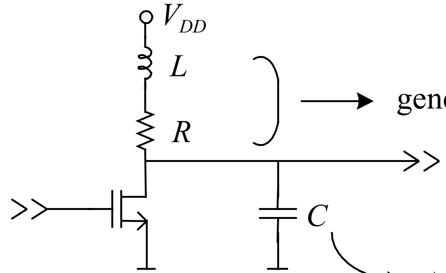
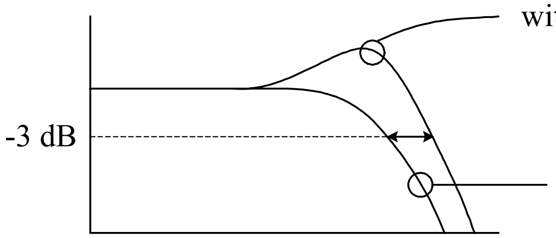
Wide Bandwidth Amplifier Design → Read chapter 8

Shunt-peaked Amps → Goal : extend BW

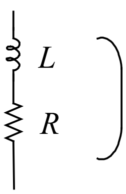


→ generate a zero in transfer function that can do gain peaking

→ typically the input capacitance of the next stage (another MOS transistor)

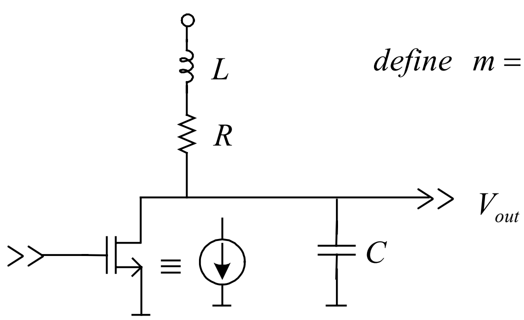


→ bandwidth increases with this method



→ since you use a series combination of inductor and resistor, you can use integrated inductors (Spiral) available on the chip without being worried about their Q

→ assumption → ignore transistor's parasitic (just a current source)



$$define \quad m = \frac{RC}{L/R}$$

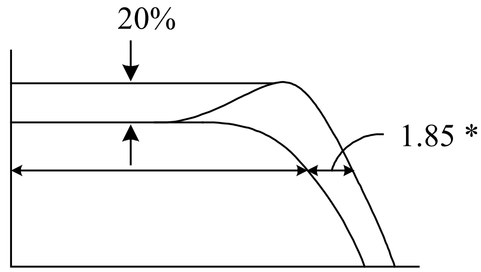
RC → time constant for capacitor

L/R → time constant for inductor

now by choosing m value you can get different results

* $m = \sqrt{2}$ $\longrightarrow$ maximum bandwidth ~ 1.85 $BW_{original}$ ← without shunt peaking

  (problem ~20% peaking in frequency response)



* $m = 2$ $\longrightarrow$ BW ~ 1.8 $BW_{original}$ $\longrightarrow$ still very good

  peaking ~ 3%

* $m = 1 + \sqrt{2}$ $\longrightarrow$ BW ~ 1.72 $BW_{original}$

  no peaking in gain will be seen $\longrightarrow$ maximally flat

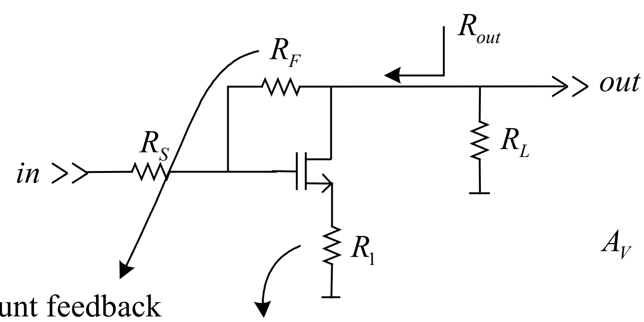
  still gives phase distortion which might be important for some cases (input amplifier of oscilloscope)

* $m = 3.1$ $\longrightarrow$ $BW = 1.6\ BW_{original}$

  no peaking

  no phase distortion (best group delay constant)

Shunt-series Amplifier → goal : convenient input/output matching



shunt feedback
(typically high
resistance value)

series feedback
(typically low resistance value)

$$A_V \approx \frac{-R_L}{R_1} \longrightarrow \text{approximate gain}$$

$$-g_m R_L \approx \frac{-1}{R_1} \cdot R_L$$

shunt feedback

$$\begin{cases} R_{in} = \dfrac{R_F}{1-A_V} \approx \dfrac{R_F}{1+R_L/R_1} \\[2ex] R_{out} = \dfrac{R_F + R_S}{1+R_S/R_1} \approx \dfrac{R_F}{1+R_S/R_1} \end{cases}$$

$if\ R_S \sim R_L = R$

$$\longrightarrow \quad R_{in} \approx R_{out} = \frac{R_F}{1+R/R_1}$$

So this circuit would give you similar input/output matching
if you have similar source and load

↓

very easy to match especially when you have several stages
of the same amplifier

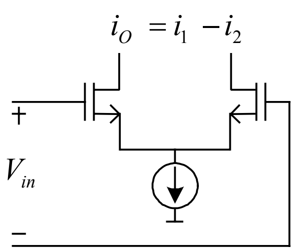
Bandwidth enhancement with f_T doublers

$$2\pi f_T = \frac{g_m}{C_{gd} + C_{gs}} \approx \frac{g_m}{C_{gs}}$$

* differential pair $\longrightarrow$ can be considered f_T doublers

$$\left.\begin{array}{l} g_{m\ total} = g_{m\ transistor} \\ \\ C_{in\ total} = C_{gs}/2 \end{array}\right\} \longrightarrow$$

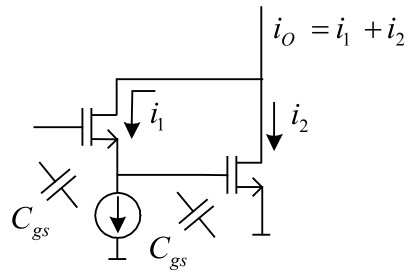
differential pair has a
cut-off frequency of
almost twice a transistor



So you expect that the
bandwidth should be
higher than a transistor (~ x2)

but it is not always possible to have differential signals

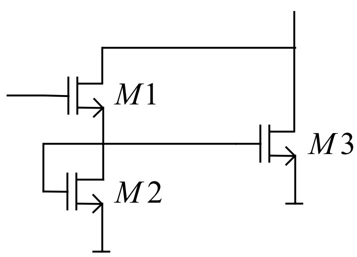
convert differential pair to single ended
(darling ton configuration)



$$C_{in\ total} = C_{gs}/2$$

$$\left\{\begin{array}{l} i_O = i_1 + i_2 \longrightarrow g_{m\ total} = g_{m\ transistor} \\ V_{in} = V_{gs1} + V_{gs2} \end{array}\right.$$

to make $i_1 = i_2$ use simple current source



$\longrightarrow$ increases BW by ~ 1.5x because now

series

$$\underbrace{(C_{gs2} + C_{gs3})}_{\text{parallel}} C_{gs1} \Big/ (C_{gs1} + C_{gs2} + C_{gs3})$$

Bandwidth enhancement through neutralization

the idea is to make $S_{12} = 0$

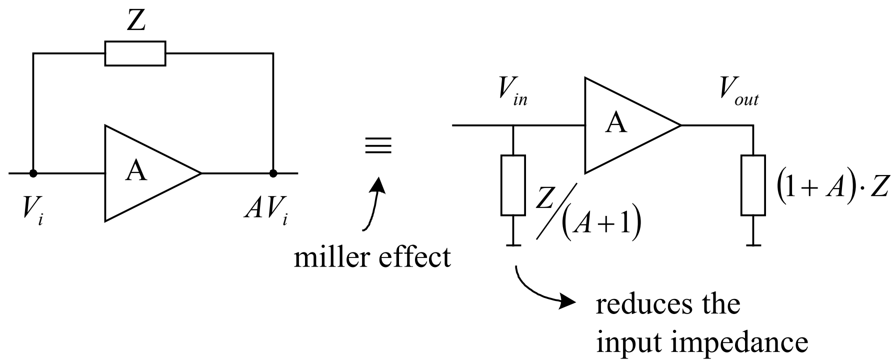
(C_{gd} is mainly responsible for S_{12})

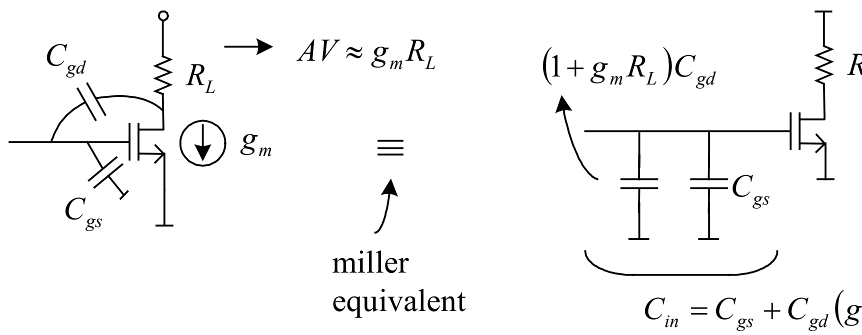
$C_{gd} \lesssim C_{gs}$

typically the case
in submicron CMOS

however effect of C_{gd}
is much more than C_{gs}
why? Because of
miller effect

Miller-Effect

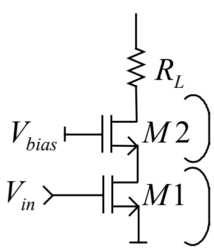




$C_{in} = C_{gs} + C_{gd}(g_m R_L + 1)$

$BW \downarrow$

to reduce the miller-effect you can use cascode



common gate $\rightarrow$ $R_{in2} \sim \dfrac{1}{g_{m2}}$

common source

assuming matched transistor

$$A_r = g_{m1} R_{in2} = \frac{g_{m1}}{g_{m2}} \approx 1$$

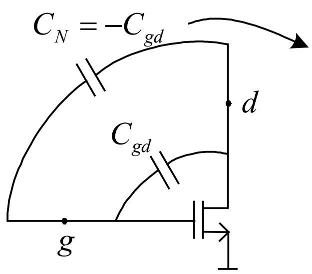
voltage gain of the first stage

but you do not have to have identical transistors so you can use cascode to improve the total Bandwidth

⟶ but there is still some feedback through C_{gd} that

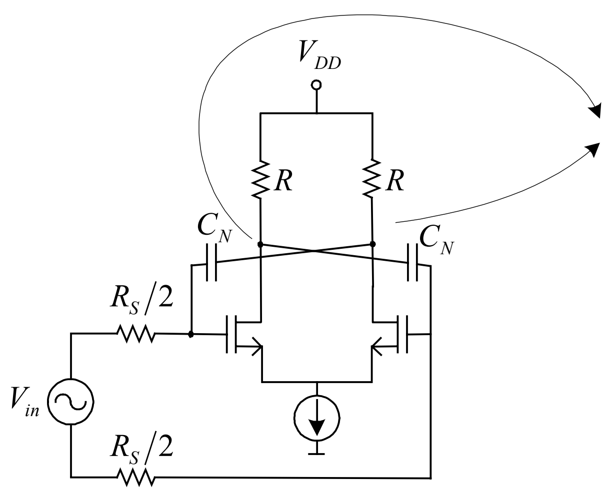
may also cause instability (high S_{12}! often causes stability problem)

to cancel C_{gd} completely you need a negative capacitor in parallel



to make this either use tapped transformer or differential pair how?

Differential pair neutralization

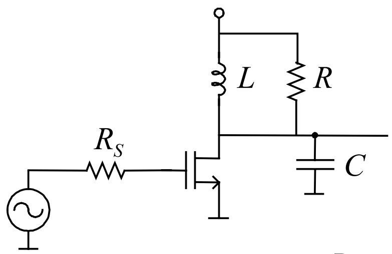


these two nodes are 180 out of phase therefore if $C_N = C_{gd}$ you can neutralize the effect of C_{gd}

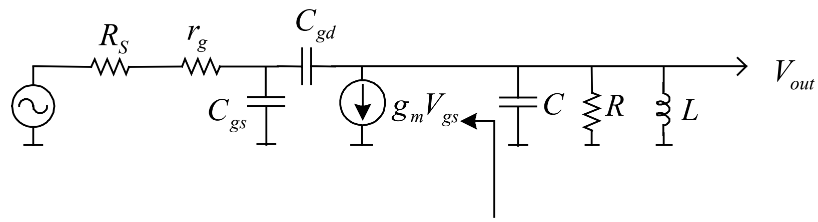
problem is that C_{gd} varies with bias etc...

* you can also use tapped transformer to do neutralization

Tuned Amplifiers



$\equiv$



$Z_{in} = ?$

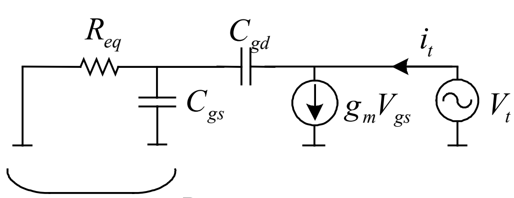
$$\left.\begin{array}{l} -3dB\ \ BW = \dfrac{1}{RC} \\ gain = g_m R \\ \quad @\,resonance \end{array}\right\} \longrightarrow \quad gain \times BW = g_m R \cdot \frac{1}{RC} = \frac{g_m}{C}$$

independent of center frequency

analyze

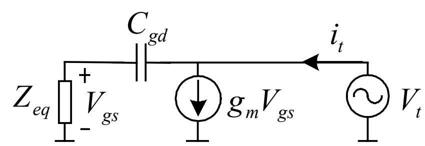
$R_{eq} = R_S + r_g$

to find Zin let's connect a test voltage source



$$Z_{eq} = \frac{R_{eq}}{1 - j\omega C_{gs} R_{eq}}$$

$\equiv$



substitute

$$\text{(I)}\quad \left.\begin{aligned} & i_t = g_m V_{gs} + \left(V_t - V_{gs}\right) j\omega C_{gd} \\ & \left(i_t - g_m V_{gs}\right)\cdot Z_{eq} = V_{gs} \end{aligned}\right\} \longrightarrow \quad V_{gs} = \frac{Z_{eq}\cdot i_t}{1+g_m Z_{eq}}$$

$$i_t = j\omega C_{gd}\cdot V_t + V_{gs}\left(g_m - j\omega C_{gd}\right)$$

$$i_t = j\omega C_{gd}\cdot V_t + \frac{Z_{eq}}{1+g_m Z_{eq}}\left(g_m - j\omega C_{gd}\right)\cdot i_t$$

$$i_t\left(1+\cancel{g_m Z_{eq}} - \cancel{g_m Z_{eq}} + j\omega C_{gd}\cdot Z_{eq}\right) = V_t\; j\omega C_{gd}\left(1+g_m Z_{eq}\right)$$

$$i_t\left(1+j\omega C_{gd} Z_{eq}\right) = V_t\; j\omega C_{gd}\left(1+g_m Z_{eq}\right)_t$$

$$Z_{in} = \frac{V_t}{i_t} = \frac{1+j\omega C_{gd} Z_{eq}}{j\omega C_{gd}\left(1+g_m Z_{eq}\right)} = \frac{\dfrac{1}{j\omega C_{gd} Z_{eq}}+1}{\dfrac{1}{Z_{eq}}+g_m} = \frac{\dfrac{1}{R_{eq}\, j\omega C_{gd}}+1-\dfrac{C_{gs}}{C_{gd}}}{g_m+\dfrac{1}{R_{eq}}-j\omega C_{gs}}$$

$$\frac{1}{Z_{eq}} = \frac{1}{R_{eq}} - j\omega C_{gs}$$

for rather small frequencies : $\quad j\omega C_{gs} << g_m + \dfrac{1}{R_{eq}}$

$$Z_{in} \approx \frac{1}{j\omega C_{gd}\left(1+g_m R_{eq}\right)} + \frac{\left(C_{gd}-C_{gs}\right)/C_{gd}}{g_m+\dfrac{1}{R_{eq}}} \qquad , \quad R_{eq} = r_g + R_S$$

$\longrightarrow$ $C_{in} = C_{gd}\left(1 + g_m\left(r_g + R_S\right)\right)$ $\longleftarrow$ equivalent capacitance can be very high

$$R_{in} = \frac{R_S + r_g}{1 + g_m\left(r_g + R_S\right)} \times \underbrace{\frac{C_{gd} - C_{gs}}{C_{gd}}}_{\text{negative since } C_{gd} < C_{gs}\ (\textit{typically})} < 0$$

$\longleftarrow$ negative resistance

$\longrightarrow$ your circuit(turned Amp) can be easily unstable

(good if you are making an oscillator)

if you did not have C_{gd}, you did not have this oscillation problem