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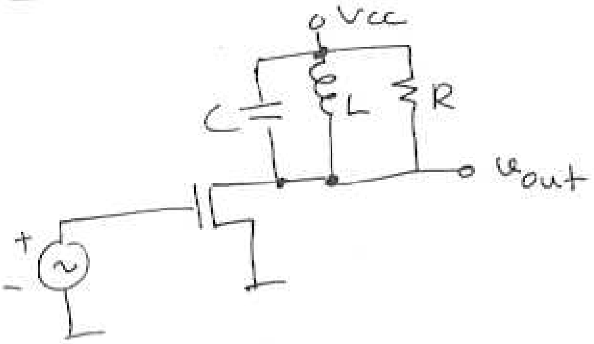
Tuned amplifiers



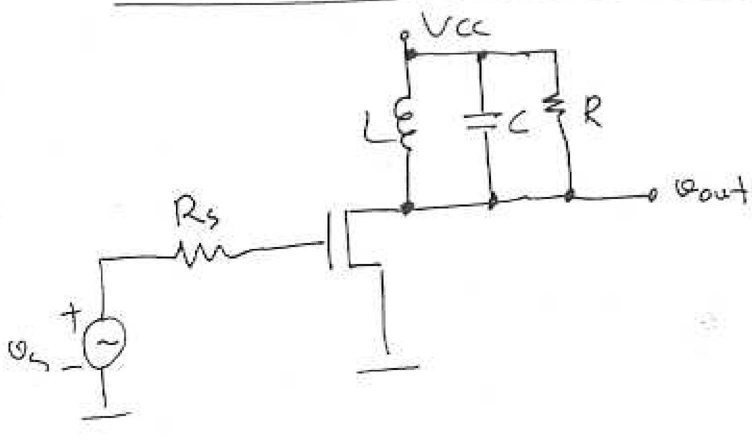
In the case where the amplifier is driven by an ideal voltage source, C_{gd} is absorbed into C, therefore at resonance:

$$\text{Gain} = g_m R$$

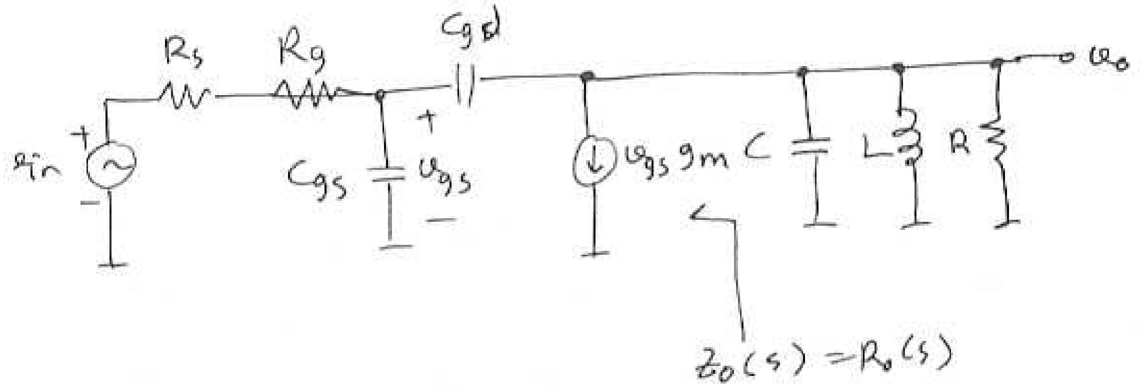
$$\frac{BW}{w_0} = \frac{1}{Q} = \frac{1}{w_0 RC} \Rightarrow BW = \frac{1}{RC}$$

$$\text{Gain} \times BW = \frac{g_m \cancel{R}}{\cancel{R} C} = \frac{g_m}{C}$$

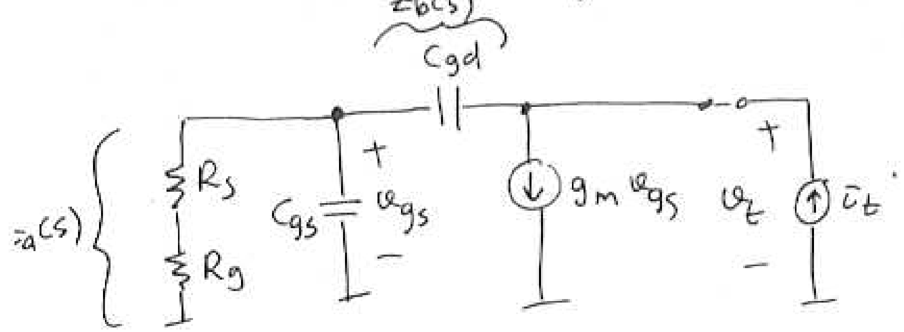
Detailed design of the tuned amplifier

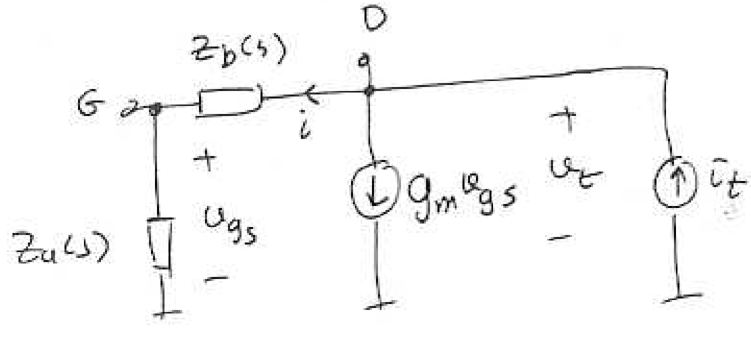


The small signal model is:



In order to analyse the circuit, we must find the loading our transistor creates on the tank circuit. The equivalent circuit to find Ro(s) is:





$$i = i_t - g_m v_{gs}$$

$$i\, Z_a(s) = v_{gs} = Z_a(s)\left[i_t - v_{gs} g_m\right]$$

$$v_{gs}\left[1+Z_a(s)\,g_m\right]=Z_a(s)\,i_t$$

$$v_{gs}=\frac{Z_a(s)}{1+Z_a(s)g_m}\,i_t$$

But

$$i\left[Z_a(s)+Z_b(s)\right]=v_t \Rightarrow$$

$$v_t=\left[Z_a(s)+Z_b(s)\right]\left[i_t-g_m v_{gs}\right]$$

$$=\left[Z_a(s)+Z_b(s)\right]\left[i_t-g_m\frac{Z_a(s)\,i_t}{1+Z_a(s)g_m}\right]$$

$$Z_o(s)=\frac{v_t}{i_t}=\left[Z_a(s)+Z_b(s)\right]\left[\frac{1+\cancel{Z_a(s)g_m}-\cancel{g_m Z_a(s)}}{1+Z_a(s)g_m}\right]$$

$$\boxed{Z_o(s)=\frac{Z_a(s)+Z_b(s)}{1+Z_a(s)g_m}}$$

$$Z_a(s)=\frac{(R_s+R_g)\cdot\frac{1}{j\omega C_{gs}}}{R_s+R_g+\frac{1}{j\omega C_{gs}}}=\frac{R_s+R_g}{1+(R_s+R_g)j\omega C_{gs}}$$

$$Z_b(s)=\frac{1}{j\omega C_{gd}}$$

ignoring C_{gs} (which means that we are well below cut-off) $Z_a(s)$ becomes R_s+R_g

$$\Rightarrow Z_o(s)=\frac{R_s+R_g+\frac{1}{sC_{gd}}}{1+(R_s+R_g)g_m}=\frac{R_s+R_g}{1+(R_s+R_g)g_m}+\frac{1}{sC_{gd}\left[1+(R_s+R_g)g_m\right]}$$

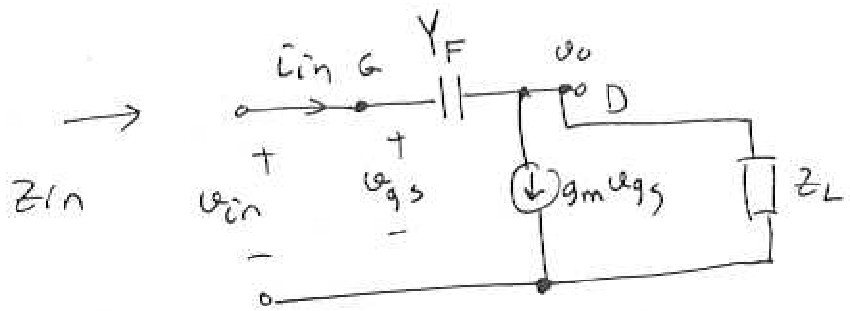
$\Rightarrow$ The reactive part of $Z_o(s)$ is created by $C_{gd}\left[1+(R_s+R_g)g_m\right]$

It means that the miller effect loads the output tank multiplied by

$1 + (R_s + R_g) g_m$

Therefore the resonant frequency of the tank is dependent on C_{gd}. Since it varies depending on different parameters it must be kept low compared to the capacitance of the tank circuit.

A more series effect of the miller capacitance on the tuned amplifier comes from the input impedance of the amplifier. Let us find the input impedance of the amplifier excluding C_{gs} :



$$\left.\begin{aligned} i_{in} &= (v_{gs} - v_o) Y_F \\ v_o &= (i_{in} - g_m v_{gs}) Z_L \end{aligned}\right\} \Rightarrow i_{in} = Y_F \left[v_{gs} - Z_L (i_{in} - g_m v_{gs}) \right]$$

$$\Rightarrow i_{in} = Y_F v_{gs} - Y_F Z_L i_{in} + Y_F Z_L g_m v_{gs}$$

$$i_{in} [1 + Y_F Z_L] = Y_F v_{gs} [Z_L g_m + 1]$$

$$Y_{in} = \frac{i_{in}}{v_{gs}} = \frac{i_{in}}{v_{in}} = \left(Y_F + Y_F Z_L g_m\right)/(1+Y_F Z_L)$$

If $Z_F \gg Z_L$ as in a properly designed amplifier

$$\Rightarrow \quad 1 + Y_F Z_L = 1 + \frac{Z_L}{Z_F} \approx 1 \Rightarrow$$

$$Y_{in} = Y_F + Y_F Z_L g_m$$

At this point we can add C_{gs} to the picture and write $Z_L = R_L + jX_L \Rightarrow$

$$Y_{in} = jw\left[C_{gs} + C_{gd}\right] + jw C_{gd}\left[R_L + jX_L\right] g_m$$
$$= jw\left[C_{gs} + C_{gd}\left[1 + g_m R_L\right]\right] - wC_{gd}(g_m X_L)$$
$$= -wC_{gd}\, g_m X_L + jw\left[C_{gs} + C_{gd}(1 + R_L g_m)\right]$$

Notice here that: 1-) if the load impedance is inductive (i.e. $X_L > 0$), than the input impedance has an negative real part

2-) C_{gd} is reflected to input multiplied by $(1 + g_m R_L)$

Corollories:

1-) The resistive part of the source admittance must be large enough to prevent oscillation at the frequencies where X_L is positive (inductive). Note that Z_L is inductive if $w < \frac{1}{(LC)^{1/2}} = w_0$

2.

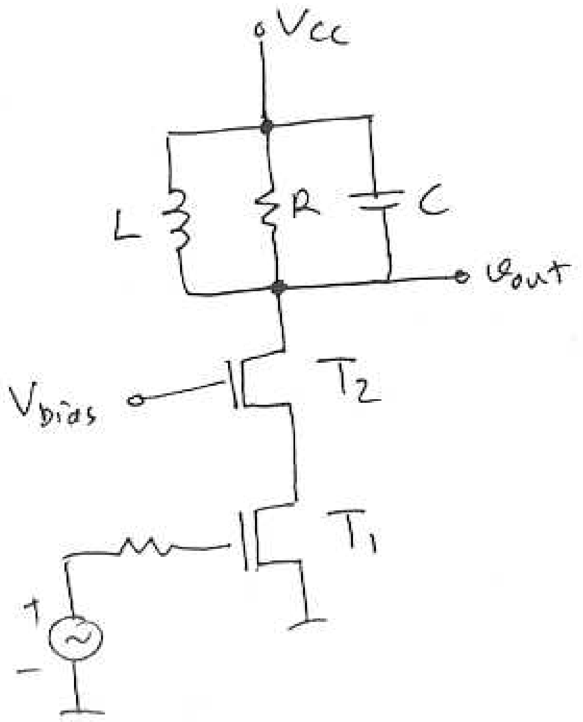
$$Z_o(j\omega) = \frac{R_s + R_g}{1 + (R_s + R_g) g_m} + \frac{1}{j\omega C_{gd}[1 + (R_s + R_g) g_m]}$$

$$Y_{in}(j\omega) = -\omega C_{gd} g_m X_L + j\omega [C_{gs} + C_{gd}(1 + R_L g_m)]$$

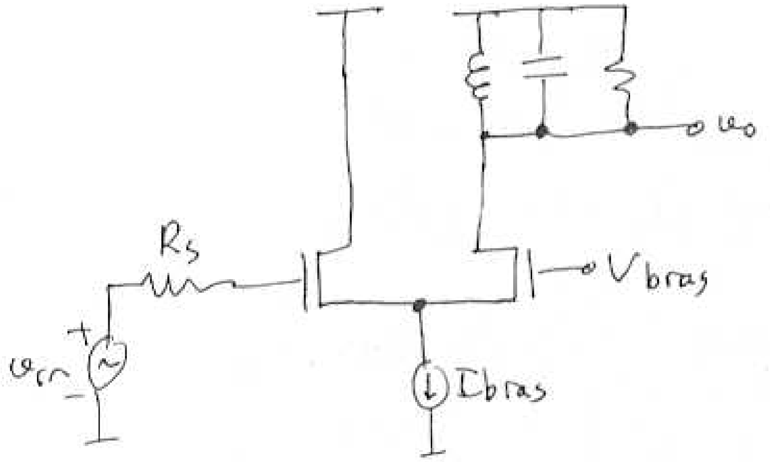
Again notice here that at both loadings C_{gd} is multiplied by a factor related to the gain of the transistor, i.e. $(1 + R_L g_m)$ or $[1 + (R_s + R_g) g_m]$

Unilateralization and Neutralization

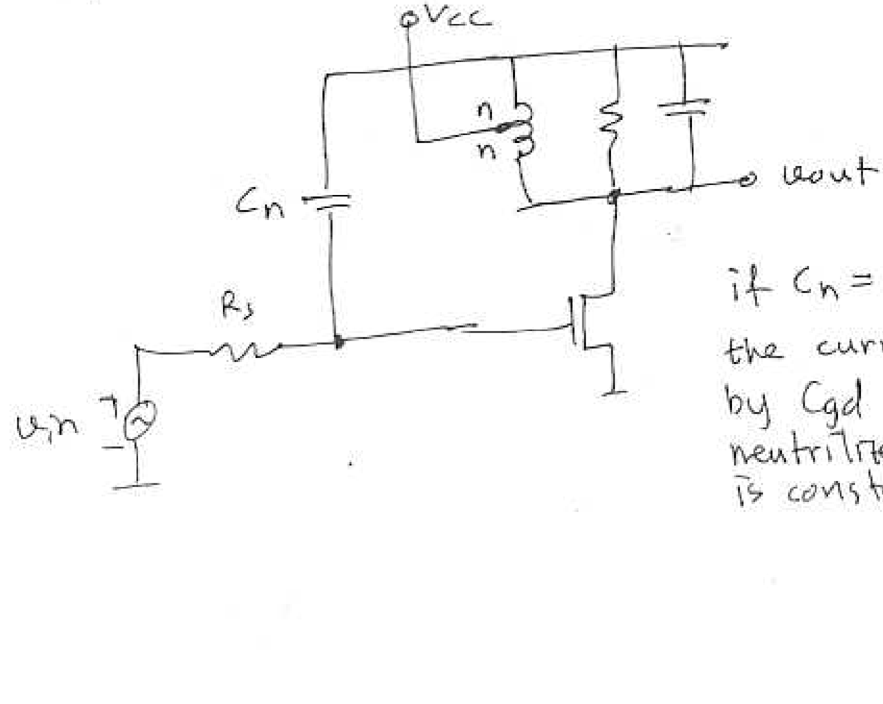
The effect of the miller capacitance C_{gd} can again be mitigated by cascode configuration which effectively ground C_{gd2} & C_{gd1}.



Or by the use of a source follover input amplifrer folloved by a common-gak confrguratron which also effectnely ground both C_{gd1} and C_{gd2}.



We can call these two amplifiers as unilateral amplifrers since the signal flow in one directron from one amplifrer to the other. One other method which cancels the effect of the miller capacitance is neutralizatron:



if $C_n = C_{gd}$
the current coupled by C_{gd} is completely neutrilized if C_{gd} is constant.

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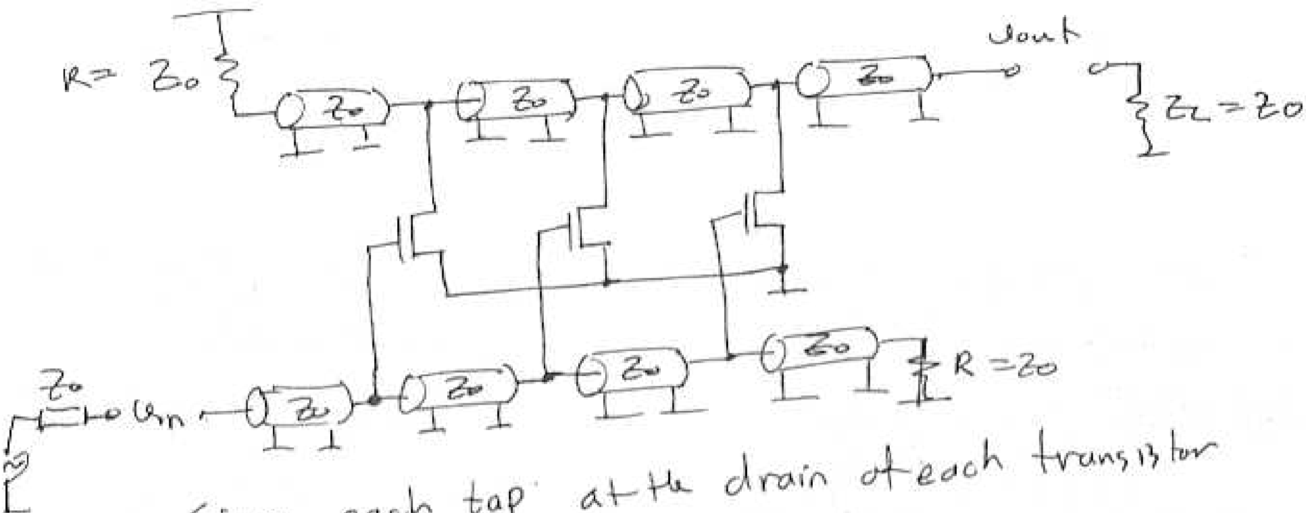
The distributed amplifier



Since each tap at the drain of each transistor shows an impedance of $Z_0/2$ and since delay from input to the output through each transistor is the same (add the output voltages created by each transistor since they are in phase)

$$v_o = n\, g_m \frac{Z_0}{2} v_{in} \Rightarrow$$

$$gain = \frac{v_o}{v_{in}} = n\, g_m \frac{Z_0}{2}$$ so gain increase with n, which means that even individual gains less than one can be combined to obtain gains substantially greater than one with suitable n. Trading delay with bandwidth.

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8-A

The condition for oscillation is:

$$\cancel{G}_S + (-wC_{gd}g_mX_L) < 0$$

$$\cancel{G}_S < wC_{gd}g_mX_L$$

$$\frac{1}{wC_{gd}g_mX_L} < \frac{1}{\cancel{G}_S} = R_S$$

$$R > \frac{1}{wC_{gd}g_mX_L}$$

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8-B

$$Y_{in} = \frac{Y_F + Y_F Z_L g_m}{1 + Y_F Z_L} = \frac{1 + Z_L g_m}{\frac{1}{Y_F} + Z_L}$$

$$= \frac{1 + Z_L g_m}{Z_F + Z_L} = \frac{1 + (R + jX) g_m}{\frac{1}{j\omega C_{gd}} + R + jX}$$

$$= \frac{(1 + R g_m) + jX g_m}{R + j\left(X - \frac{1}{\omega C_{gd}}\right)} = \cancel{\left[(1 + R g_m) + jX g_m\right]}$$

$$= \frac{\left[(1 + R g_m) + jX g_m\right]\left[R - j\left(X - \frac{1}{\omega C_{gd}}\right)\right]}{R^2 + \left(X - \frac{1}{\omega C_{gd}}\right)^2}$$

$$Y_{in} = \frac{\left[(1 + R g_m) R + X g_m \left(X - \frac{1}{\omega C_{gd}}\right)\right] + j\left(R X g_m + (1 + R g_m)\left(\frac{1}{\omega C_{gd}} - X\right)\right)}{R^2 + \left(X - \frac{1}{\omega C_{gd}}\right)^2} + \frac{-j}{\omega C_{gs}}$$

$$(1 + R g_m) R - X g_m \left(\frac{1}{\omega C_{gd}} - X\right) \begin{matrix} < 0 \Rightarrow \text{oscillation} \\ > 0 \Rightarrow \text{stability} \end{matrix}$$

$$(1 + R g_m) R \lessgtr X g_m \left(\frac{1}{\omega C_{gd}} - X\right) \Rightarrow \text{stability} \quad //$$

$$(1 + g_m R) R < g_m \frac{X}{\omega C_{gd}} - g_m X^2 \quad \Rightarrow \quad // \quad \Rightarrow \text{stability}$$

$$(1 + g_m R) R > g_m \frac{X}{\omega C_{gd}} - g_m X^2 \quad \Rightarrow \text{stability}$$

$$//$$

$$g_m X^2 - \frac{g_m}{\omega C_{gd}} X + (1 + g_m R) R > 0$$

$$X^2 + X_c X + \underbrace{\frac{1 + g_m R}{g_m} R}_{c =} > 0$$

$a = 1$ $\quad b = X_c$

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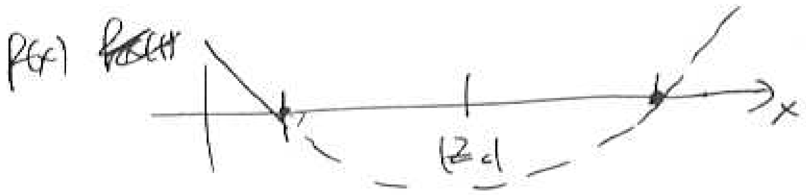
$$x_{1,2} = \frac{-b \pm \sqrt{b^2-4ac}}{2a} = +\left(\frac{-b}{2a}\right) \pm \sqrt{\left(\frac{b}{2a}\right)^2 - \frac{c}{a}} = +\left(\frac{-b}{2}\right) \pm \sqrt{\left(\frac{-b}{2}\right)^2 - c}$$

if $a=1$

$$\frac{1}{2}\left[-b \pm \frac{1}{2}\sqrt{b^2 - \frac{c}{4}}\right]$$

$$= +\left(\frac{-X_c}{2}\right) \pm \sqrt{\left(\frac{-X_c}{2}\right)^2 - \frac{1+g_m R}{g_m} R}$$

$$= \frac{1}{2wC_{gd}} \pm \sqrt{\left(\frac{1}{2wC_{gd}}\right)^2 - \frac{1+g_m R}{g_m} R}$$



$$\Rightarrow \quad x < \frac{1}{2wC_{gd}} - \sqrt{\left(\frac{1}{2wC_{gd}}\right)^2 - \frac{1+g_m R}{g_m} R}$$

and

$$x > \frac{1}{2wC_{gd}} + \sqrt{\left(\frac{1}{2wC_{gd}}\right)^2 - \frac{1+g_m R}{g_m} R}$$