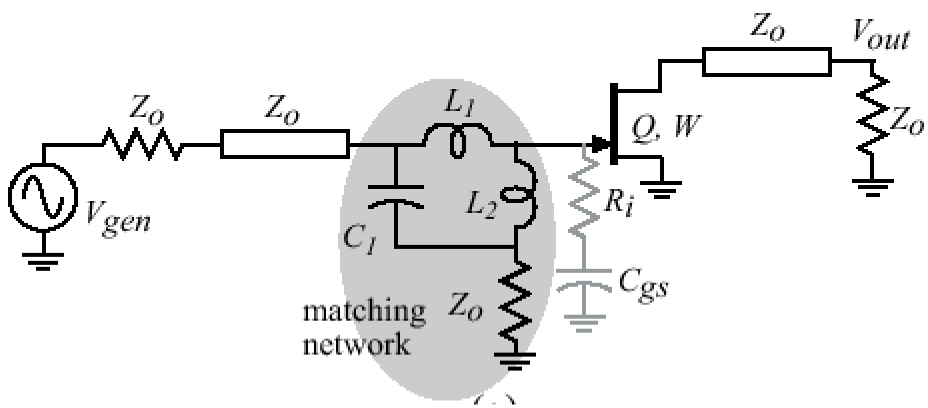


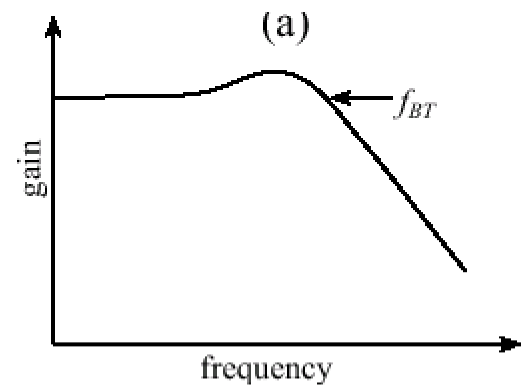
$$C_{in} = C_{be} = C = \tau / Z_o$$

$$L = \tau Z_o = (2C_{in} Z_o) Z_o$$

Insofar as the transistor has an input impedance modeled by Cbe, S11 is exactly zero. S21 nevertheless rolls off at about the same frequency as the T-section case...

Bridged-T-section input tuning....





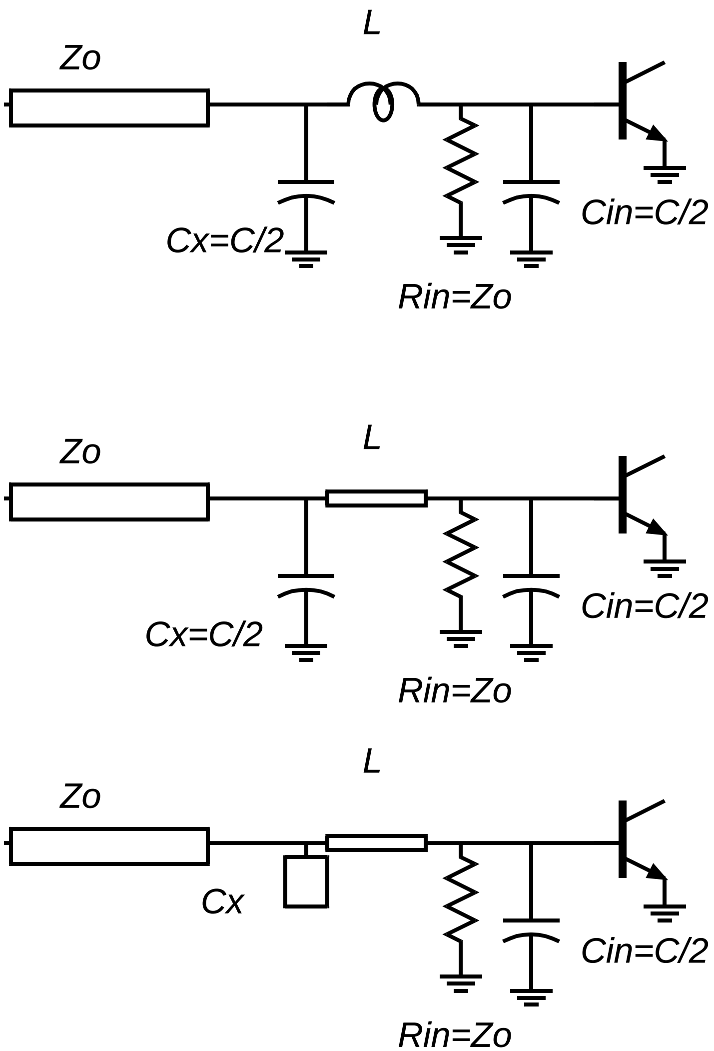
$$\begin{aligned} L_1 &= (Z_o - R_i)Z_oC_{gs}/2 \,, \\ L_2 &= (Z_o + R_i)Z_oC_{gs}/2 \,, \\ C_1 &= (1 - R_i^2/Z_o^2)C_{gs}/4 \,. \end{aligned} \tag{3.19}$$

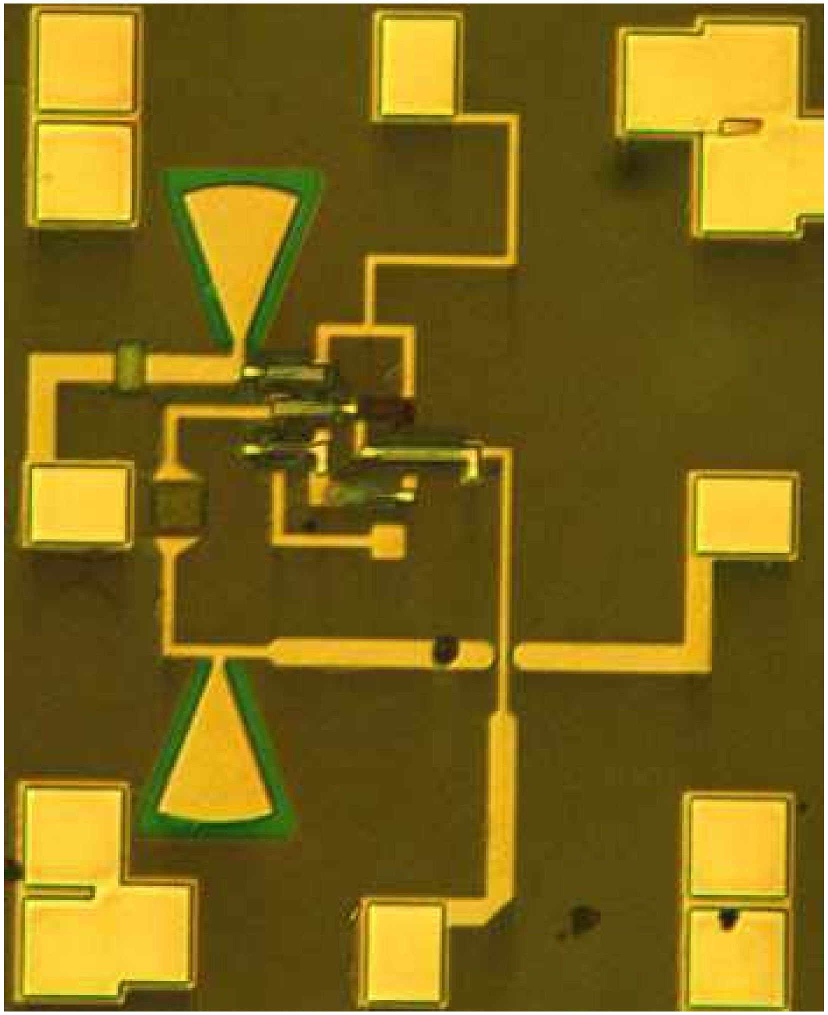
The mid-band voltage gain and transducer power gain are same as the common source case (eq. 3.7), but the amplifier exhibits a gain peaking and the gain falls to the low frequency value at

$$\Delta f_{BT} = \frac{1}{\pi Z_o C_{gs}} \sqrt{\frac{(1 - 3(R_i/Z_o))}{(1 - (R_i/Z_o))\,(1 - (R_i/Z_o)^2)}} \simeq \frac{1}{\pi Z_o C_{gs}} \,, \tag{3.20}$$

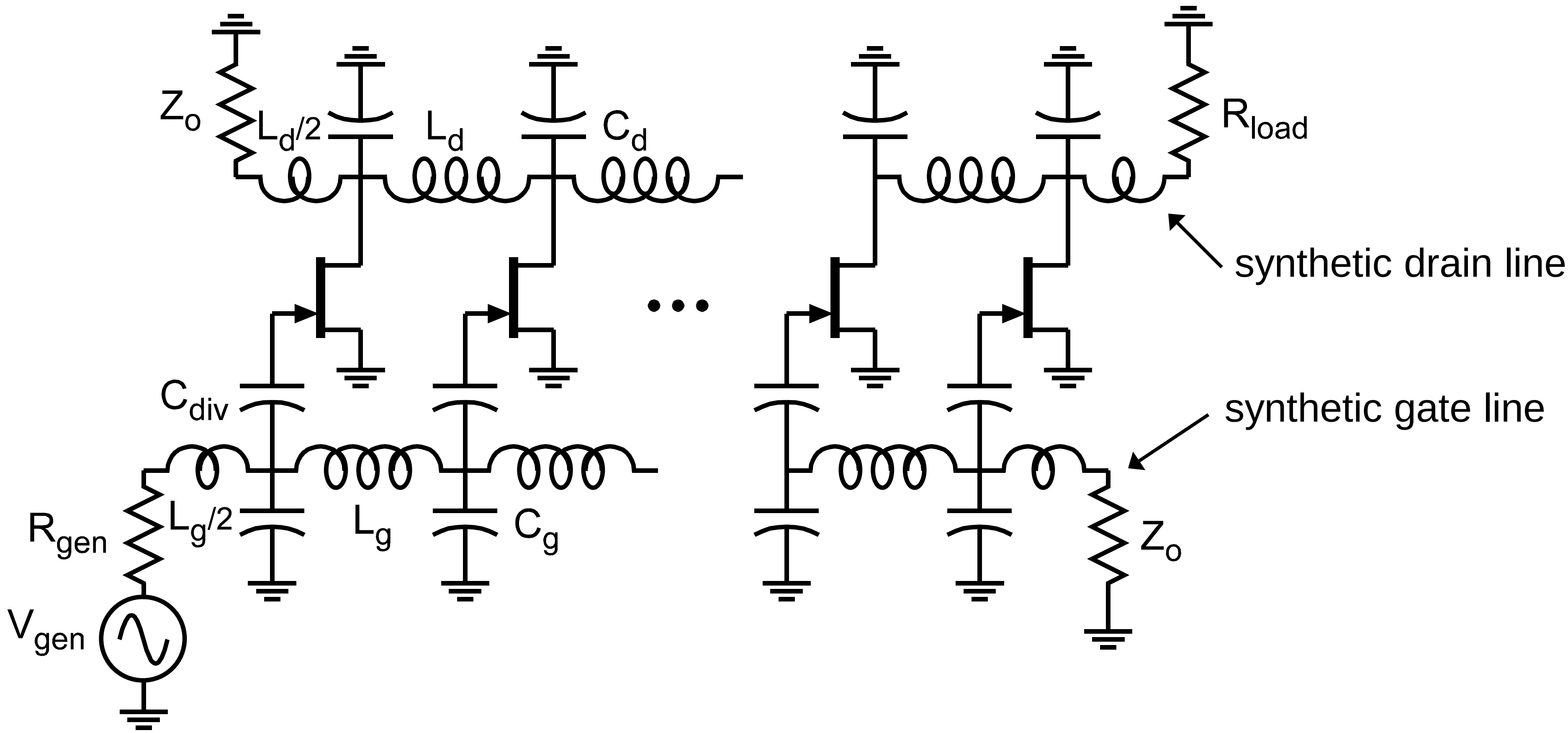
for $R_i \ll Z_o$. Note that the -3 dB bandwidth is somewhat larger than Δf_{BT} (exact expressions for f_{3dB} are intractable).

Tuning Networks use Microstrip elements....





Distributed Amplifiers : Theory



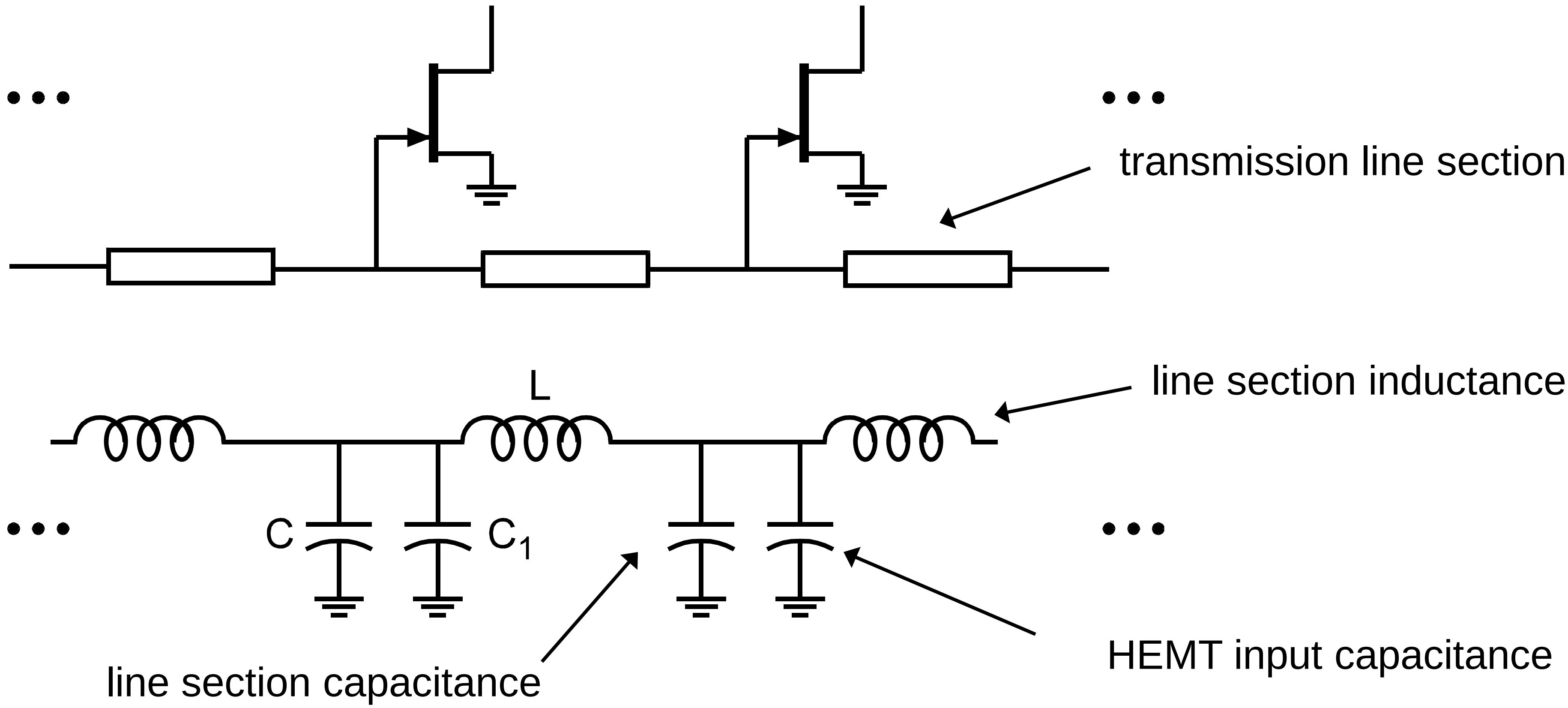
- HEMT input/output capacitances absorbed into artificial input/output lines
- broadband circuit ; gain/bandwidth limited by

  HEMT resistive parasitics (f_{max})

  distributed structure Bragg frequency

  input/output transmission line skin-effect losses

Synthetic Transmission Lines



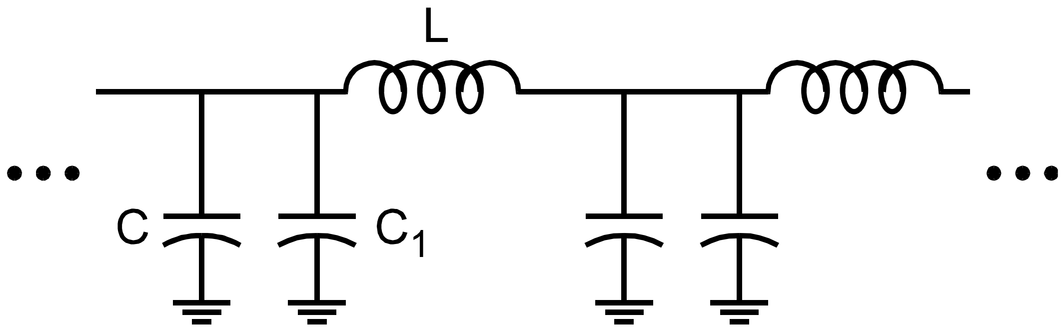
- synthetic input/output line formed by transmission line sections
- transmission line

  characteristic impedance $Z = \sqrt{L/C}$

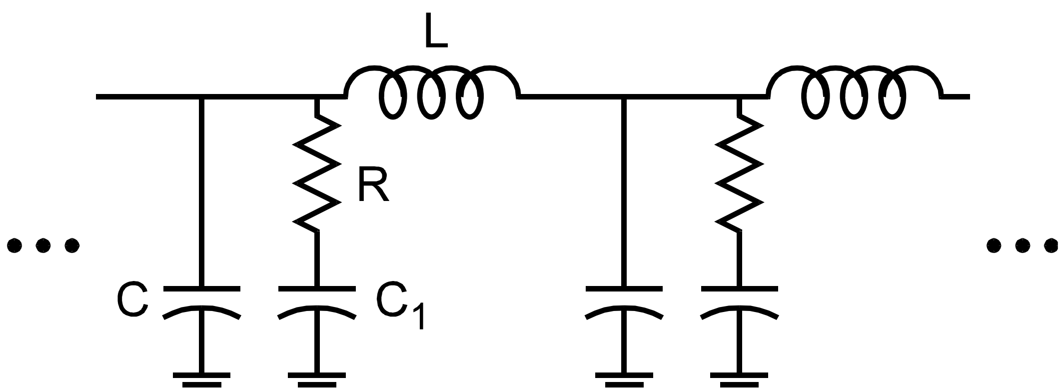
  cutoff (Bragg) frequency $f_B = 1/\pi\sqrt{LC}$

  delay $T = \sqrt{LC}$

Transmission Line Loss

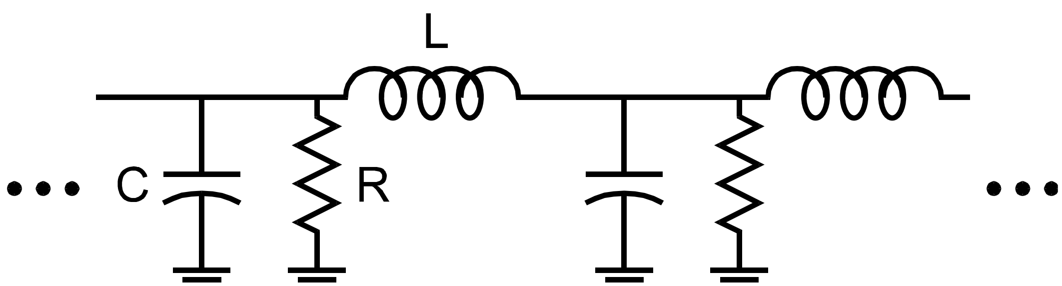


lossless line



frequency-dependent loss

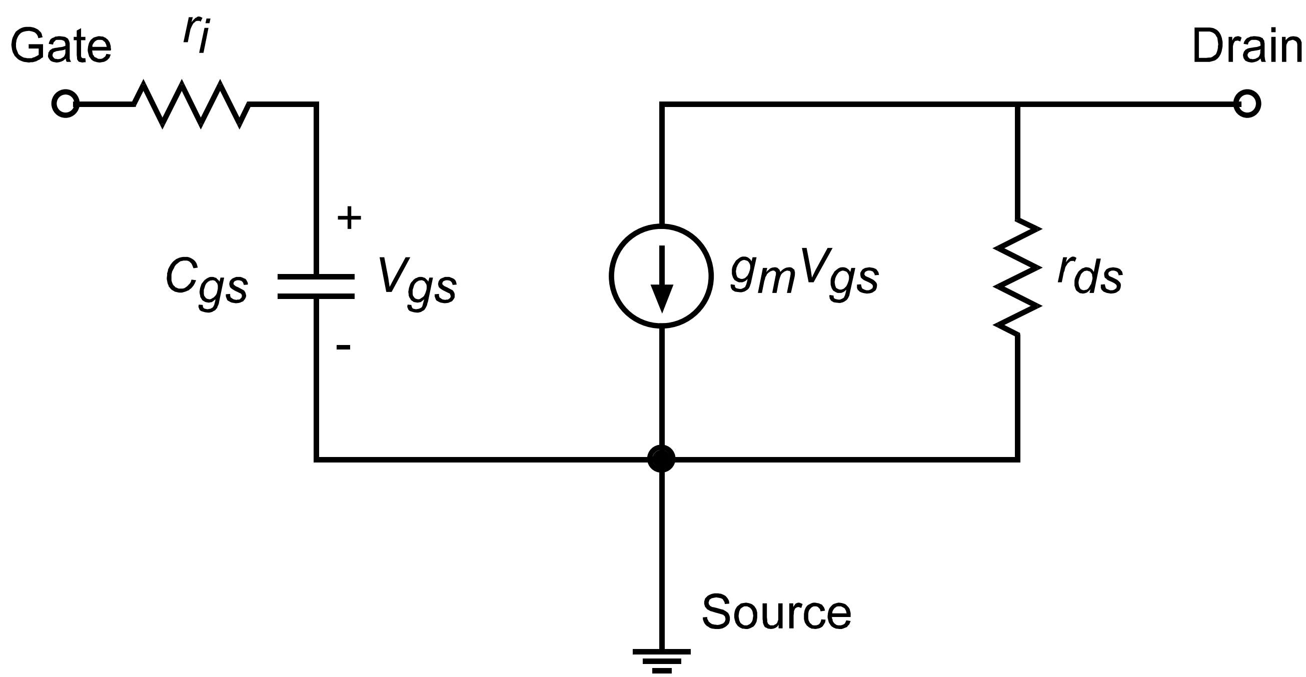
$$\alpha = 4\pi^2 f^2 C_1^2 R Z_0 / 2$$



frequency-independent loss

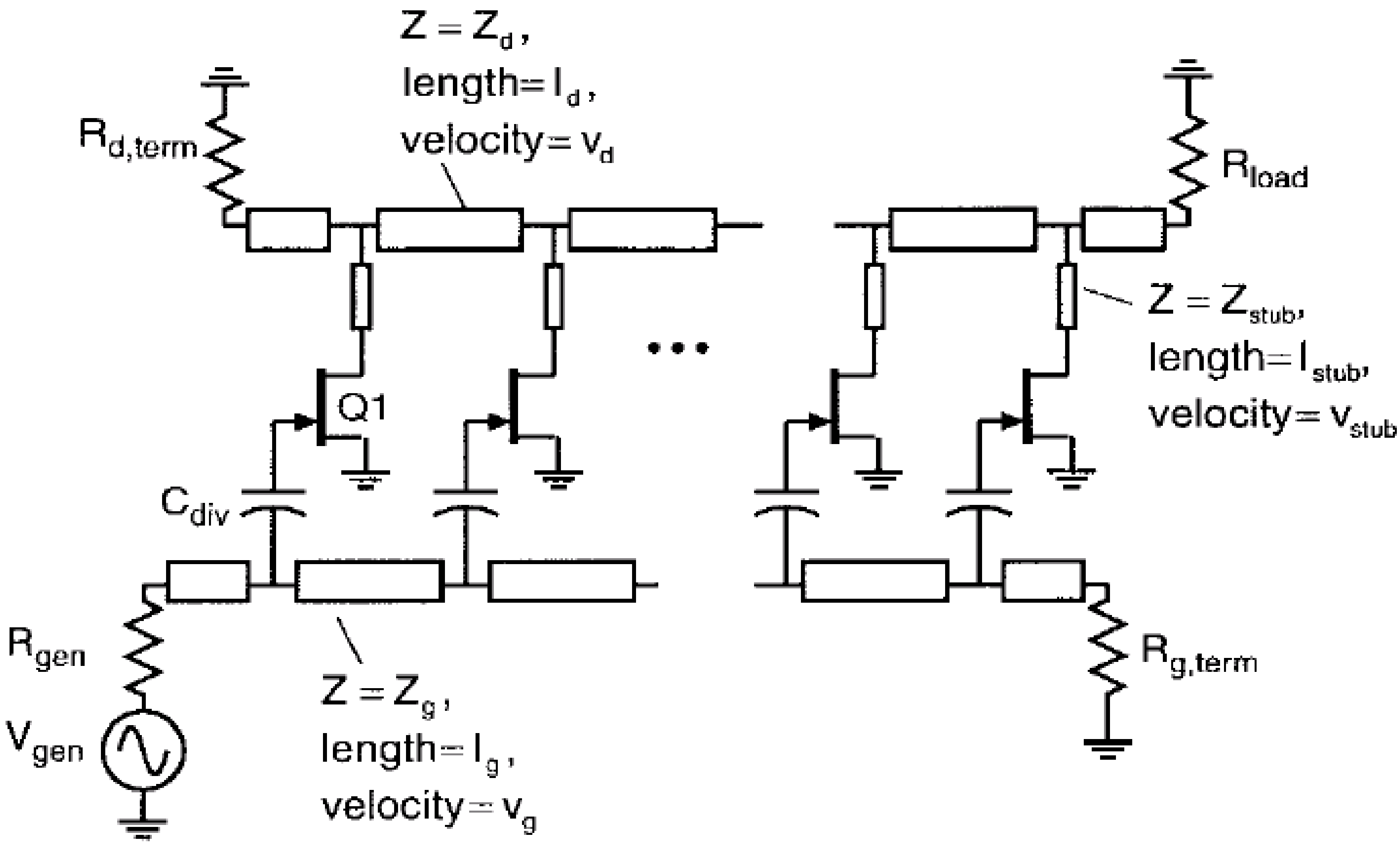
$$\alpha = Z_0 / 2R$$

HEMT Equivalent Circuit



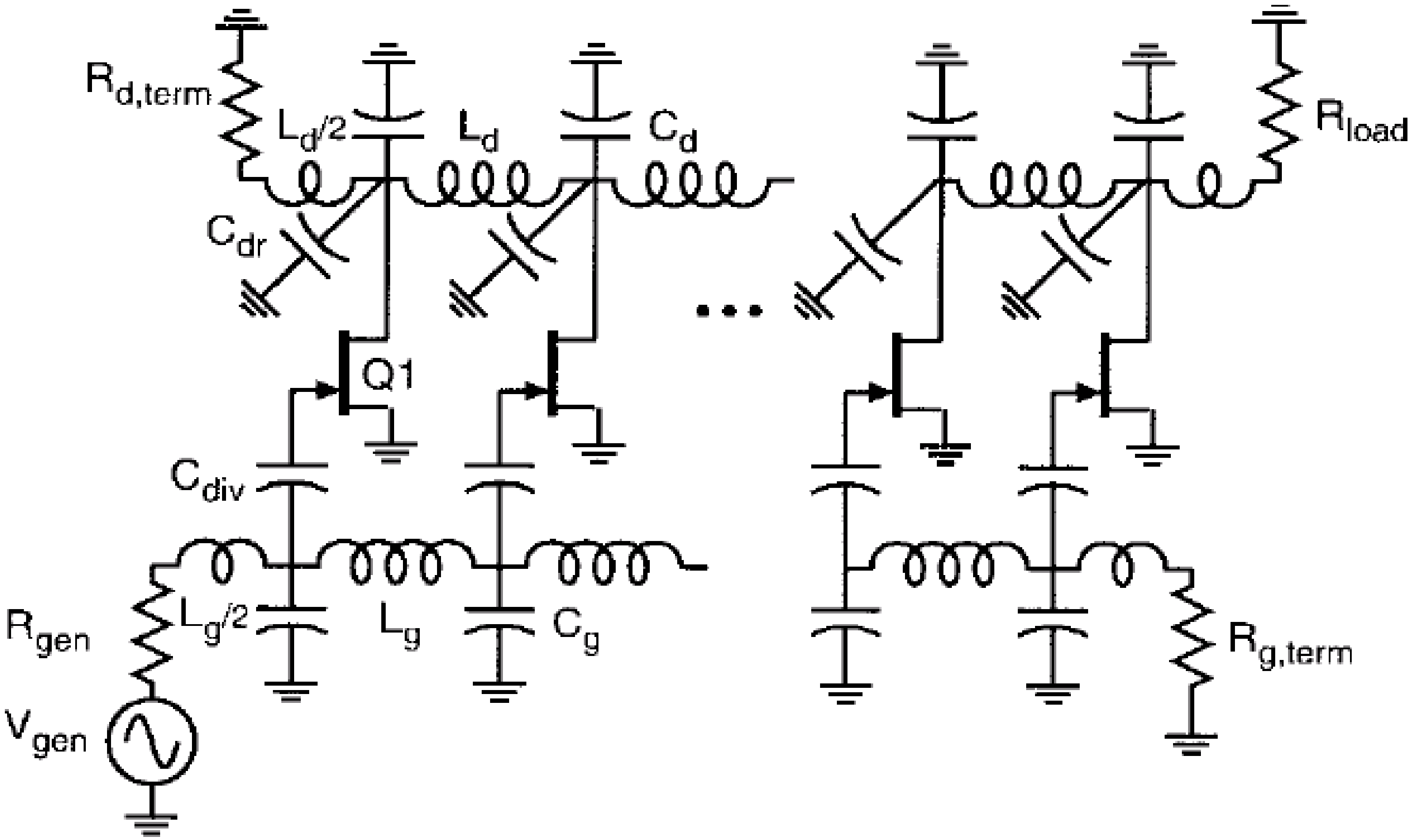
- simplified equivalent circuit
- element values scale with HEMT size
- $f_\tau = g_m / 2\pi C_{gs}$ $\qquad f_{max} = f_\tau \sqrt{r_{ds} / 4r_i}$

CS HEMT TWA

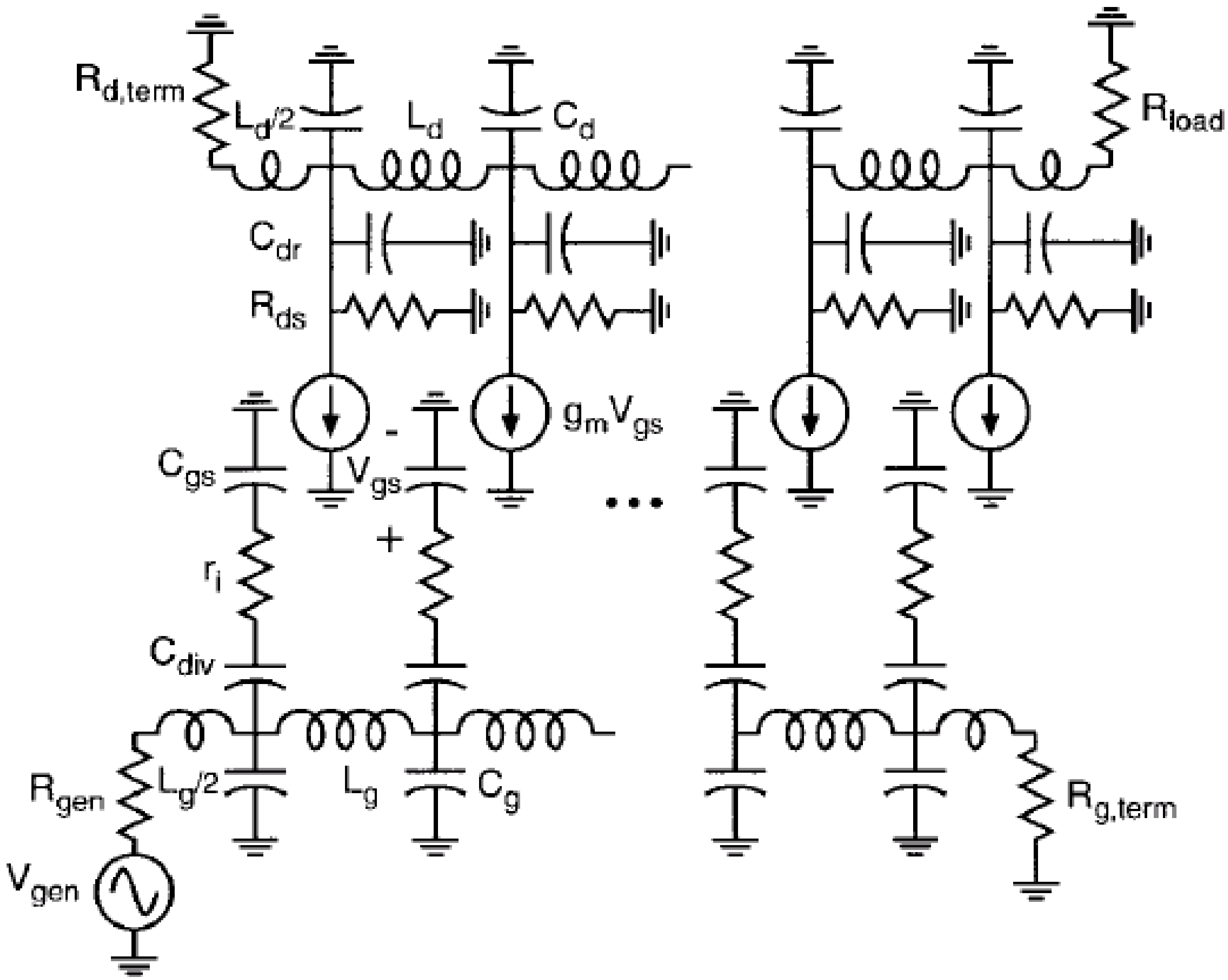


Z = Z_d ,
length= l_d ,
velocity= v_d
R_d,term
R_load
Z = Z_stub,
length= l_stub,
velocity= v_stub
Q1
C_div
R_gen
V_gen
Z = Z_g ,
length= l_g ,
velocity= v_g
R_g,term

CS HEMT TWA--LC equivalent



CS HEMT TWA--SS model



TWA...limits to gain and bandwidth...

Gate (input)line loss :

$\mathrm{N}\alpha_{\mathrm{g}} < 1/2$ ...sets a high frequency limit because $\alpha_g = \omega^2 C^2 R_i$...

Drain (output) Line Loss

$\mathrm{N}\alpha_d < 1/2$...sets a gain limit because $\alpha_{\mathrm{d}} = Z_o / 2R_{ds}$

Input - Output Delay Mismatch

$N(T_{gate} - T_{drain}) < 90^0$...just requires we get design right...

Low - Frequency Gain

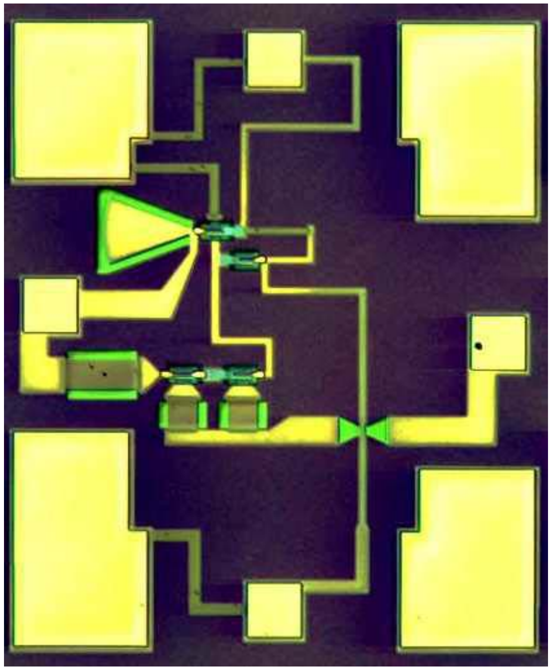
$S_{21} = Ng_m(Z_o/2)$(capacitive degeneration ratio)

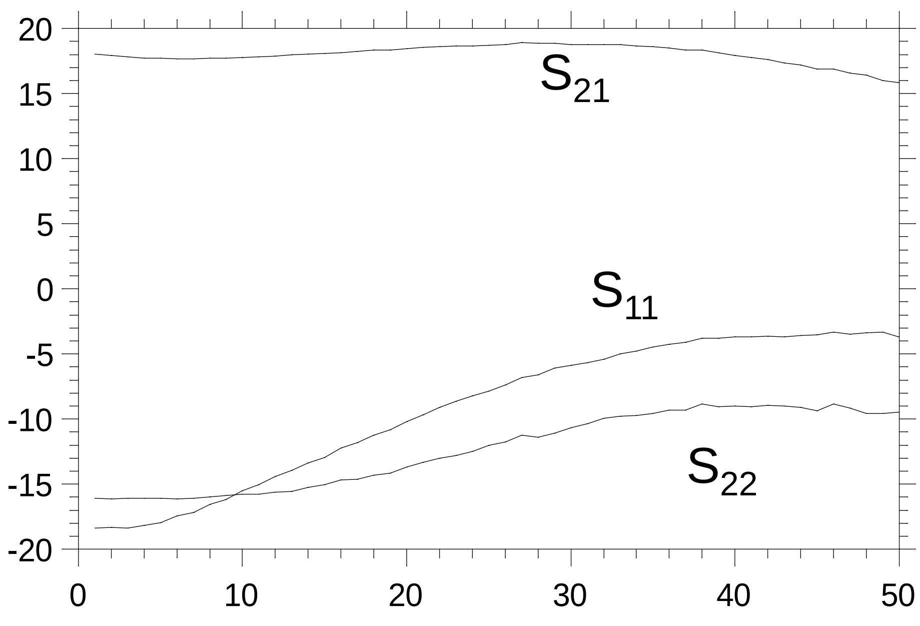
....together with above limits, sets feasible gain - bandwidth

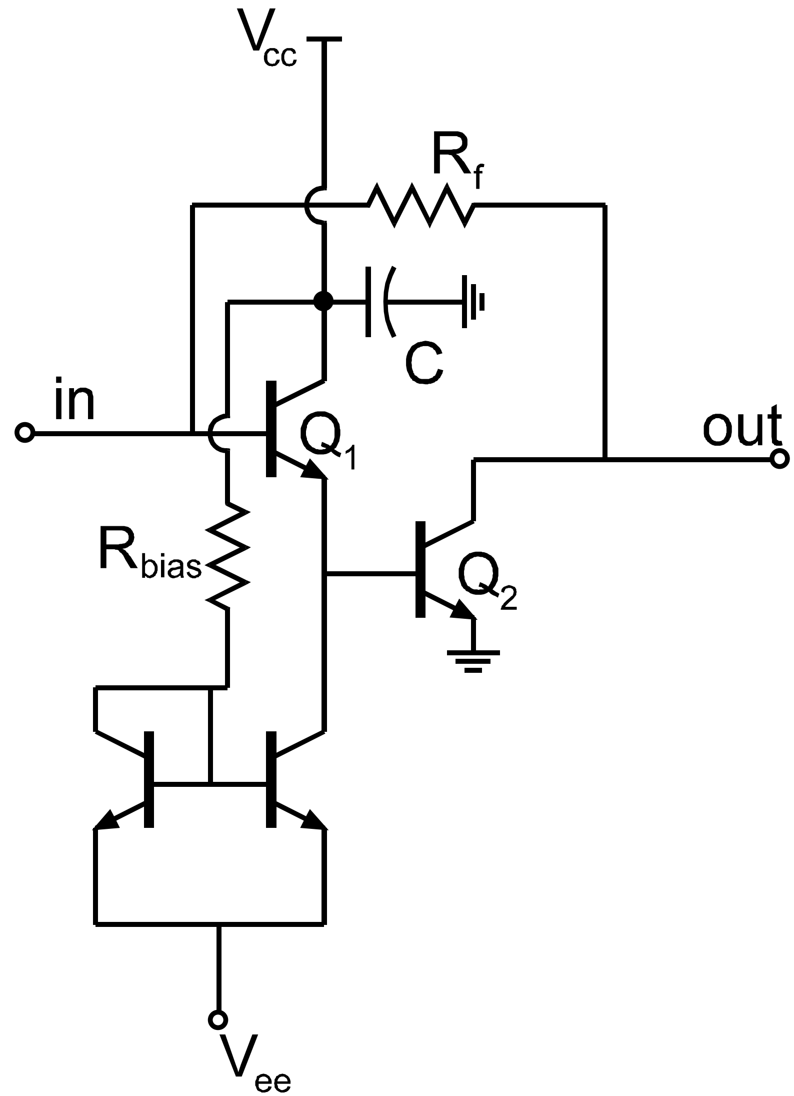
Bragg Frequency

$f < 1/\pi\sqrt{LC}$...can be made very high with lots of very small transistors....no problem...

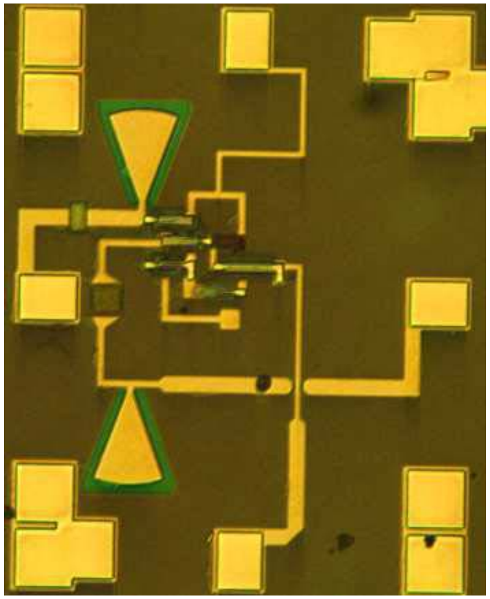
Undamped Darlington FBA

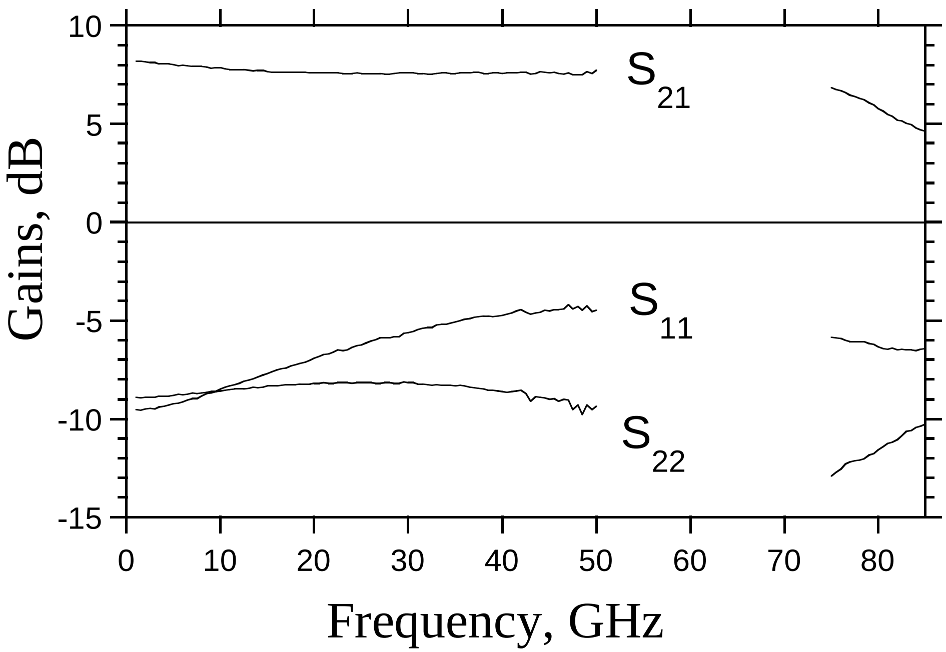


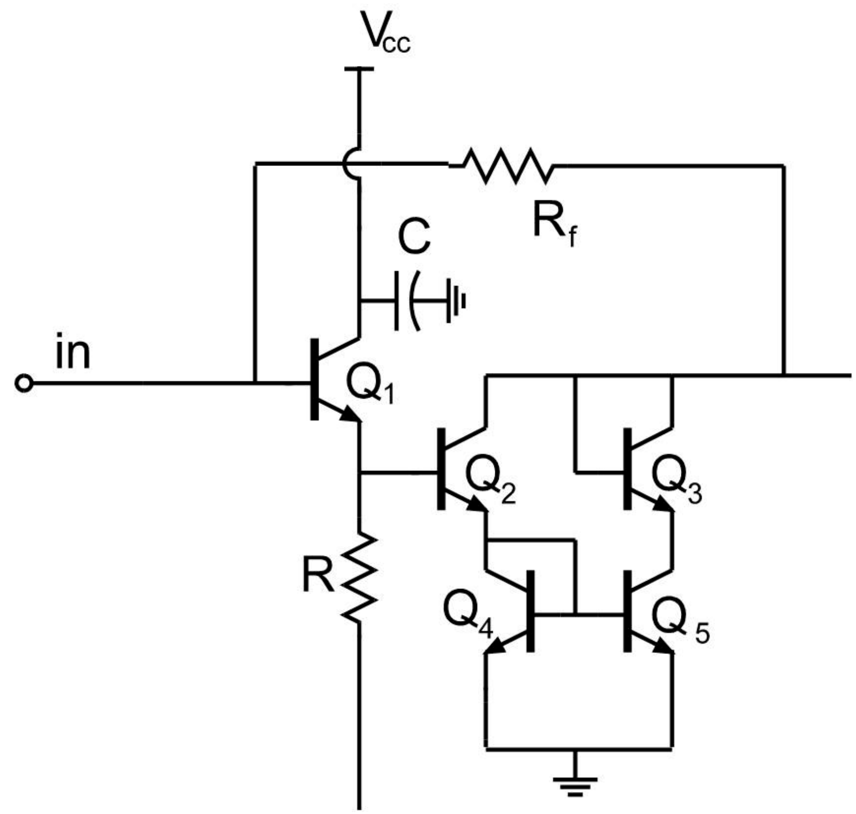




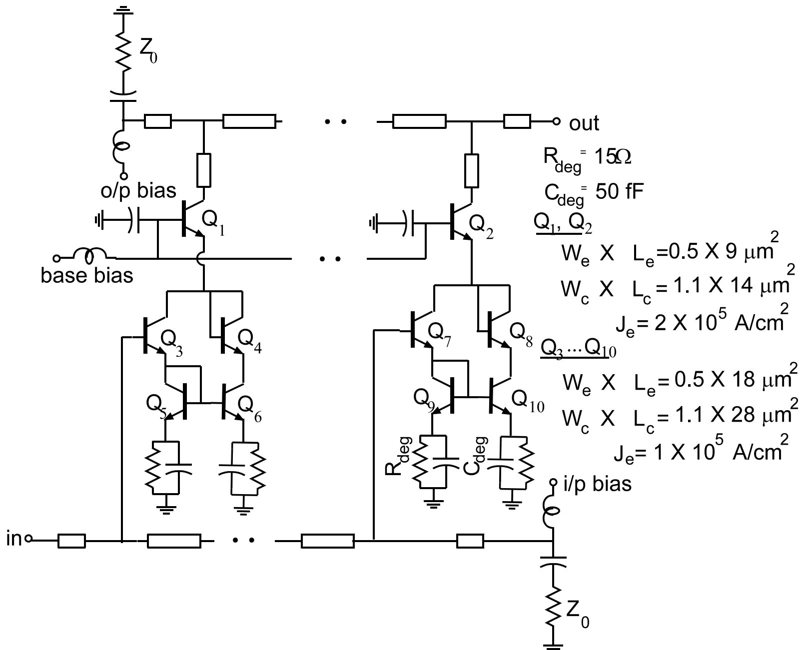
Feedback Amplifier with Follower Driving ft-doubler

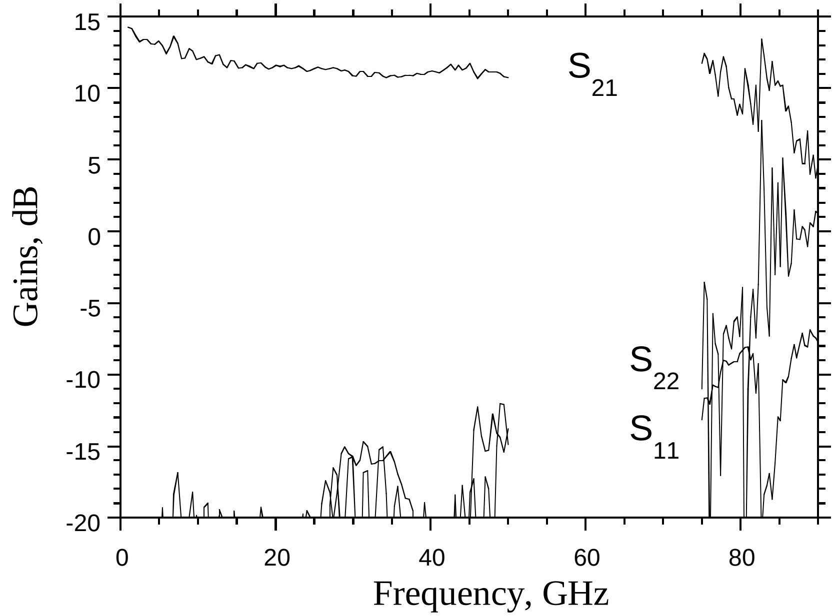


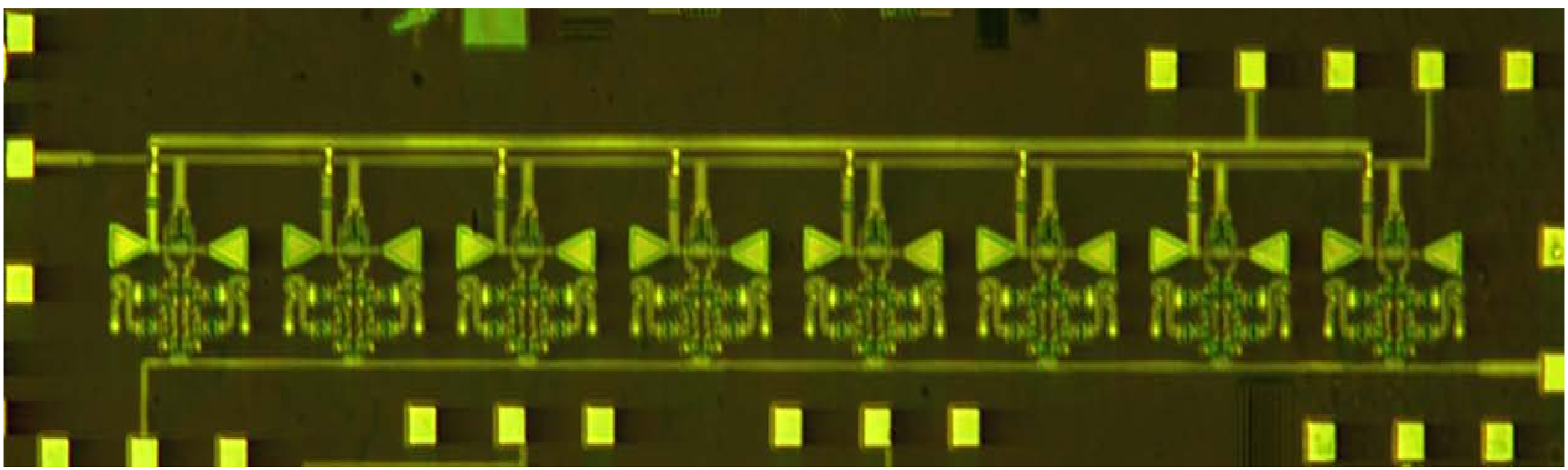




80 GHz HBT Distributed Amplifier







5 dB, 180 GHz Bandwidth HEMT TWA