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Analog Circuits Basics

Content

Single Ended Amplifiers

Current Mirrors

Differential Amplifiers

Content

Single Ended Amplifiers
- DC/Low Frequency
- AC/High Frequency

Current Mirrors

Differential Amplifiers

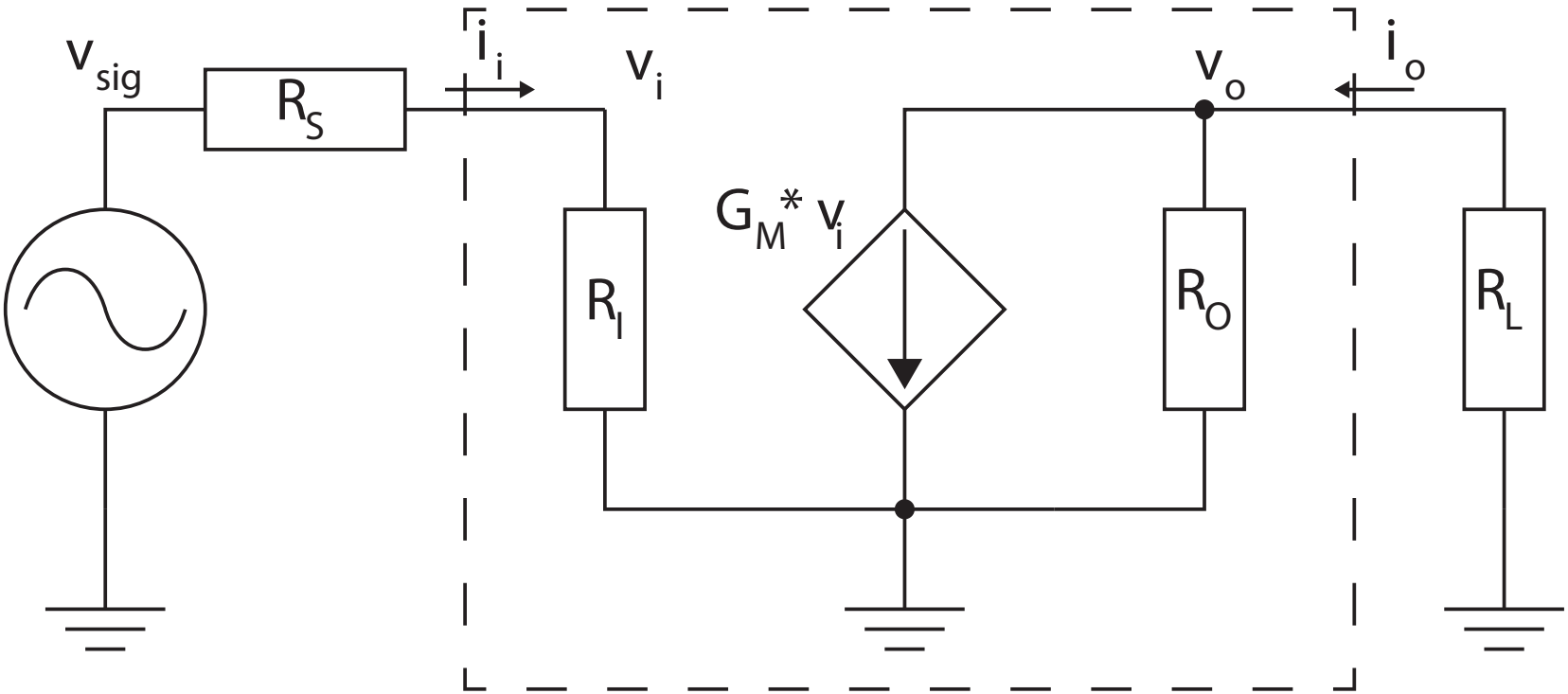
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Single Ended Amplifiers
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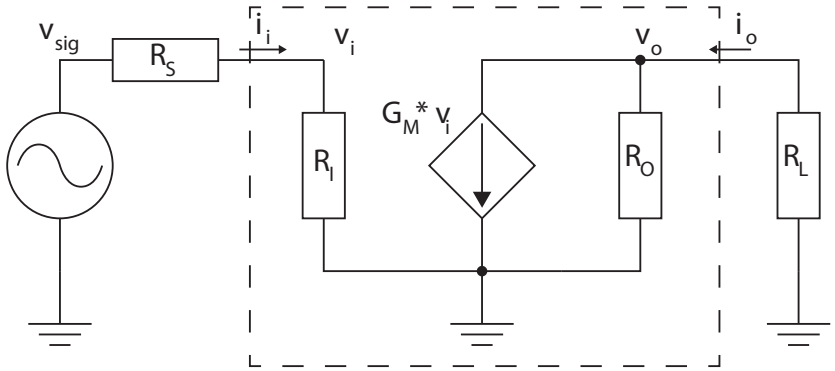
Current Mirrors

Differential Amplifiers
- Differential pair
- Resistor loaded diff pair
- Current mirror loaded diff pair
- Folded Cascode Operational Amplifier

General LF Model



Find the parameters

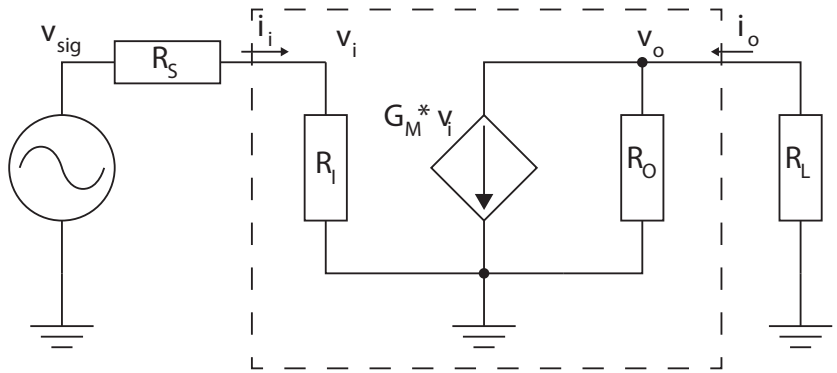


$$R_I := \frac{v_i}{i_i} = \frac{\partial V_I}{\partial I_I}$$

$$G_M := \frac{i_o}{v_i} = \frac{\partial I_O}{\partial V_I} \text{ for } V_O \text{ held constant and } v_o = 0 \text{ !}$$

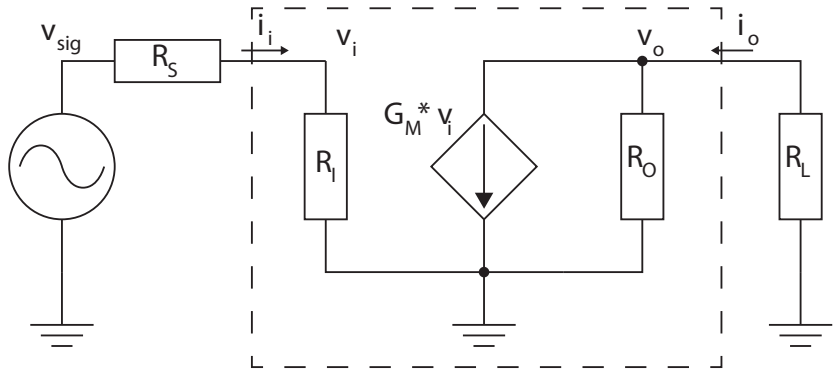
$$R_O := \frac{v_o}{i_o} = \frac{\partial V_O}{\partial I_O} \text{ for } V_{SIG} \text{ held constant and } v_{sig} = 0 \text{ !}$$

Intrinsic Gain



$$A_O = -G_M R_O$$

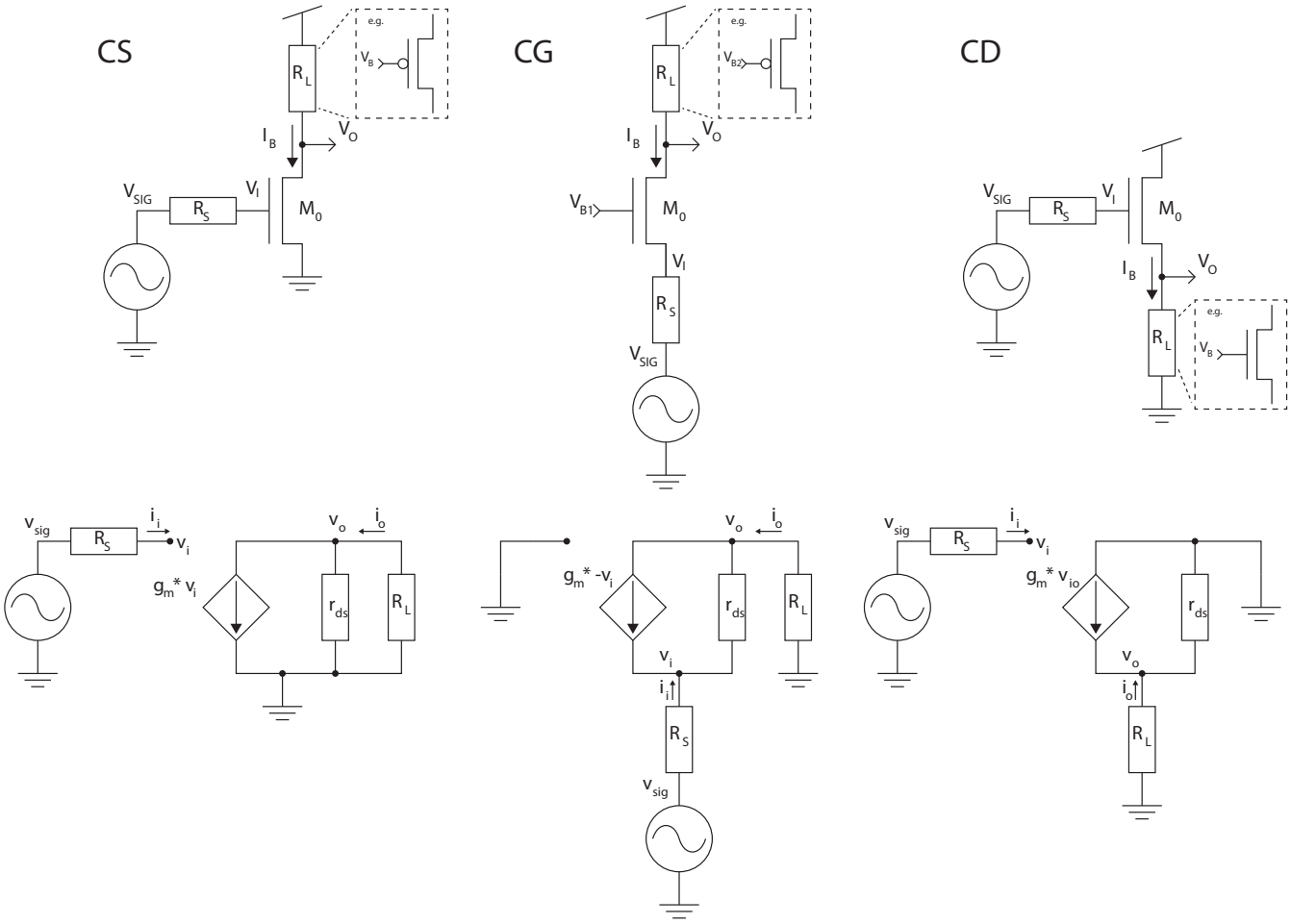
Total Gain



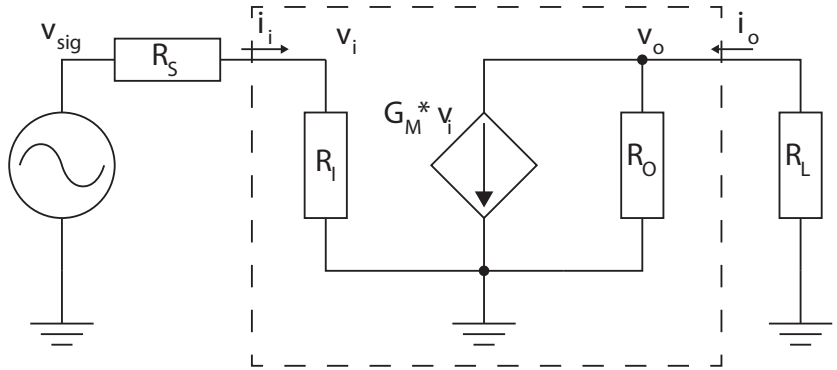
$$\frac{i_o}{v_{sig}} := G'_M = G_M \frac{R_I}{R_S + R_I}$$

$$A' := \frac{v_o}{v_{sig}} = -G'_M(R_O||R_L) = \frac{R_I}{R_S + R_I} A_O \frac{R_L}{R_O + R_L}$$

Single Transistor Gain Stages



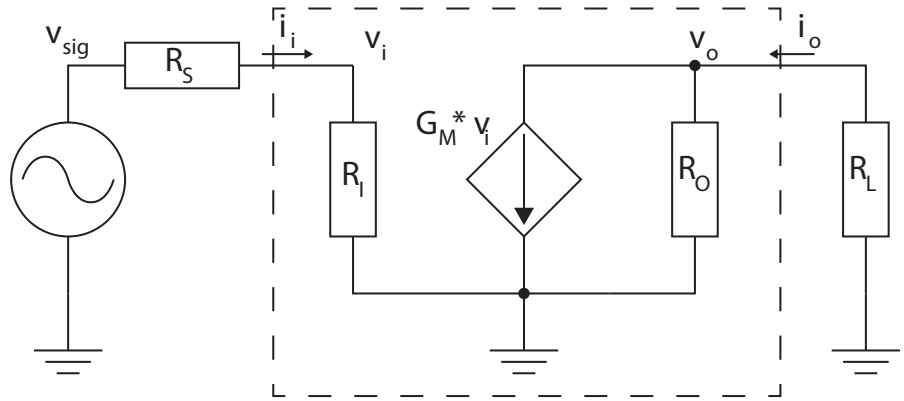
Single Transistor Gain Stages



	R_I	G_M	R_O
CS	∞	g_m	r_{ds}
CG	$\frac{r_{ds}+R_L}{1+g_m r_{ds}}$	$-(\frac{1}{r_{ds}}+g_m)$	$r_{ds}+R_S(g_m r_{ds}+1)$
CD	∞	$-g_m$	$\frac{1}{g_m}\|\|r_{ds}=\frac{r_{ds}}{1+g_m r_{ds}}$

Single Transistor Gain Stages: CS

	R_I	G_M	R_O
CS	∞	g_m	r_{ds}

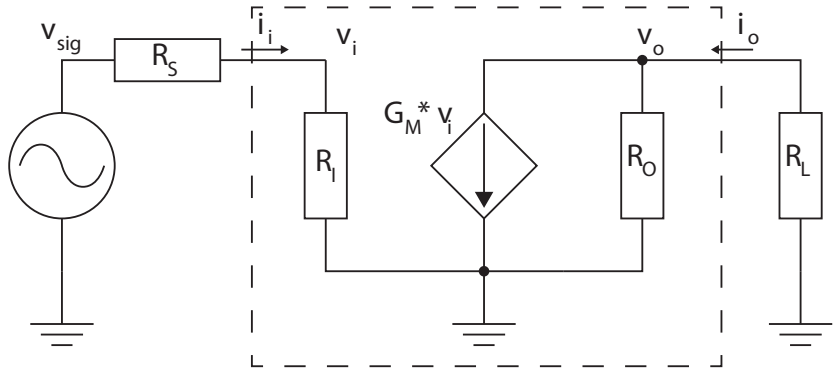


One (not so common) way of looking at it is as a decent voltage controlled current source, since the output resistance is high (makes a good/robust current output).

The more common way of looking at it is as a high gain voltage amplifier, especially in LF/DC even with a "bad" voltage source (i.e. a good voltage receiver with $R_I = \infty$). However, it cannot drive low R_L loads (or significant C_L at HF), i.e. low R_L severely diminish the voltage gain.

These two views are quite equivalent: a good controlled current source with high output resistance needs a huge voltage swing at the output for small changes in output current.

Single Transistor Gain Stages: CG



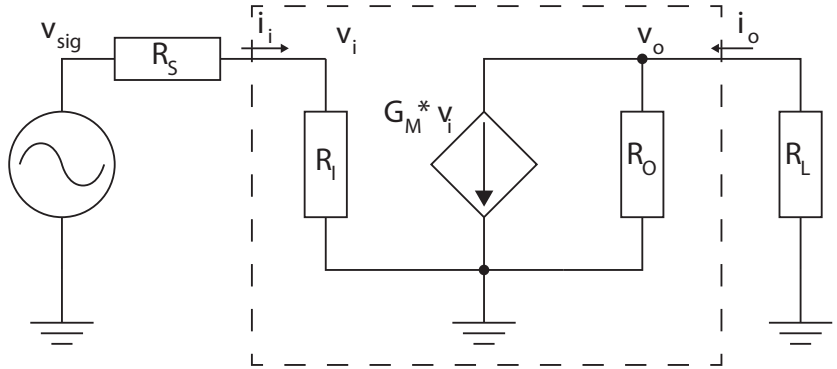
	R_I	G_M	R_O
CG	$\frac{r_{ds}+R_L}{1+g_m r_{ds}}$	$-(\frac{1}{r_{ds}}+g_m)$	$r_{ds}+R_S(g_m r_{ds}+1)$

Good as current conveyor: $i_o = -i_i$ (evident in the single transistor stages overview (fig. 9.2 in the compendium) and using a Norton signal source), while $R_I << R_L$ and $R_O >> R_S$, i.e. it turns the signal source into a better current source.

$$A_I' = \frac{-i_o}{i_{sig}} = \frac{i_i}{i_{sig}} = \frac{\frac{1}{R_I}}{\frac{1}{R_I}+\frac{1}{R_S}} \overset{R_I << R_S}{\approx} 1$$

With a very good voltage source (i.e. $R_S << R_I$) it *could* also serve as a voltage amplifier ...

Single Transistor Gain Stages: CD/SF

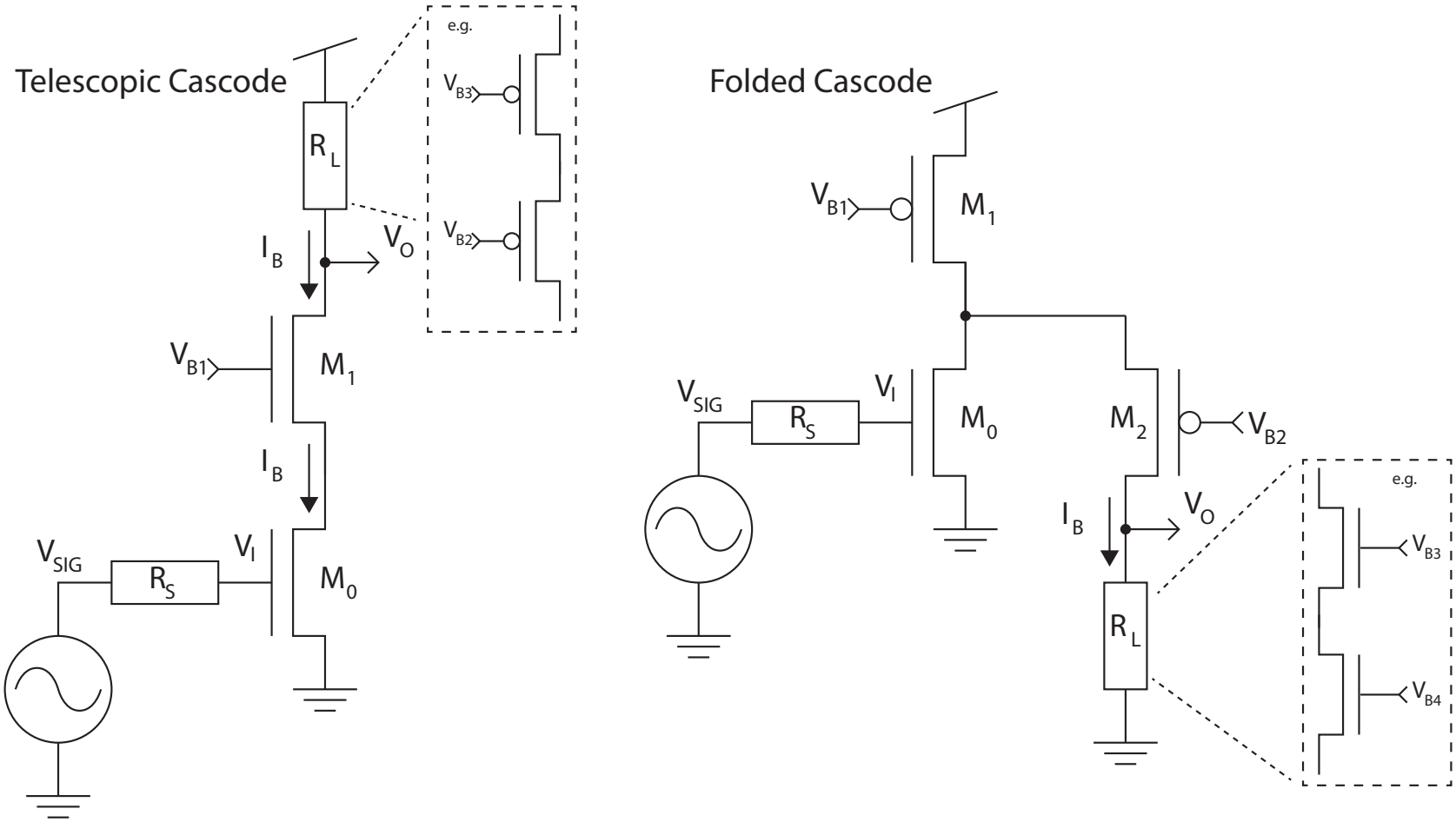


	R_I	G_M	R_O
CD	∞	$-g_m$	$\frac{1}{g_m}\|\|r_{ds} = \frac{r_{ds}}{1+g_m r_{ds}}$

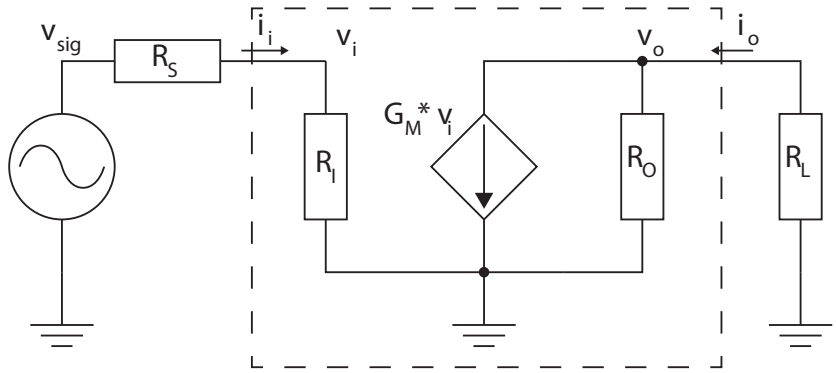
The source follower makes the output (the source) follow the input (the gate) with a gain close to 1. It can be used to turn a bad voltage (large R_S) source into a better voltage source (small R_O). Remember: in the large signal world there is an offset between input and output.

$$A' = g_m \left(\frac{1}{g_m} || r_{ds} || R_L \right) \approx 1$$

Cascode Gain Stage

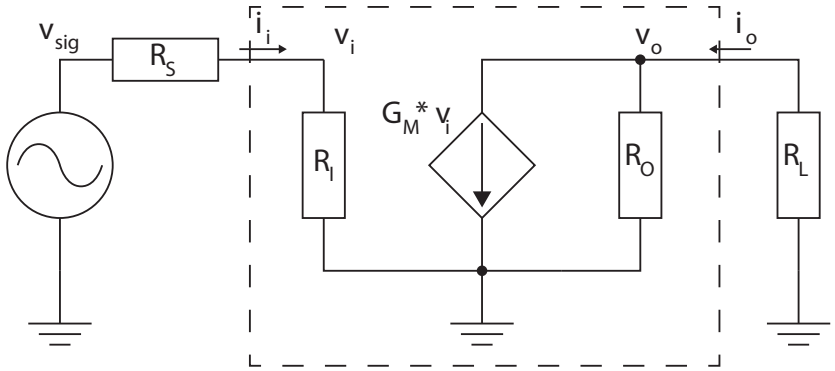


Cascode Gain Stages



	R_I	G_M	R_O
Cascode	∞	$g_{m1} r_{ds1} \frac{\frac{r_{ds2}}{1+g_{m2} r_{ds2}}}{r_{ds1} + \frac{r_{ds2}}{1+g_{m2} r_{ds2}}} \left(\frac{1}{r_{ds2}} + g_{m2} \right)$	$r_{ds2} + r_{ds1}(g_{m2} r_{ds2} + 1)$

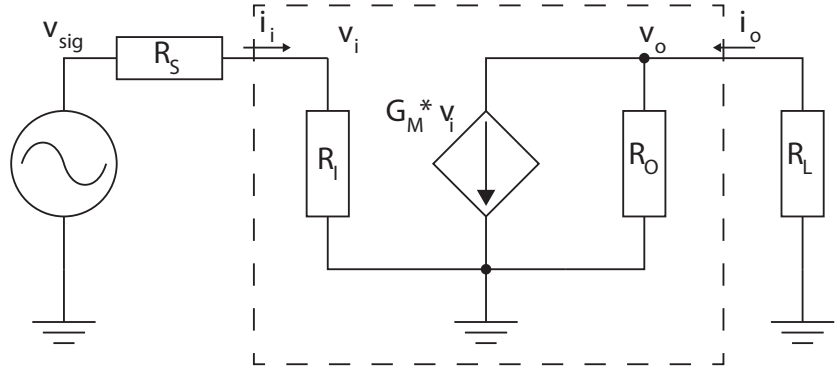
Cascode Gain Stages (1/2)



If one makes the assumptions that in general $g_m r_{ds} >> 1$ and $g_{m1} \approx g_{m2}$ and $r_{ds1} \approx r_{ds2}$ one can simplify these set of equations to:

$$\begin{aligned} R_O &\approx g_m r_{ds}^2 \\ G_M &\approx g_m \\ R_I &= \infty \end{aligned}$$

Cascode Gain Stages (2/2)



Note that $A' = -G_M(R_O||R_L)$, so it will depend a lot on R_L. A single pFET as load will not do: it needs to be a Cascode current source that also has an $R_O \approx g_m r_{ds}^2$ to get maximum gain.

$$A' = -G_M(R_O||R_L) = g_m(g_m r_{ds}^2||g_m r_{ds}^2) = \frac{1}{2} g_m^2 r_{ds}^2$$

With just a single pFET as load one is left again with about the same gain as a simple CS stage:

$$A' = -G_M(R_O||R_L) = g_m(g_m r_{ds}^2||r_{ds}) \approx g_m r_{ds}$$

(However: one can eliminate the Miller effect at the input that way, moving the pole at the input node to higher frequencies. See later.)

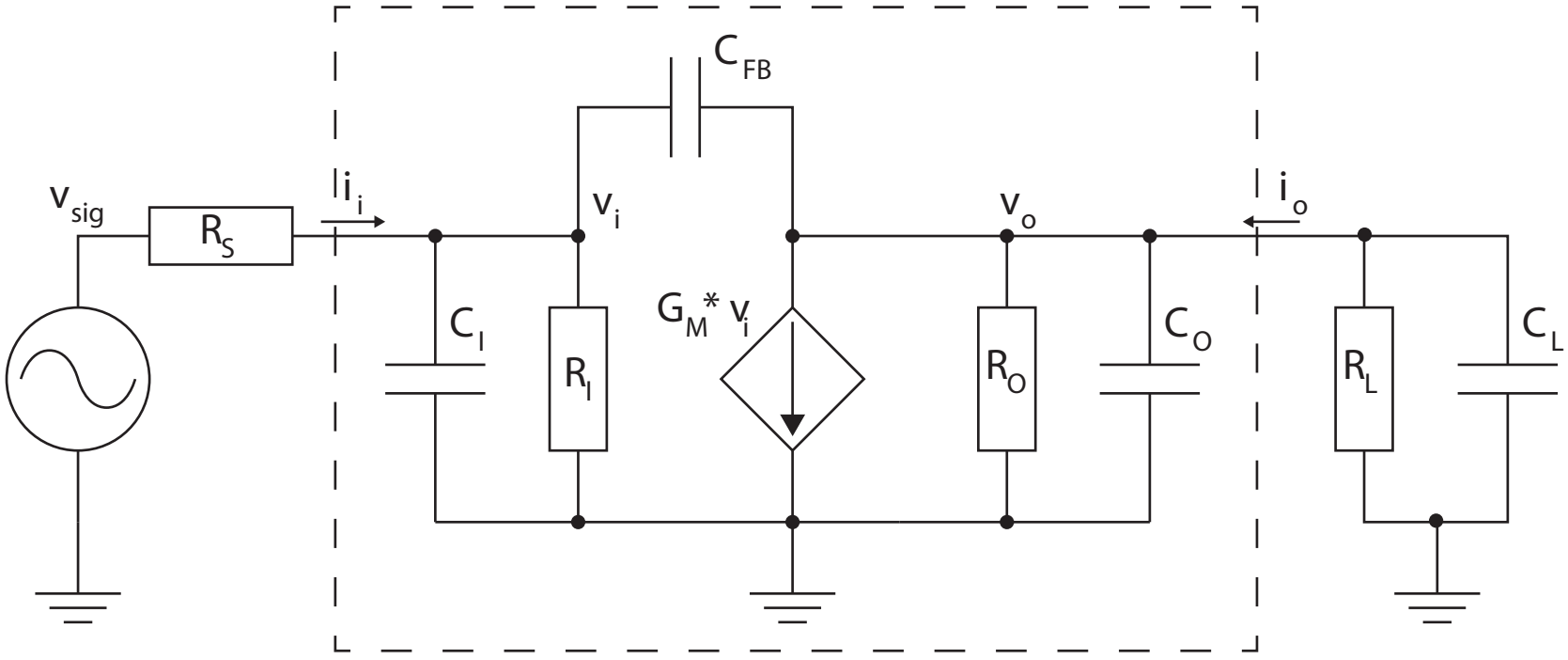
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General HF model for single transistor configurations



NOTE: we already deduced a general transfer function here in the linear systems chapter. Making the assumption of a dominant pole at either input node or output node we can use (5.36) and (5.30) and insert the appropriate parameters for CS, CG, and CD.

Assuming dominant pole at input

	$\tau_i = \frac{1}{\omega_{p_i}}$
CS	$R_S\left(C_{GS} + C_{GD}(g_m(r_{ds}\|\|R_L) + 1)\right)$
CG	$\left(R_S \| \frac{r_{ds}+R_L}{1+g_m r_{ds}}\right) C_{GS}$
CD	$R_S\left(C_{GD} + C_{GS}\left(1 - g_m\left(R_L \| \frac{1}{g_m} \| r_{ds}\right)\right)\right)$

CS suffers from the Miller effect, while the other two do not.

Assuming dominant pole at output

	$\tau_o = \frac{1}{\omega_{po}}$	ω_{zo}
CS	$(r_{ds}\|\|R_L)(C_{GD} + C_{DB} + C_L)$	$\frac{g_m}{C_{GD}}$
CG	$[(r_{ds} + R_S(g_m r_{ds} + 1))\|\|R_L](C_{GD} + C_{DB} + C_L)$	NA
CD	$\left(R_L\|\|\frac{1}{g_m}\|\|r_{ds}\right)(C_{GS} + C_{SB} + C_L)$	$\frac{-g_m}{C_{GS}}$

Remember from (5.34) that the zero is beyond the unity gain frequency, so not really of consequence. CG has a large output resistance so can result in a quite low pole frequency (dependent on R_L), so is the most likely candidate for a dominant pole at the output. CD has a very small output resistance, so is less likely to have a dominant pole at the output.

How to analyse

1. Compute both the potential dominant poles at the input and output.
2. If one of them has a much lower frequency (more than 4 times), use that one for a first order approximation.
3. If they are similar, the real cut off frequency will be somewhat lower than the lowest of the two. Use simulation tools or the full transfer function (5.26) to compute the Bode plot numerically.

Intrinsic Speed of CS stage (1/2)

Analogous to the *intrinsic gain* of a CS stage being the maximum gain under ideal load conditions (i.e. infinite input resistance load), one can ask the question: what is the maximum signal frequency that can pass through under ideal conditions for a CS stage? 'Passig through' here is defined as the *unity gain bandwidth* ω_T. the Sedra & Smith book answers that question by deriving the unity **current** gain, i.e. connect an ideal current source at the input and short the output to Gnd:

$$|i_o| = |i_i|$$
$$g_m|v_i| = |v_i s(C_{GS} + C_{GD})| = |v_i|\omega_T(C_{GS} + C_{GD})$$

so it follows that:

$$\omega_T = \frac{g_m}{C_{GS} + C_{GD}}$$

Intrinsic Speed of CS stage (2/2)

One gets the same result deriving the unity **voltage** gain using the observation in section 5.6:

$$\omega_T = |A_{DC}|\omega_{p_1} = \frac{G_M}{C_{FB} + C_L^*}$$

Assuming the total load capacitance C_L^* is of the same magnitude as the input capacitance $C_I = C_{GS}$, and where $C_{FB} = C_{GD}$, $A_{DC} = g_m(r_{ds}||R_L)$ and $\omega_{p_1} = \frac{1}{(r_{ds}||R_L)(C_L^* + C_{FB})}$

Trade off gain vs speed (1/2)

Note the *gain bandwidth product* (GBW) is the same as the unity voltage gain frequency ω_T of a first order system. It often expresses the trade off you can make between BW and gain, e.g. by employing a negative feedback loop.
(*drawing*)

Trade off gain vs speed (2/2)

Also note the inverse influence of changing L and V_{OV} on intrinsic GBW ω_T

$$\omega_T = \frac{\mu V_{OV}}{L\left(L_{OV} + \frac{2}{3}L\right)} \text{ strong inversion}$$

$$\omega_T = \frac{2\mu U_T e^{\frac{V_{OV}}{U_T}}}{L\left(L_{OV} + \frac{2}{3}L\right)} \text{ weak inversion}$$

and intrinsic gain A_O of a CS/FET (remember: λL is constant):

$$A_0 = \frac{2L}{\lambda L V_{OV}} \text{ strong inversion}$$

$$A_0 = \frac{L}{\lambda L U_T} \text{ weak inversion}$$

Stability

Q-factor from the general transamp transfer function (9.46):

$$Q = \frac{\sqrt{A}}{B} = \frac{\sqrt{R_I^* R_L^* \left[(C_{FB} + C_L^*)(C_{FB} + C_I) - C_{FB}^2\right]}}{R_L^*(C_{FB} + C_L^*) + R_I^* \left(C_I + C_{FB}(1 + G_M R_L^*)\right)}$$

No C_{FB}, no problem (CG)

Previous definitions of the time constants for dominant poles at input and output can be used for simplified expressions. With $C_{FB} = 0$ they reduce to $\tau_i = R_I^* C_I$ and $\tau_o = R_L^* C_L^*$, and we get

$$Q^2 = \frac{\tau_i \tau_o}{(\tau_i + \tau_o)^2} = \begin{cases} < \frac{1}{4} \text{ if } \tau_i \neq \tau_o \\ = \frac{1}{4} \text{ if } \tau_i = \tau_o \end{cases}$$

$$\Rightarrow Q \leq \frac{1}{2}$$

With C_{FB} and $G_M \geq 0$ (CS)

Again using the dominant pole approximations at I/O nodes for simpler notation, now with C_{FB}: $\tau_i = R_I^* \left(C_I + C_{FB}(G_M R_L^* + 1)\right)$ and $\tau_o = R_L^*(C_{FB} + C_L^*)$

$$
\begin{aligned}
Q^2 &= \frac{R_I^* \left(C_I + C_{FB}\right)\tau_o - R_I^* R_L^* C_{FB}^2}{\left(\tau_o + \tau_i\right)^2} \\
&= \frac{(\tau_i - R_I^* C_{FB} G_M R_L^*)(\tau_o) - R_I^* R_L^* C_{FB}^2}{\left(\tau_o + \tau_i\right)^2} \\
&\leq \frac{(\tau_i - R_I^* C_{FB} G_M R_L^*)(\tau_o)}{\left(\tau_o + \tau_i\right)^2} \\
&\leq \frac{\tau_i \tau_o}{\left(\tau_o + \tau_i\right)^2} \\
&\Rightarrow Q \leq \frac{1}{2}
\end{aligned}
$$

With C_{FB}, $G_M < 0$, and $(1 + G_M R_L^*) \approx 0$ (CD/SF)

Assuming exact $(1 + G_M R_L^*) = 0$ again $\tau_i = R_I^* C_I$ and we define $\tau_o' := \tau_o - R_L^* C_{FB} = R_L^* C_L^*$, we get

$$Q^2 = \frac{\tau_i \tau_o' + \tau_i R_L^* C_{FB} + \tau_o' R_I^* C_{FB}}{(\tau_o + \tau_i)^2} > \frac{1}{4}$$

$$\Leftrightarrow$$

$$\frac{1}{2}\left(\frac{\tau_i}{\tau_o} + \frac{\tau_o}{\tau_i}\right) - 1 < \frac{2 C_L^* C_{FB}}{C_I (C_L^* + C_{FB})}$$

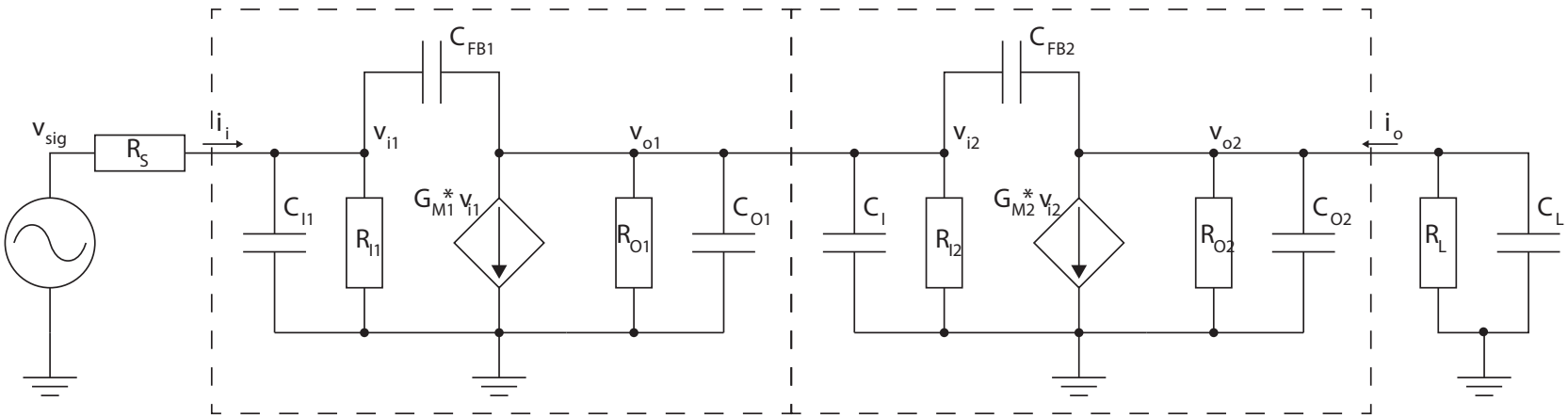
The left-hand side is minimally 0 if

$$\tau_i = \tau_o$$

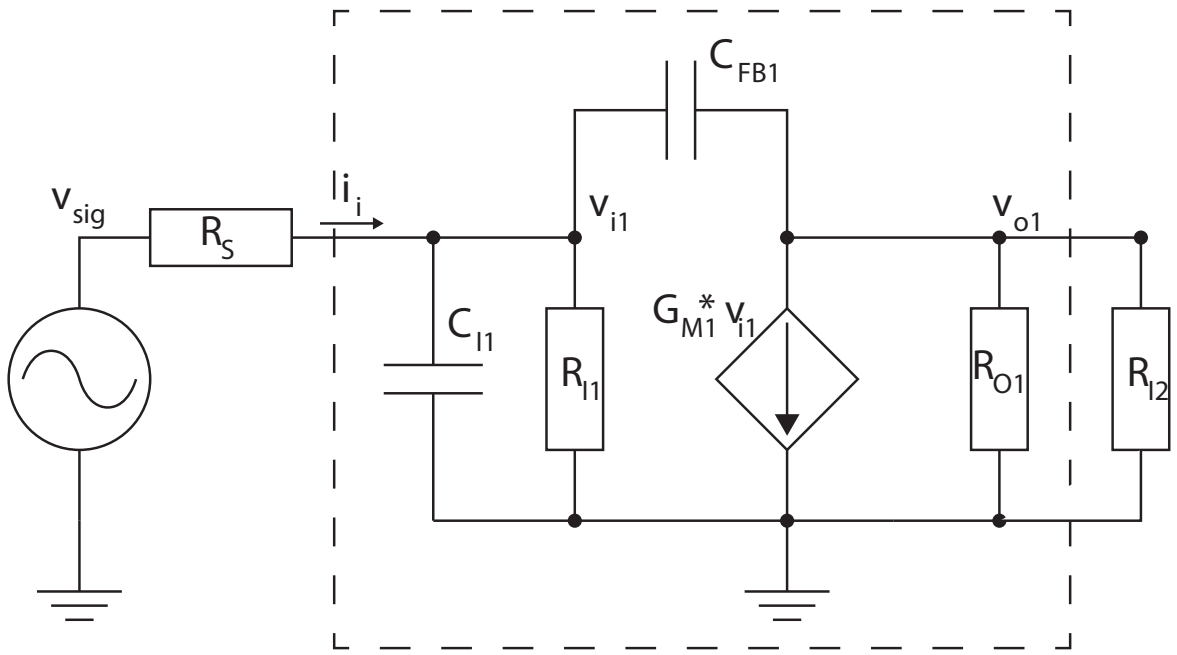
and since the righ-thand term above is >0, resonance $Q^2 > \frac{1}{4}$ is guaranteed.

Cascode gain stage frequency behaviour

Mapping onto single transamp equivalent HF model is unfortunately not possible due to the internal nodes capacitances ...

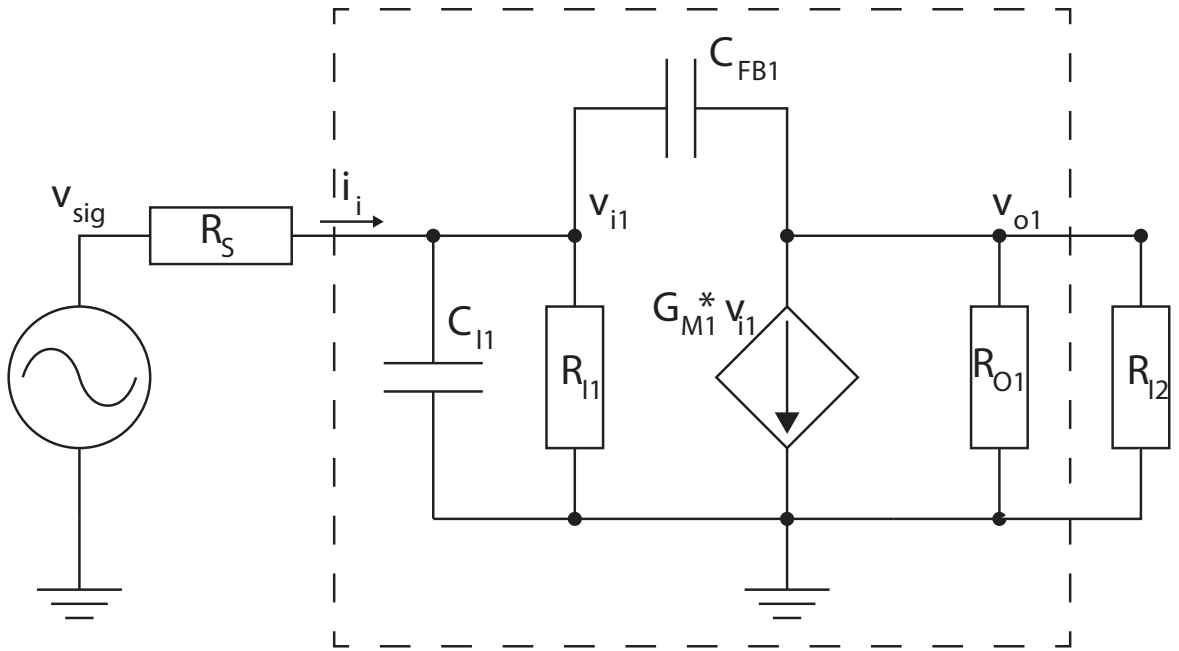


Cascode gain stage dominant pole at input (1/3)



However, once again we can draw some conclusions making the strong assumption of a dominant pole at the input. Tweaking R_L and thus R_{I2} offers some interesting gain/bandwidth trade off, increasing the gain-bandwidth product (GBW) beyond just a CS stage.

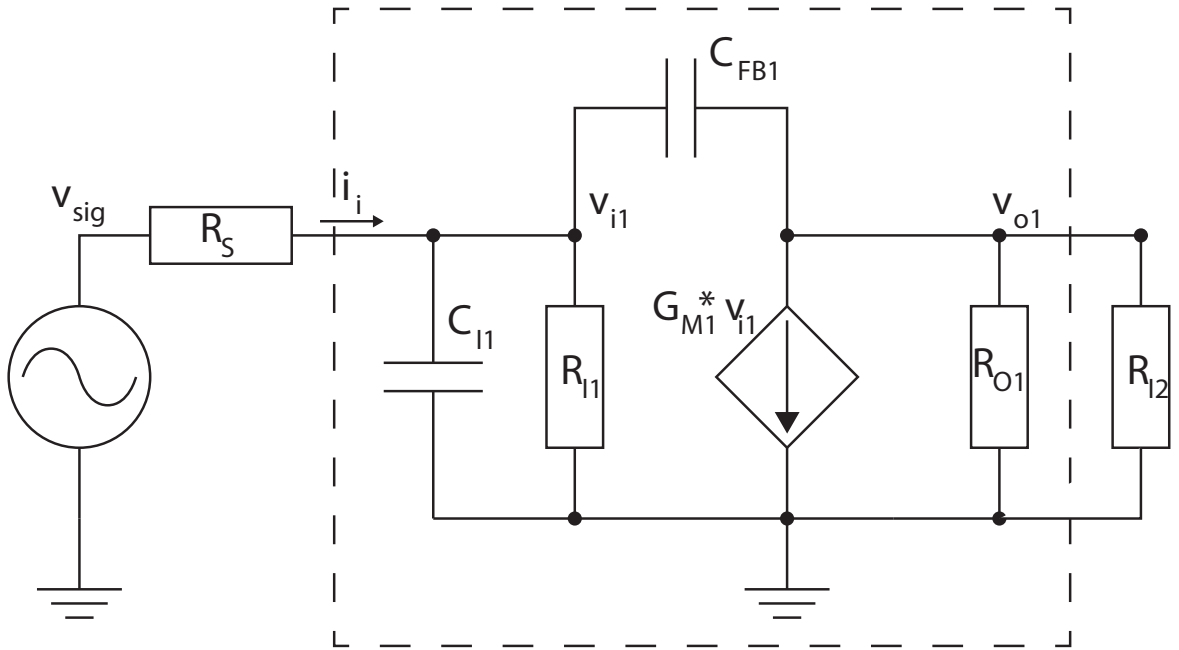
Cascode gain stage dominant pole at input (2/3)



$$R_{I2} = \frac{r_{ds2} + R_L}{1 + g_{m2} r_{ds2}} \tag{1}$$

So if $R_L \approx r_{ds}$ we get $R_{I2} \approx \frac{1}{g_m}$ and minimal gain in this first stage and thus minimal Miller effect, while the total gain is still $A'_{DC} \approx g_m r_{ds}$.

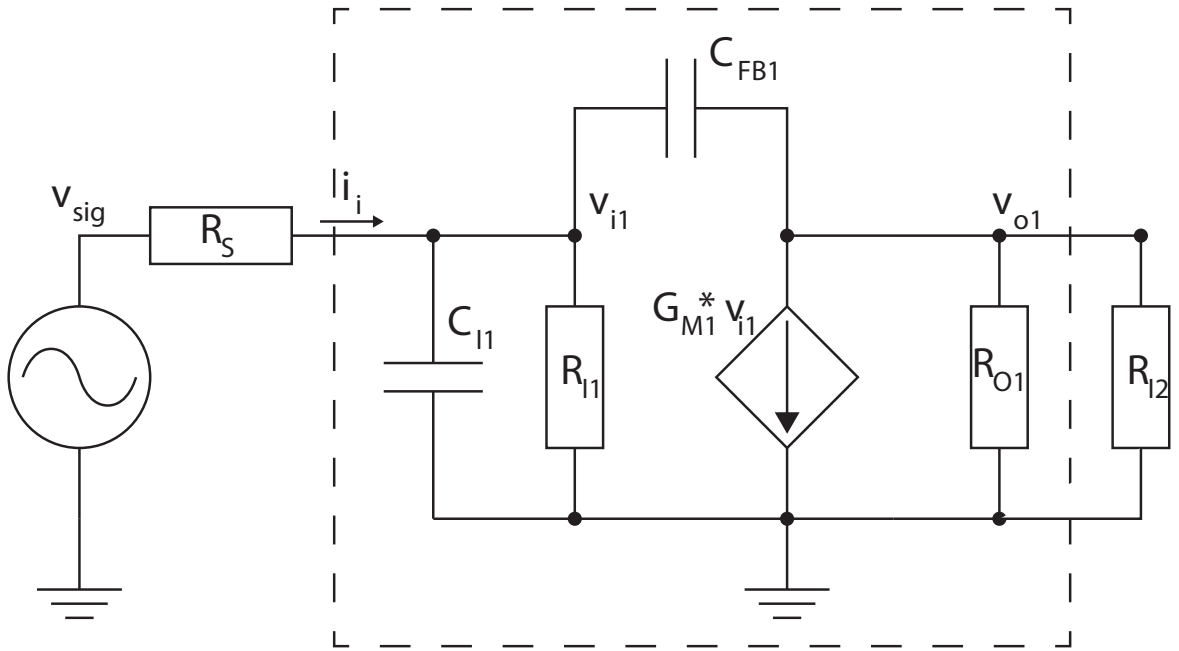
Cascode gain stage dominant pole at input (2/3)



$$R_{I2} = \frac{r_{ds2} + R_L}{1 + g_{m2} r_{ds2}} \tag{2}$$

So if $R_L \approx r_{ds}$ we get $R_{I2} \approx \frac{1}{g_m}$ and minimal gain in this first stage and thus minimal Miller effect, while the total gain is still $A' \approx g_m r_{ds}$.

Cascode gain stage dominant pole at input (3/3)

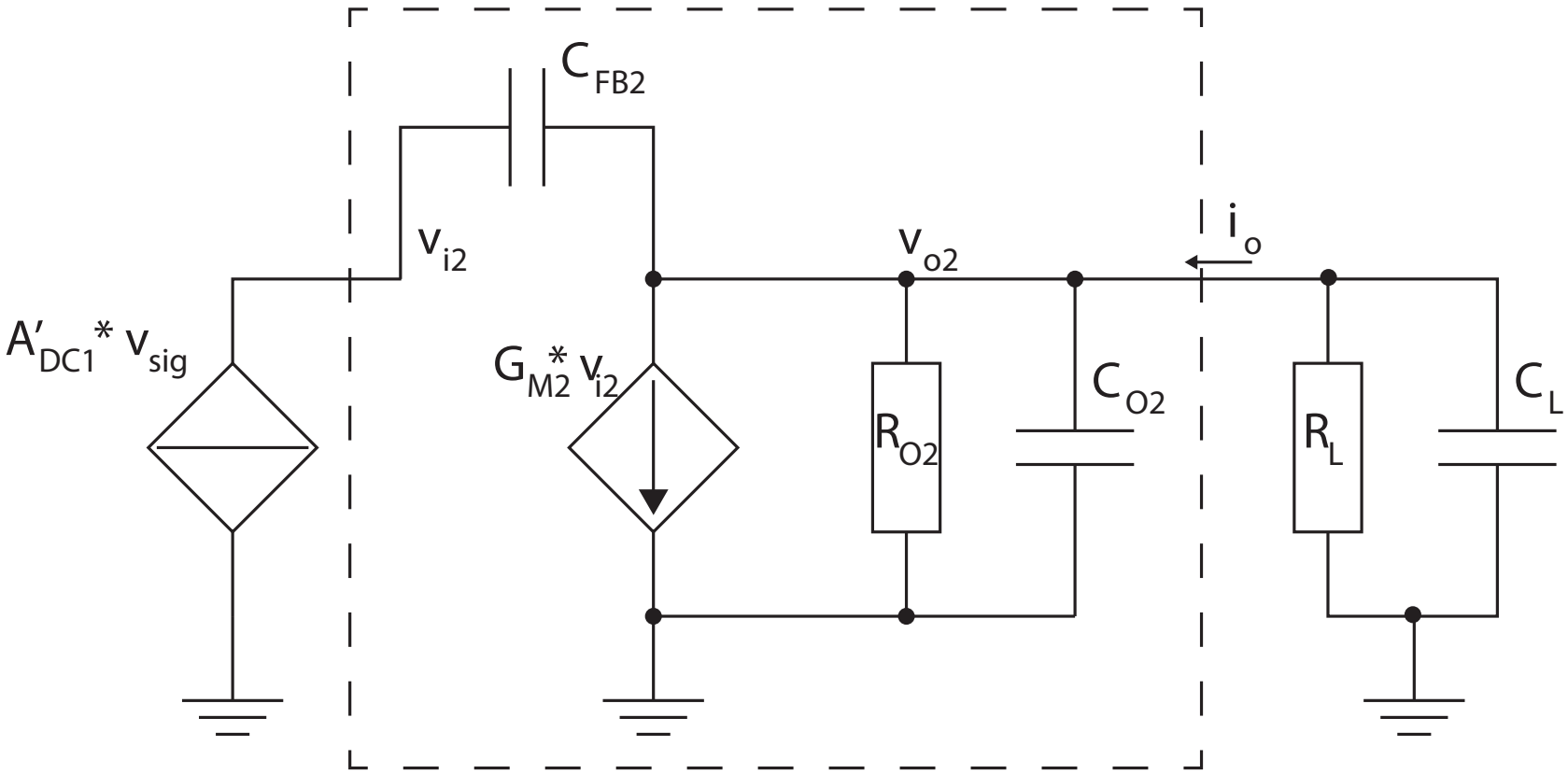


$$R_{I2} = \frac{r_{ds2} + R_L}{1 + g_{m2} r_{ds2}} \tag{3}$$

If on th other hand $R_L \approx g_m r_{ds}^2$ we get $R_{I2} \approx r_{ds}$ and 'normal' CS gain ($\approx \frac{1}{2} g_m r_{ds}$) in this first stage and thus the same Miller effect and BW like a normal CS stage, but the total gain now $A'_{DC} \approx \frac{1}{2} g_m^2 r_{ds}^2$, i.e. the square of a 'normal' CS stage.

Cascode gain stage dominant pole at output

Somewhat less beneficial trade off when assuming a dominant pole at the output, without improving the GBW, when assuming the dominant pole at the output.



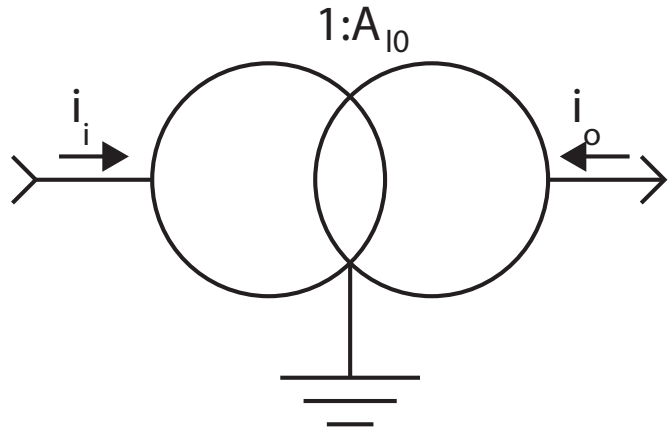
Content

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Current Mirror Symbol

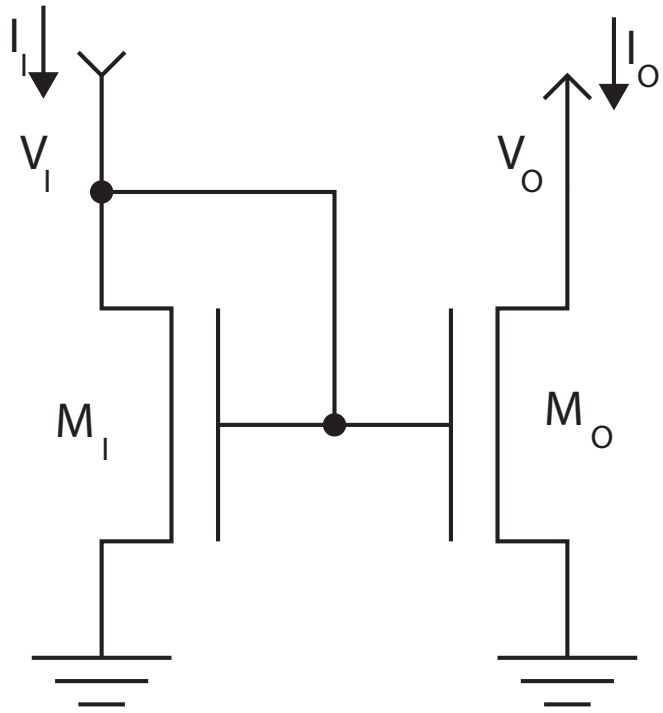


A current amplifier with intrinsic current gain A_{I0}. Often just copying a current with $A_{I0} = 1$. Also useful to generate bias currents from a master reference current.

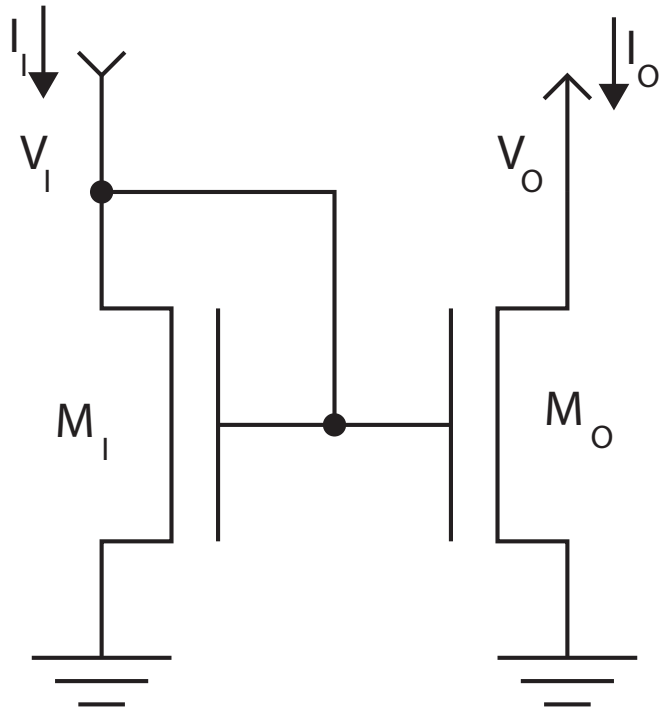
$$A_I := \frac{i_o}{i_{sig}} = A_{I0} \frac{R_O}{R_O + R_L} \frac{R_S}{R_I + R_S}$$

Ideally one would have $R_O = R_S = \infty$ and/or $R_I = R_L = 0$.

Simple Implementation



Simple Implementation

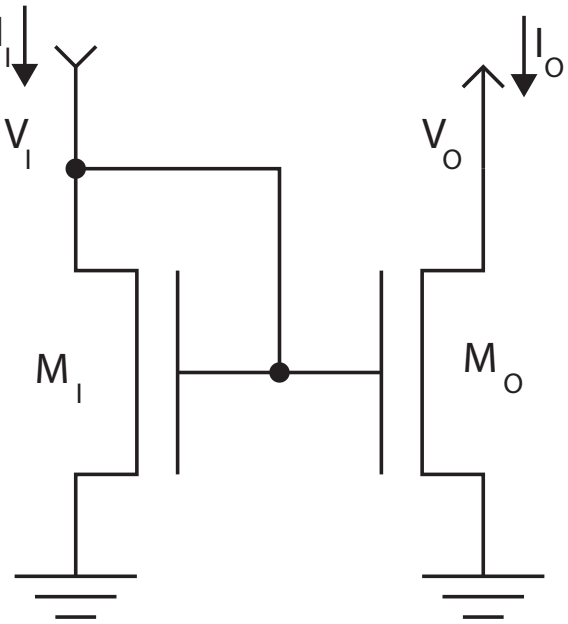


$$K := \frac{W_O L_I}{L_O W_I}$$

$$I_O \approx K I_I$$

$$V_D \overset{!}{\geq} V_{sat}$$

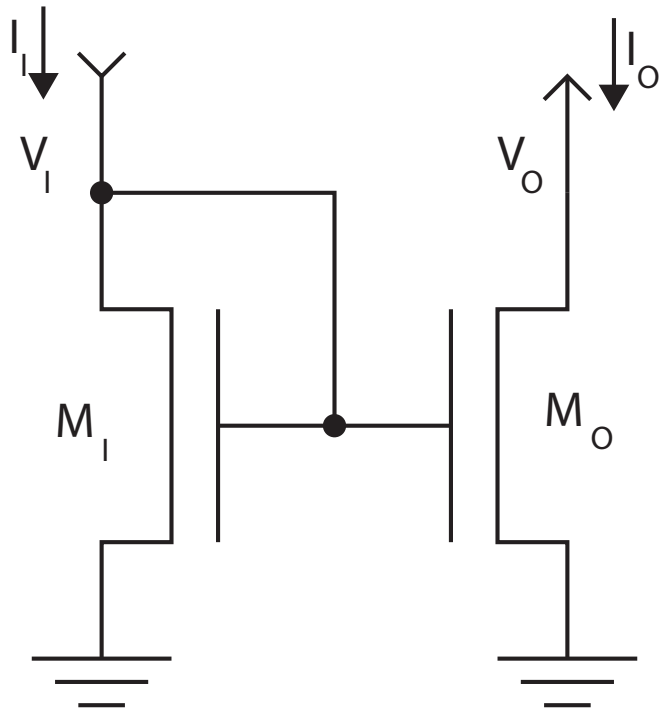
Simple Implementation Large Signal Considerations



I_O will be affected by an offset. One way to analyse it is assuming we know V_I and V_O (will depend on R_S and R_L!) and defining I'_D as the I_I if λ were zero, and using r_{ds} in saturation:

$$\begin{aligned} KI_I &= KI'_D(1 + V_{DI}\lambda) \\ I_O &= KI'_D(1 + V_{DO}\lambda) \\ I_O - KI_I := \epsilon &= KI'_D(\Delta V_D \lambda) = K\frac{\Delta V_D}{r_{dsI}} = \frac{\Delta V_D}{r_{dsO}} \end{aligned} \tag{4}$$

Simple Implementation Small Signal Considerations

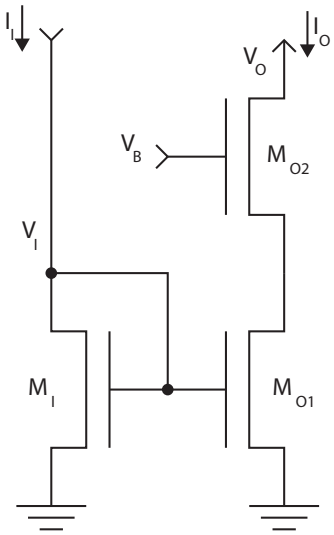


Once the large signal PO is found with a small deviation from $I_O \approx KI_I$ (or often just ignoring that small deviation) one can more precisely determine the small signal behaviour (or also just assume that $i_o \approx Ki_i$):

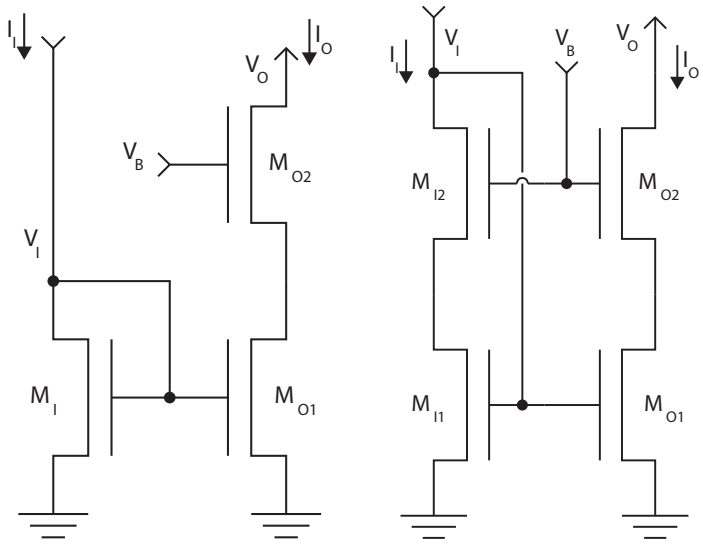
$$A_I(0) = \frac{i_o}{i_{sig}} = K \underbrace{\frac{R_S}{\left(r_{dsI} || \frac{1}{g_{mI}}\right) + R_S}}_{\text{due to } R_S \neq 0} \underbrace{\frac{g_{mI} r_{dsI}}{1 + g_{mI} r_{dsI}}}_{\text{due to } A_{IO} \neq K} \underbrace{\frac{r_{dsO}}{r_{dsO} + R_L}}_{\text{due to } R_O \neq \infty} \quad (5)$$

Improving R_O

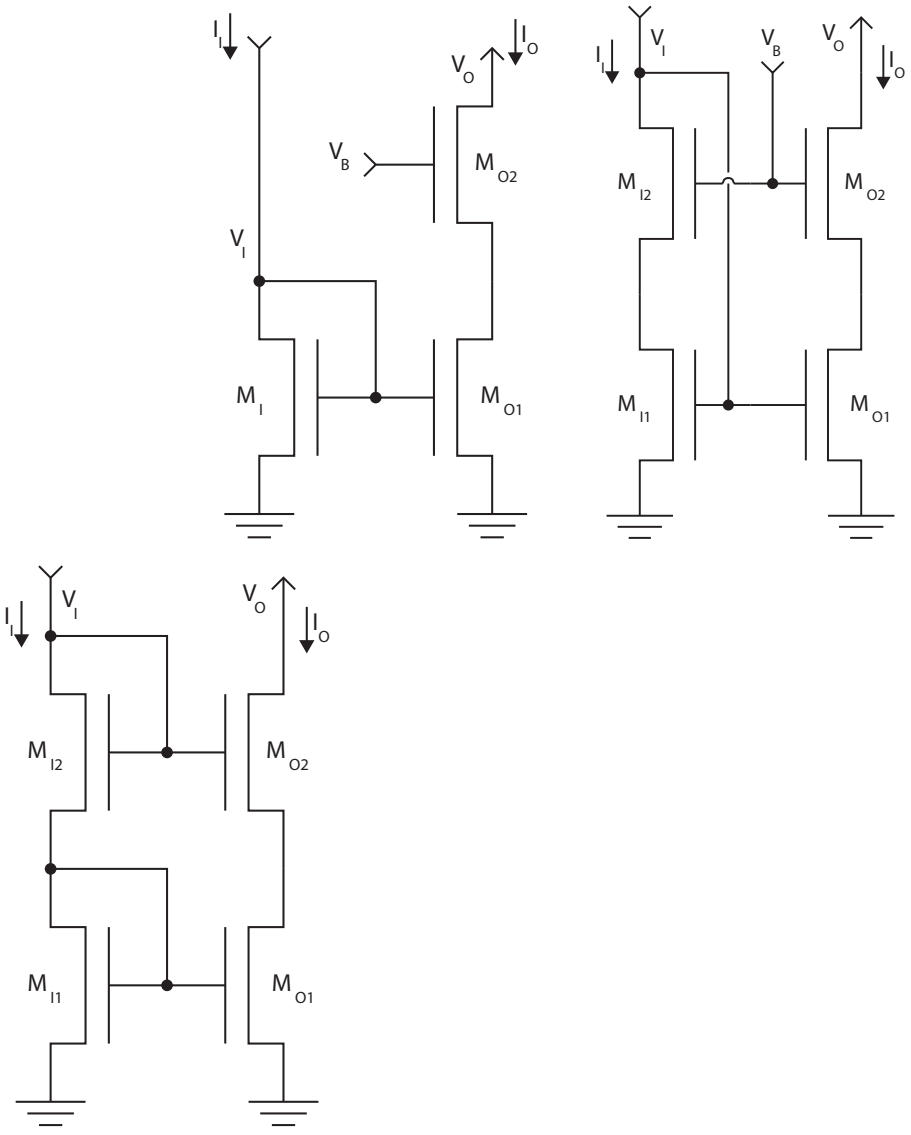
Improving R_O



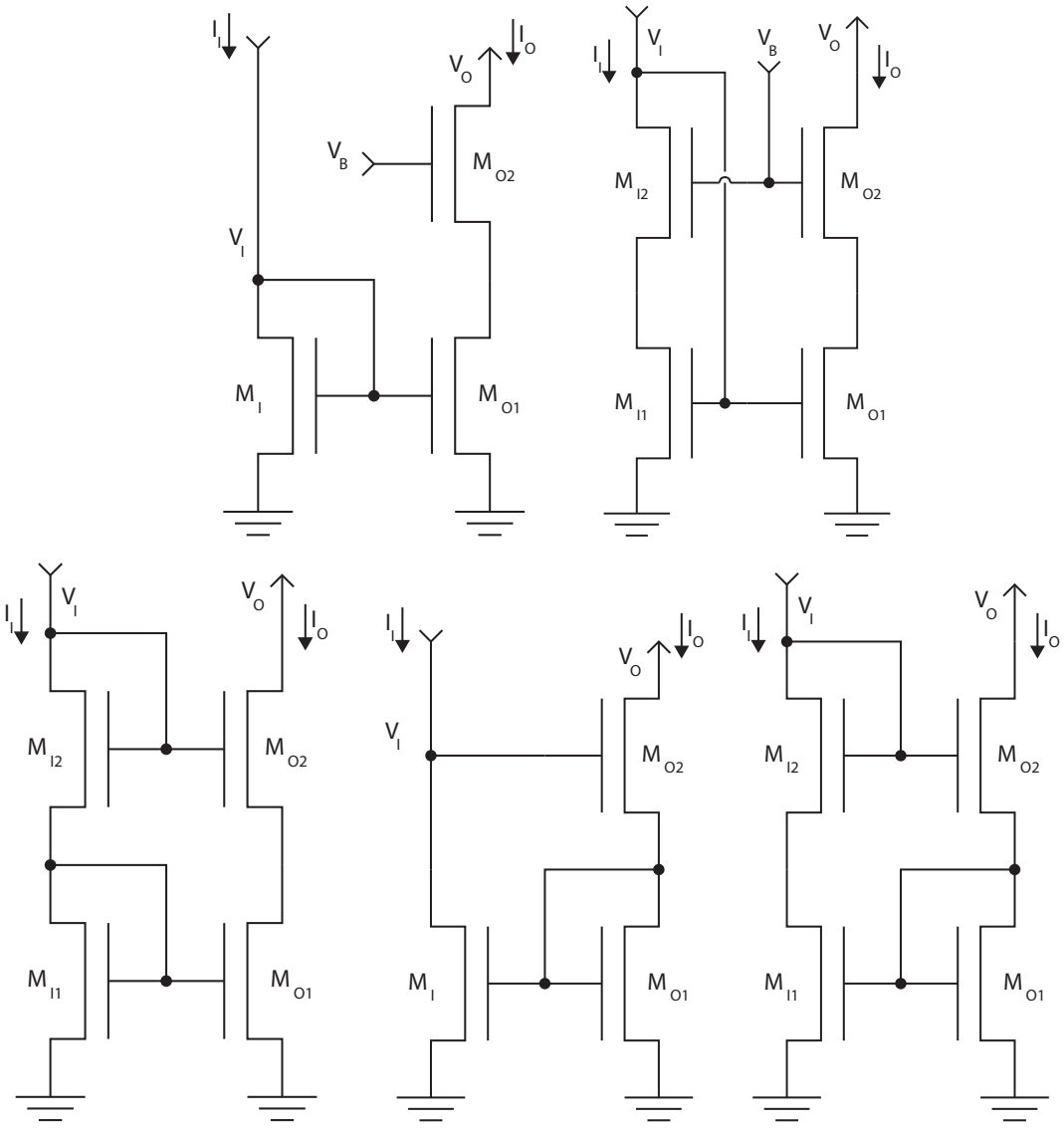
Improving R_O



Improving R_O



Improving R_O



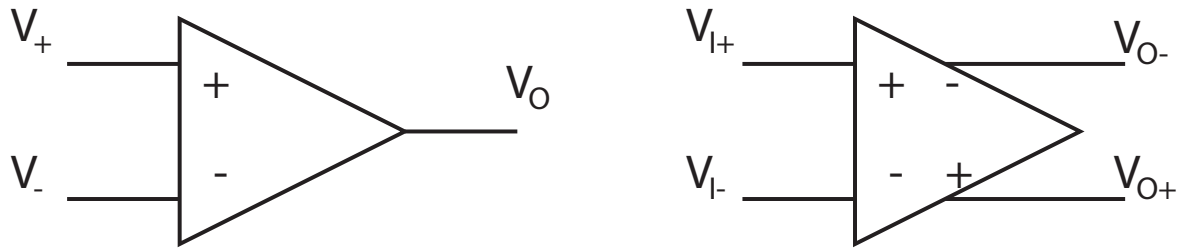
Content

Single Ended Amplifiers

Current Mirrors

Differential Amplifiers
- Differential pair
- Resistor loaded diff pair
- Current mirror loaded diff pair
- Folded Cascode Operational Amplifier

Differential amplifiers



Differential amplifiers are a more versatile tool than single ended ones. Differential signalling offers some noise immunity and the possibility to interface different signal domains, as well as more possibilities for amplifier circuits, e.g. negative feedback circuits.

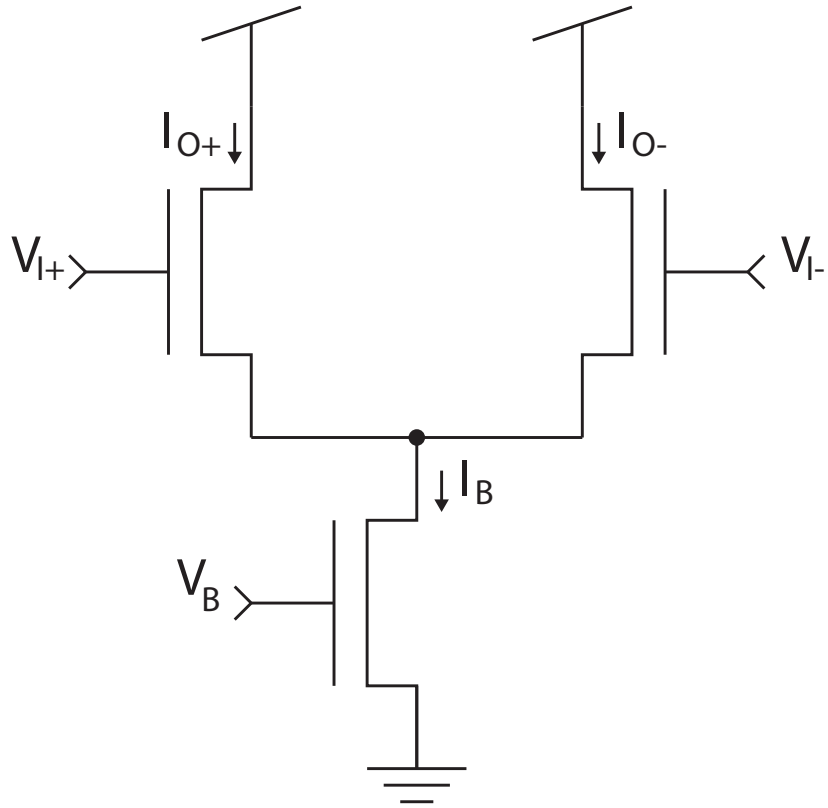
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The differential pair



Splits a bias current between two branches dependent on the differences of input voltages: approximately linearly when the input voltages are close (fully differential **transconductance amplifier**), and digitally if the difference exceeds a certain value (**comparator**).

Small signal analysis

Considering two types of correlated inputs: **common mode input $\hat{v}_i$** and **differential input Δv_i**:

$$\hat{v}_i := \frac{v_{i+} + v_{i-}}{2}$$

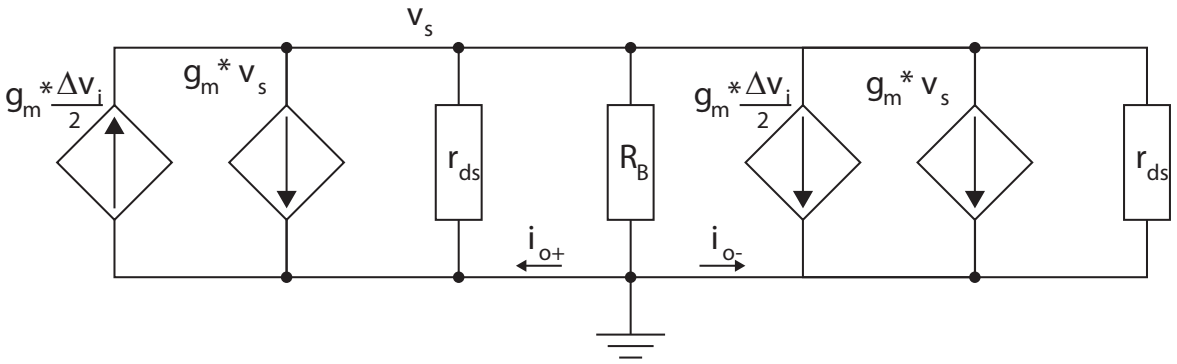
$$\Delta v_i := v_{i+} - v_{i-}$$

So:

$$v_{i+} = \hat{v}_i + \frac{\Delta v_i}{2}$$
$$v_{i-} = \hat{v}_i - \frac{\Delta v_i}{2}$$

We conduct the analysis for them changing separately. PO: the two large signal inputs are equal (transconductance amplifier).

Analysis for differential input



Kirchhoff:

$$i_{o+} + v_s\left(g_m + \frac{1}{r_{ds}}\right) = g_m\frac{\Delta v_i}{2}$$

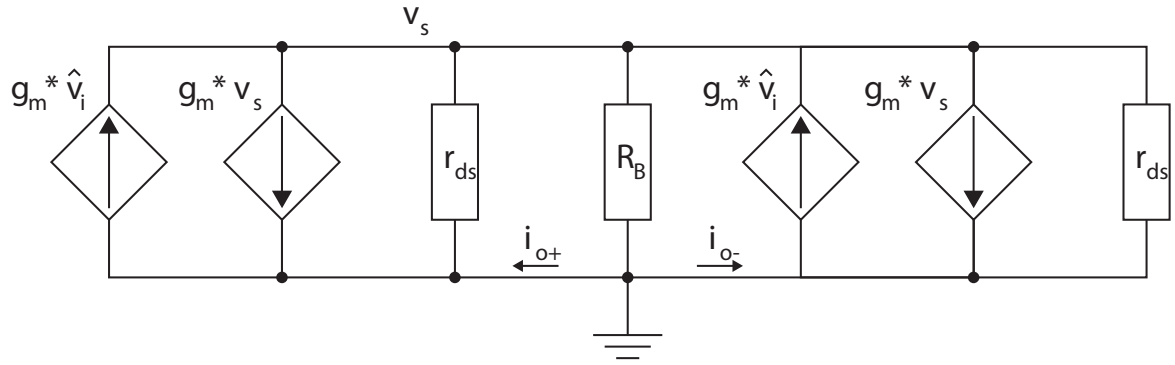
$$i_{o-} + v_s\left(g_m + \frac{1}{r_{ds}}\right) = -g_m\frac{\Delta v_i}{2}$$

$$v_s\left(2g_m + 2\frac{1}{r_{ds}} + \frac{1}{R_B}\right) = \frac{\Delta v_i}{2}(g_m - g_m) = 0$$

Result:

$$\Delta i_o = g_m \Delta v_i$$

Analysis for common mode input



Kirchhoff:

$$i_{o+} = i_{o-} = \hat{v}_i g_m - v_s \left(g_m + \frac{1}{r_{ds}} \right)$$

$$v_s \left(2g_m + 2\frac{1}{r_{ds}} + \frac{1}{R_B} \right) = 2\hat{v}_i g_m$$

Result:

$$i_{o+} = i_{o-} = \hat{v}_i \frac{\frac{g_m}{R_B}}{2g_m + 2\frac{1}{r_{ds}} + \frac{1}{R_S}}$$

No effect on difference between the two output currents (because of symmetry). CAVEAT: change of PO for significant $\hat{V}_i$ necessitating reevaluation of g_m!

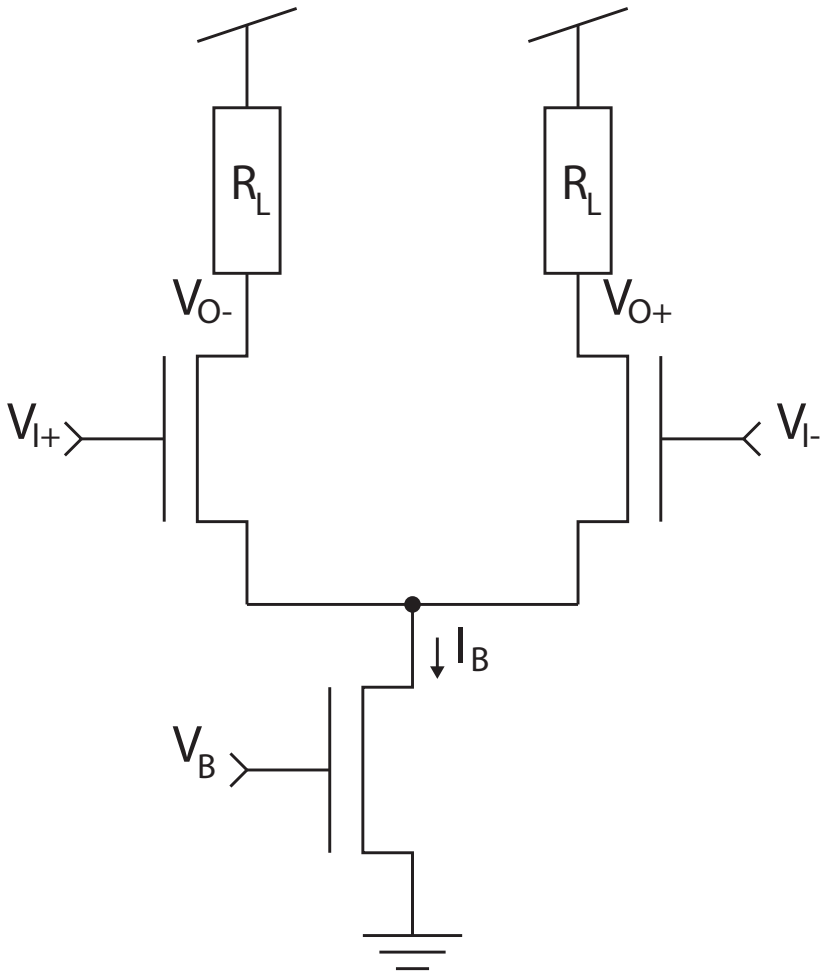
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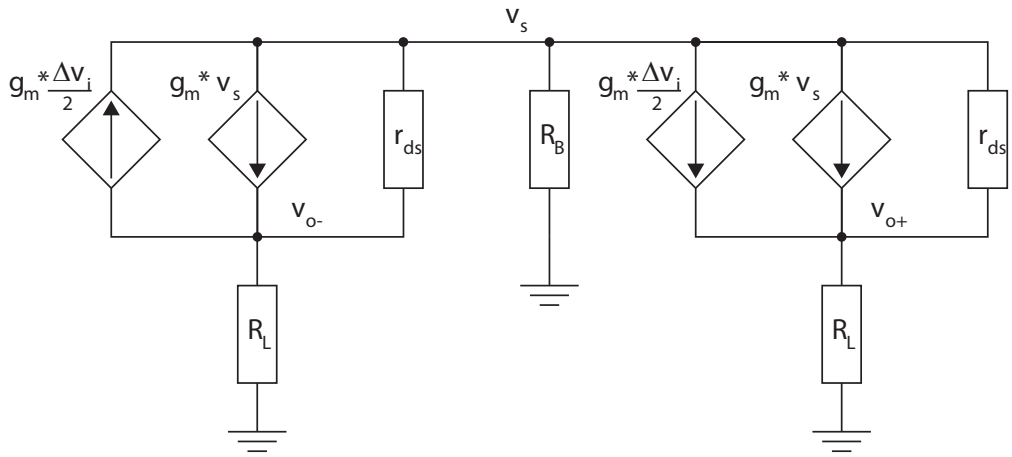
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Resistor load diff output



Resistor load small signal analysis



Kirchhoff:

$$v_{o+}(\frac{1}{R_L}+\frac{1}{r_{ds}})-g_m*\frac{\Delta v_i}{2}-g_m v_s=v_s\frac{1}{r_{ds}}$$

$$v_{o-}(\frac{1}{R_L}+\frac{1}{r_{ds}})+g_m*\frac{\Delta v_i}{2}-g_m v_s=v_s\frac{1}{r_{ds}}$$

$$v_s(\frac{1}{R_B}+\frac{2}{r_{ds}}+2g_m)=v_{o+}\frac{1}{r_{ds}}+v_{o-}\frac{1}{r_{ds}}$$

Result:

$$A'=\frac{\Delta v_o}{\Delta v_i}=g_m(R_L||r_{ds})$$

Common mode rejection ratio (CMRR)

The common mode gain is defined as

$$A_{\mathrm{CM}} = \frac{\Delta v_o}{\hat{v}_i}$$

for differential output and later for single ended output:

$$A_{\mathrm{CM}} = \frac{v_o}{\hat{v}_i}$$

Here for the resistor load differential output, trivially (because of symmetry) $A_{\mathrm{CM}} = 0$ and the so called Common mode rejection ratio (CMRR)

$$CMMR := \frac{|A|}{|A_{\mathrm{CM}}|} = \infty$$

is infinite. But again BEWARE a gain error that will result from changes of the PO, i.e. $\hat{V}_i$ that are not reflected by the CMRR!

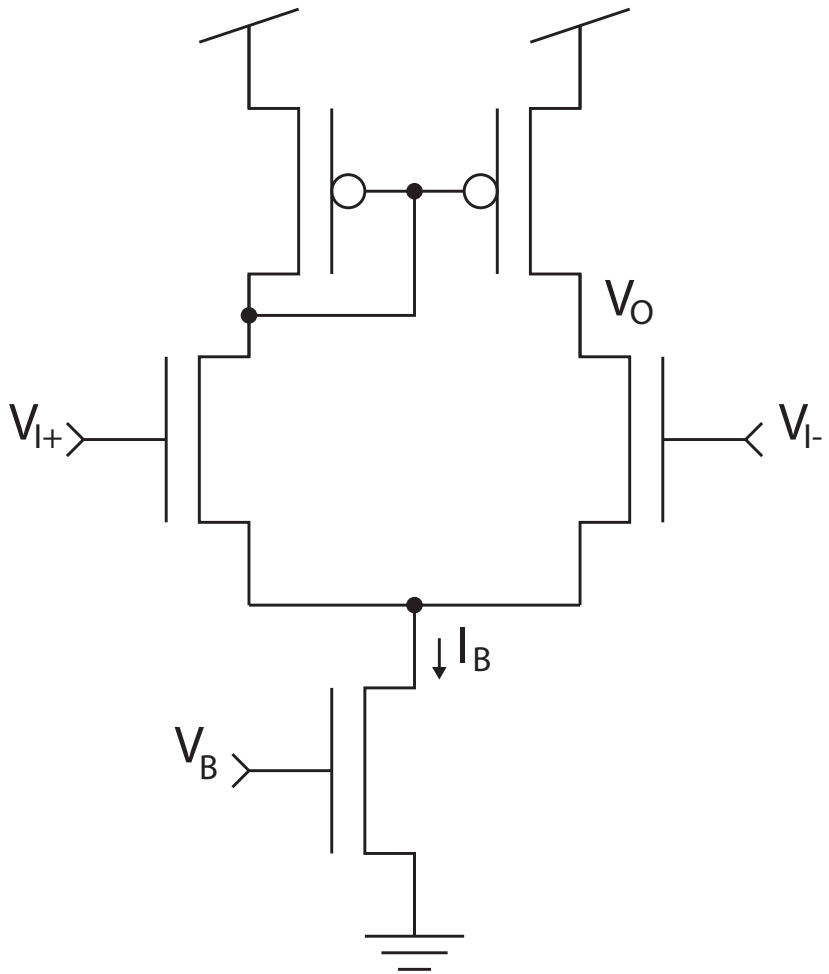
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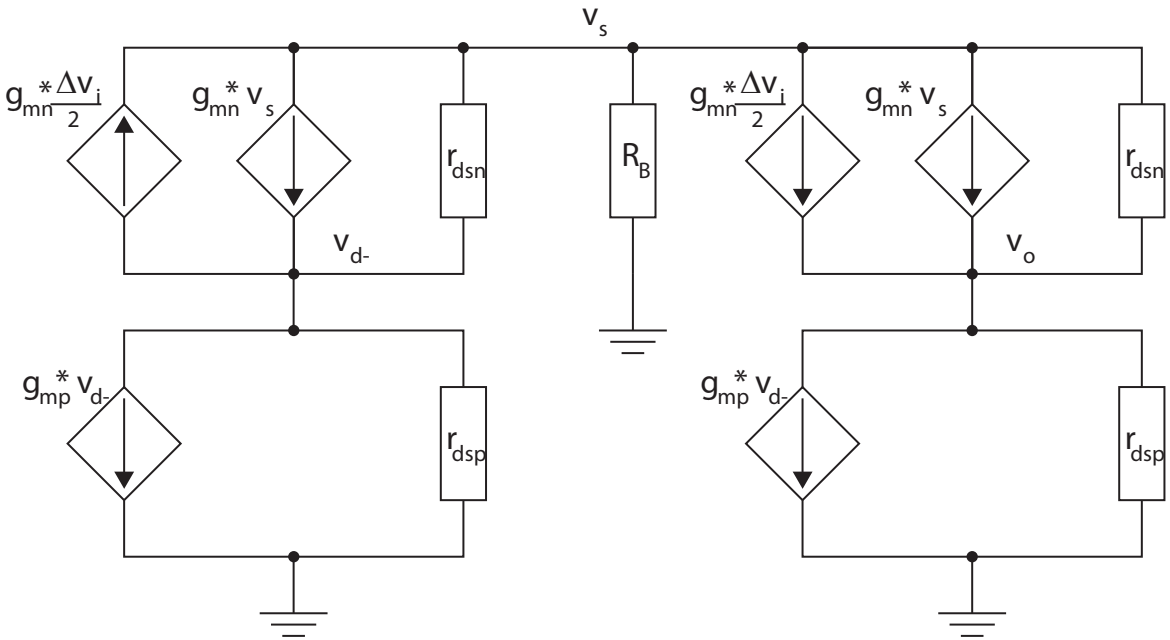
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Current mirror loaded diff pair



Δv_i small signal (1/3)



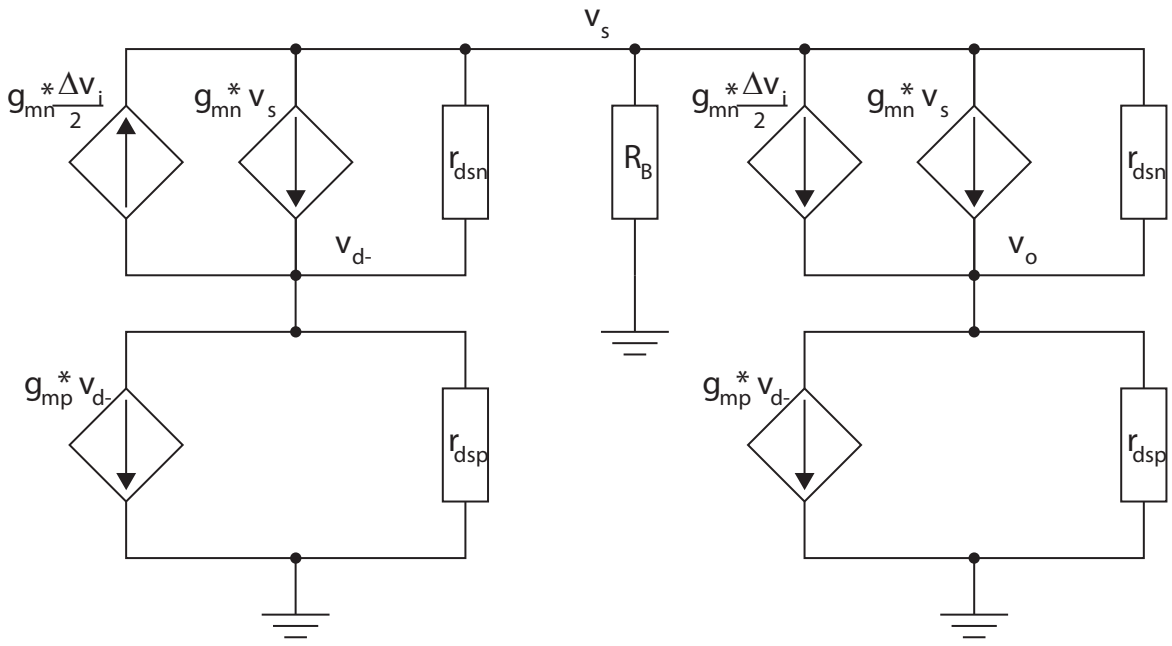
Kirchhoff:

$$v_d\left(\frac{1}{r_{dsn}} + \frac{1}{r_{dsp}} + g_{mp}\right) + \frac{\Delta v_i}{2} g_{mn} = v_s\left(g_{mn} + \frac{1}{r_{dsn}}\right)$$

$$v_s\left(2\frac{1}{r_{dsn}} + 2g_{mn} + \frac{1}{R_B}\right) = v_d\frac{1}{r_{dsn}} + v_o\frac{1}{r_{dsn}}$$

$$v_o\left(\frac{1}{r_{dsn}} + \frac{1}{r_{dsp}}\right) + v_d g_{mp} = v_s\left(g_{mn} + \frac{1}{r_{dsn}}\right) + \frac{\Delta v_i}{2} g_{mn}$$

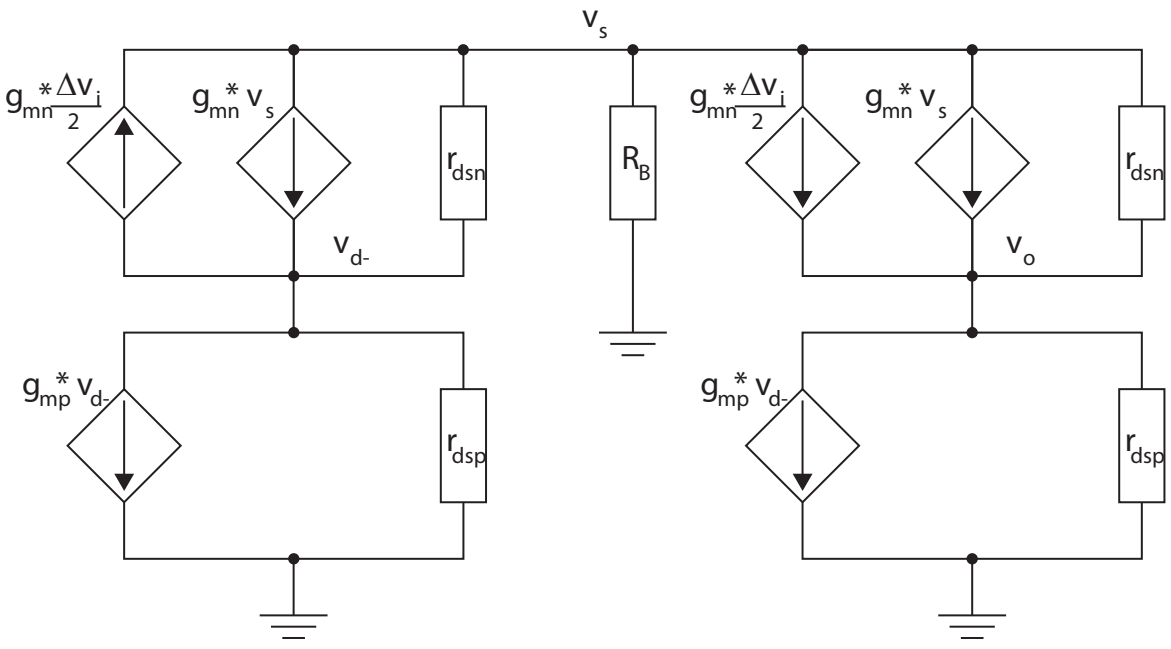
Δv_i small signal (2/3)



Result:

$$A = \frac{v_o}{\Delta v_i} = g_{mn}\left(r_{dsn}||r_{dsp}\right) * \frac{2g_{mp}g_{mn} + 2\frac{g_{mp}}{r_{dsn}} + \frac{g_{mp}}{R_B} + \left(\frac{g_{mn}}{r_{dsp}} + \frac{1}{r_{dsn}r_{dsp}} + \frac{1}{2r_{dsn}R_B} + \frac{1}{2r_{dsp}R_B}\right)}{2g_{mn}g_{mp} + 2\frac{g_{mp}}{r_{dsn}} + \frac{g_{mp}}{R_B} + 2\left(\frac{g_{mn}}{r_{dsp}} + \frac{1}{r_{dsp}r_{dsn}} + \frac{1}{2r_{dsn}R_B} + \frac{1}{2r_{dsp}R_B}\right)}$$

Δv_i small signal (3/3)

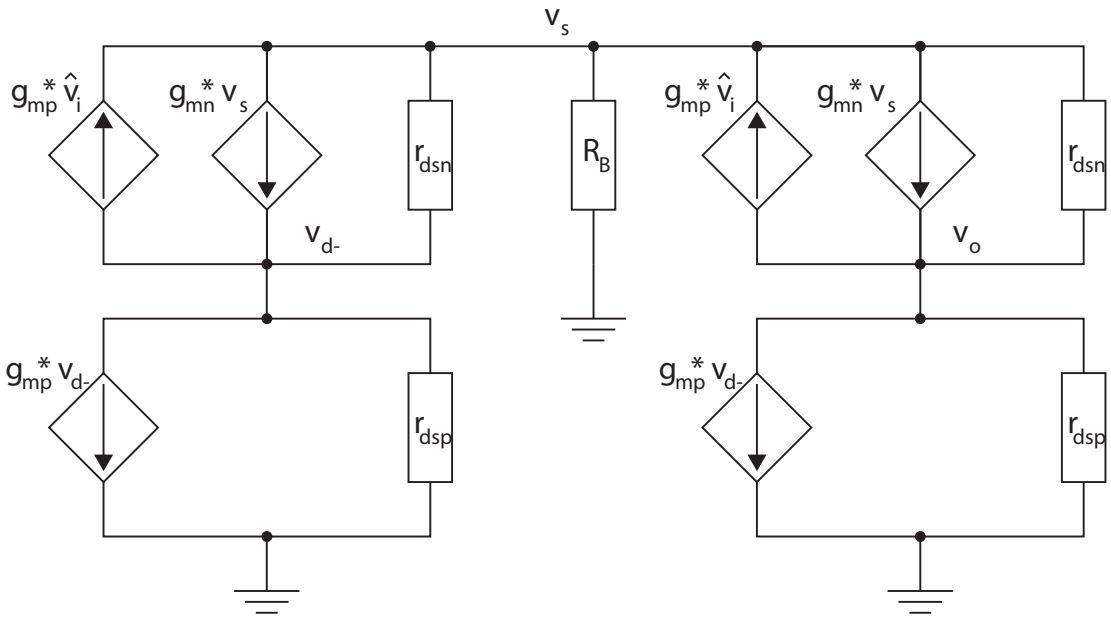


simplified assuming $g_m r_{ds} >> 1$ and $R_B \approx r_{dsp} \approx r_{dsn}$ and $g_{mn} \approx g_{mp}$

$$A = \frac{v_o}{\Delta v_i} \approx g_{mn} \left(r_{dsn} || r_{dsp} \right)$$

For example if $g_m r_{ds} = 10$ then $A = g_{mn} \left(r_{dsn} || r_{dsp} \right) * 0.953$

$\hat{v}_i$ small signal (1/3)



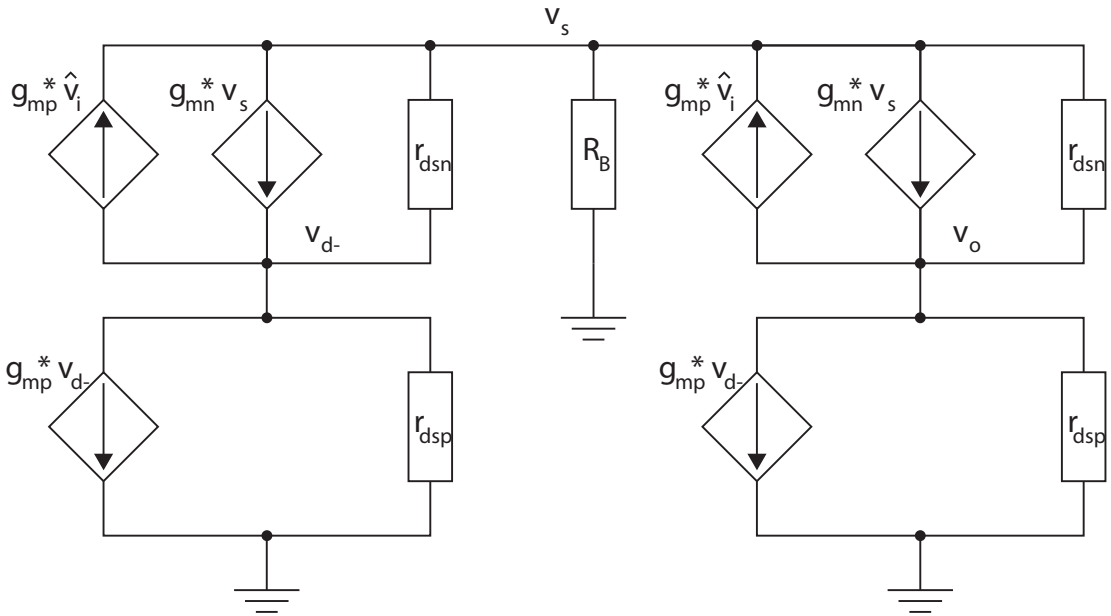
Kirchhoff:

$$v_d\left(\frac{1}{r_{dsn}} + \frac{1}{r_{dsp}} + g_{mp}\right) + \hat{v}_i g_{mn} = v_s\left(g_{mn} + \frac{1}{r_{dsn}}\right)$$

$$v_s\left(2\frac{1}{r_{dsn}} + 2g_{mn} + \frac{1}{R_B}\right) = v_d\frac{1}{r_{dsn}} + v_o\frac{1}{r_{dsn}} + 2\hat{v}_i g_{mn}$$

$$v_o\left(\frac{1}{r_{dsn}} + \frac{1}{r_{dsp}}\right) + \hat{v}_i g_{mn} + v_d g_{mp} = v_s\left(g_{mn} + \frac{1}{r_{dsn}}\right)$$

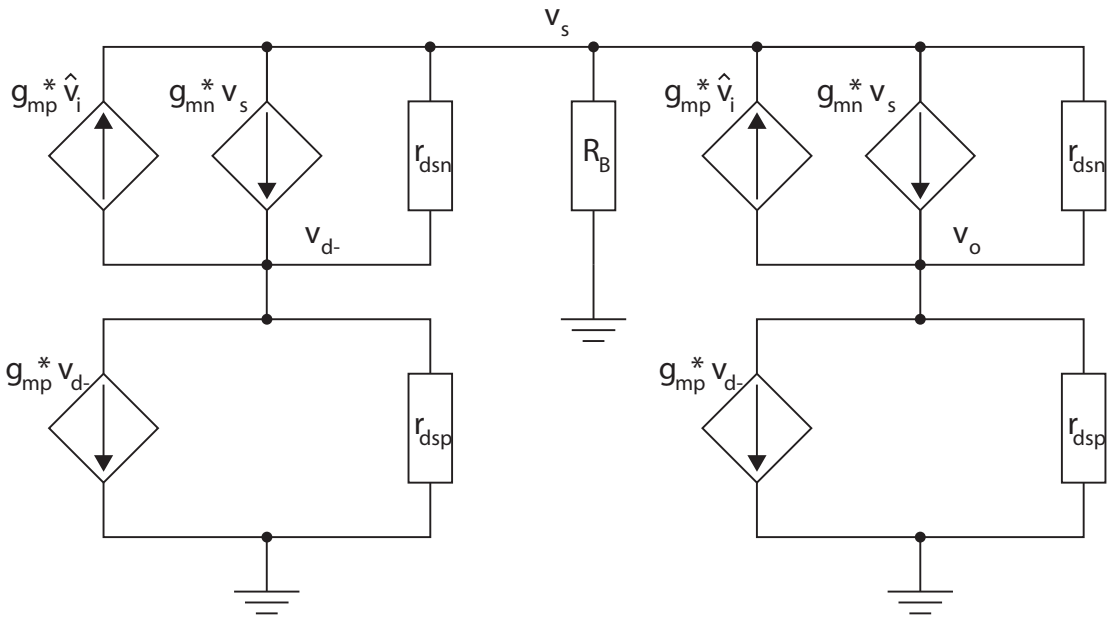
$\hat{v}_i$ small signal (2/3)



Result:

$$A_{\text{CM}} = \frac{v_o}{\hat{v}_i} = -\frac{\frac{g_{mn}}{R_B}}{2g_{mp}g_{mn} + 2\frac{g_{mp}}{r_{dsn}} + 2\frac{g_{mn}}{r_{dsp}} + \frac{g_{mp}}{R_B} + 2\frac{1}{r_{dsn}r_{dsp}} + \frac{1}{r_{dsn}R_B} + \frac{1}{r_{dsp}R_B}}$$

$\hat{v}_i$ small signal (3/3)

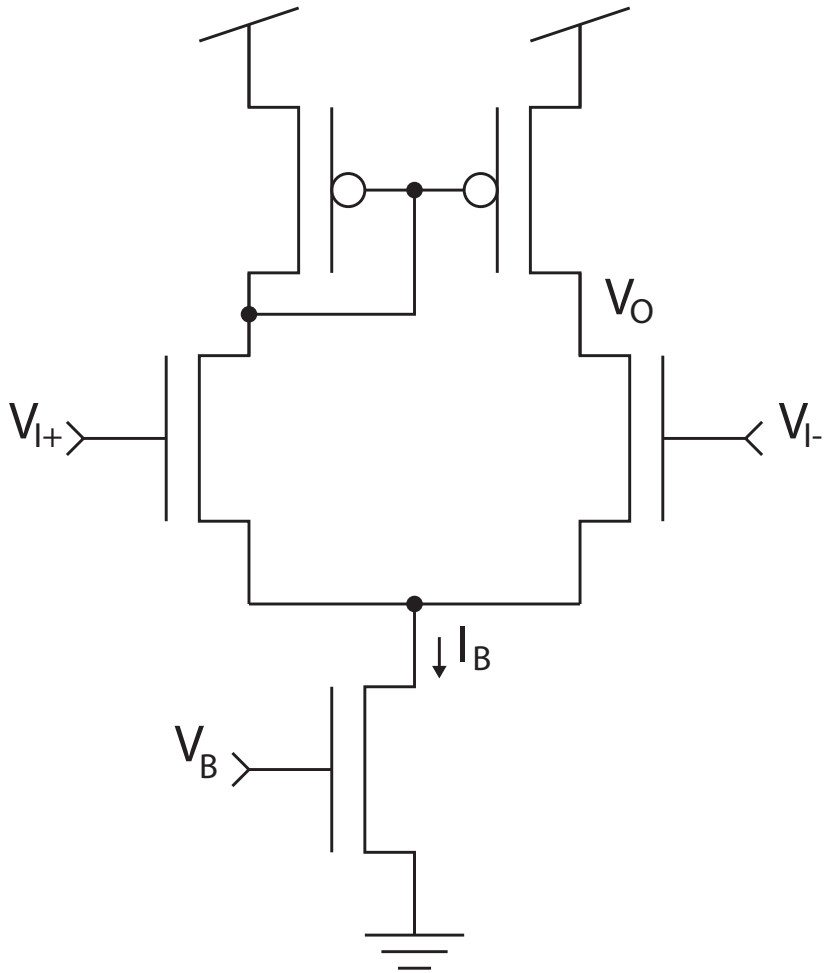


Same previous simplifying assumption:

$$\frac{v_o}{\hat{v}_i} \approx -\frac{1}{2R_B g_{mp}}$$

Systematic offset

(and an easier way to deduce the common mode gain A_{CM})



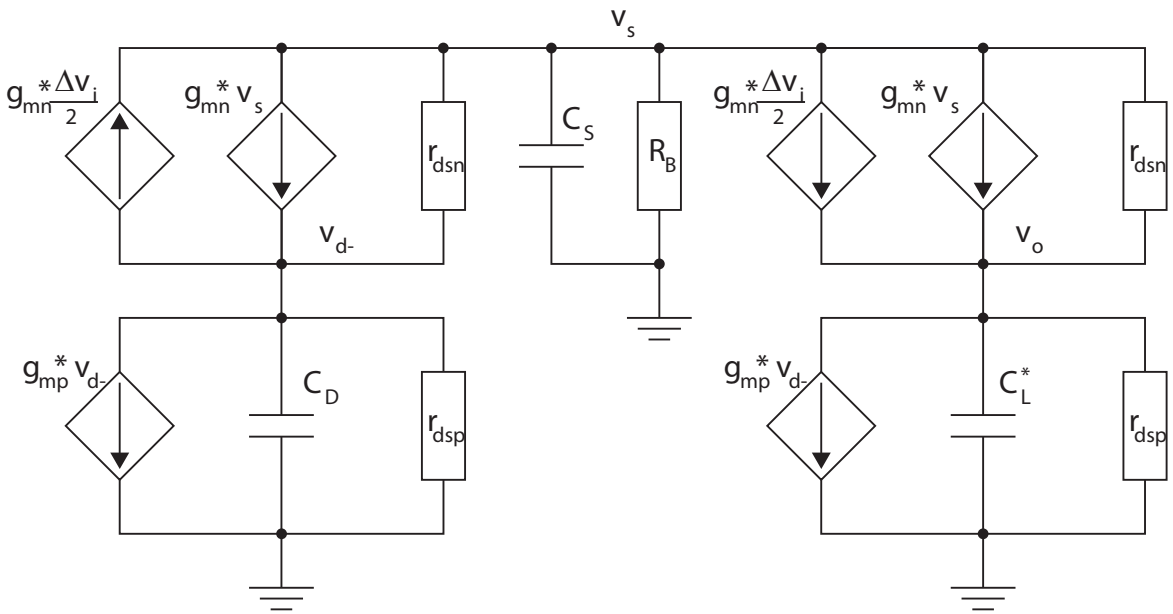
Common mode rejection ratio (CMRR)

Here:

$$A_{\mathrm{CM}} = \frac{v_o}{\hat{v}_i}$$

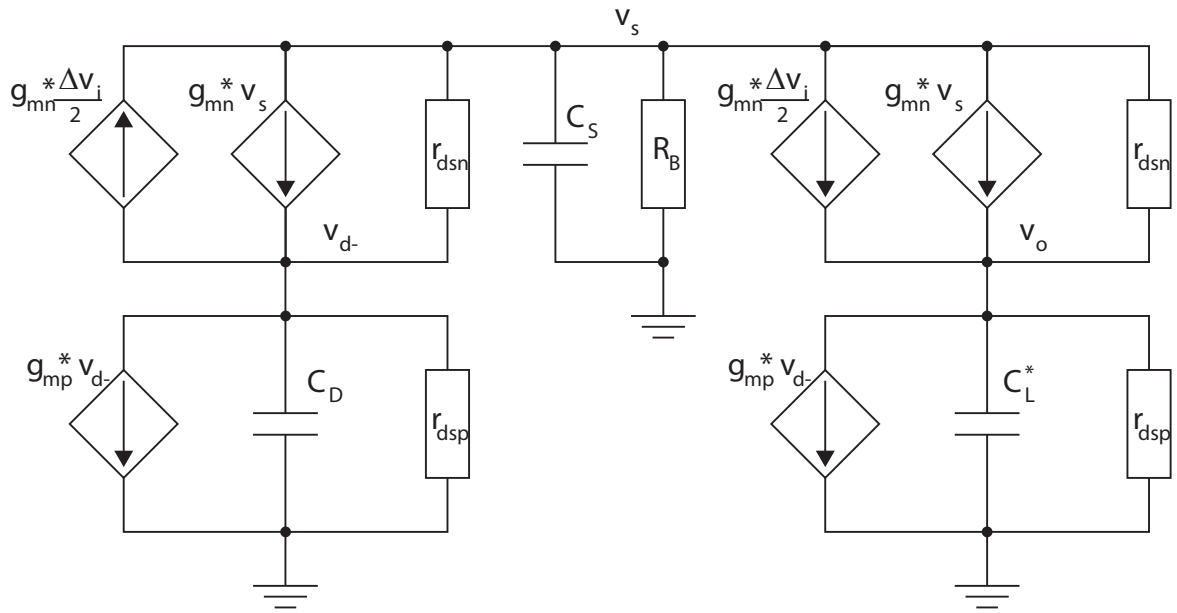
$$\mathrm{CMRR} \approx g_{mn} \left(r_{dsn} || r_{dsp} \right) 2 g_{mp} R_B$$

Frequency response: small signal



Neglecting some terms with capacitors C_{GD} of the input transistors and the current mirror output transistor.

Frequency response: Kirchhoff



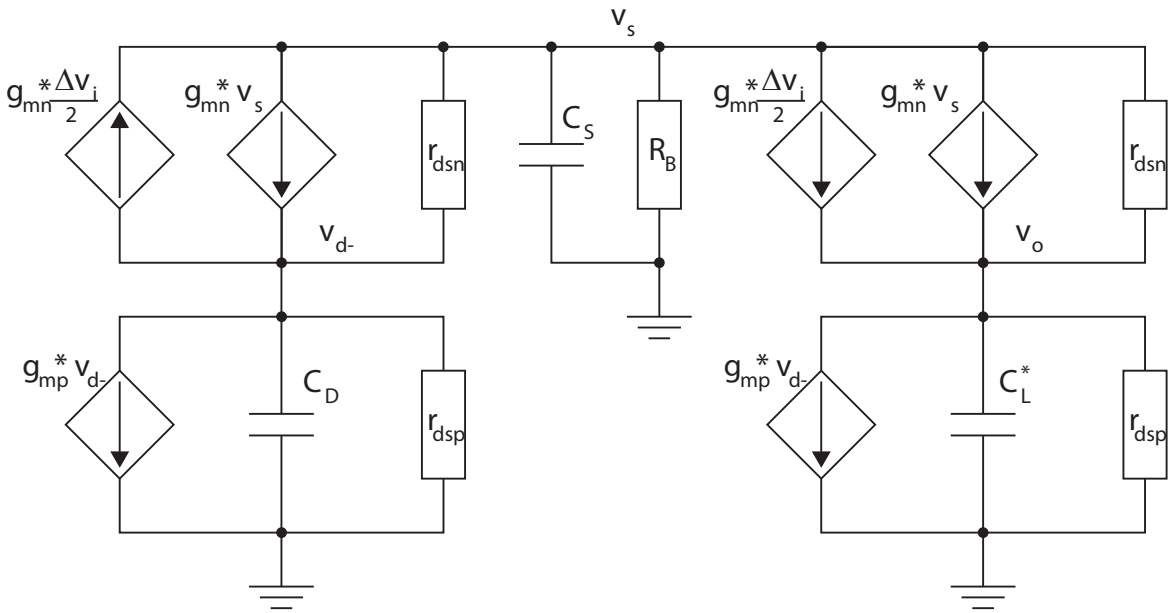
Kirchhoff:

$$v_d\left(\frac{1}{r_{dsn}}+\frac{1}{r_{dsp}}+g_{mp}+sC_D\right)+\frac{\Delta v_i}{2}g_{mn}=v_s\left(g_{mn}+\frac{1}{r_{dsn}}\right)$$

$$v_s\left(2\frac{1}{r_{dsn}}+2g_{mn}+\frac{1}{R_B}+sC_S\right)=v_d\frac{1}{r_{dsn}}+v_o\frac{1}{r_{dsn}}$$

$$v_o\left(\frac{1}{r_{dsn}}+\frac{1}{r_{dsp}}+sC_L^*\right)+v_d g_{mp}=v_s\left(g_{mn}+\frac{1}{r_{dsn}}\right)+\frac{\Delta v_i}{2}g_{mn}$$

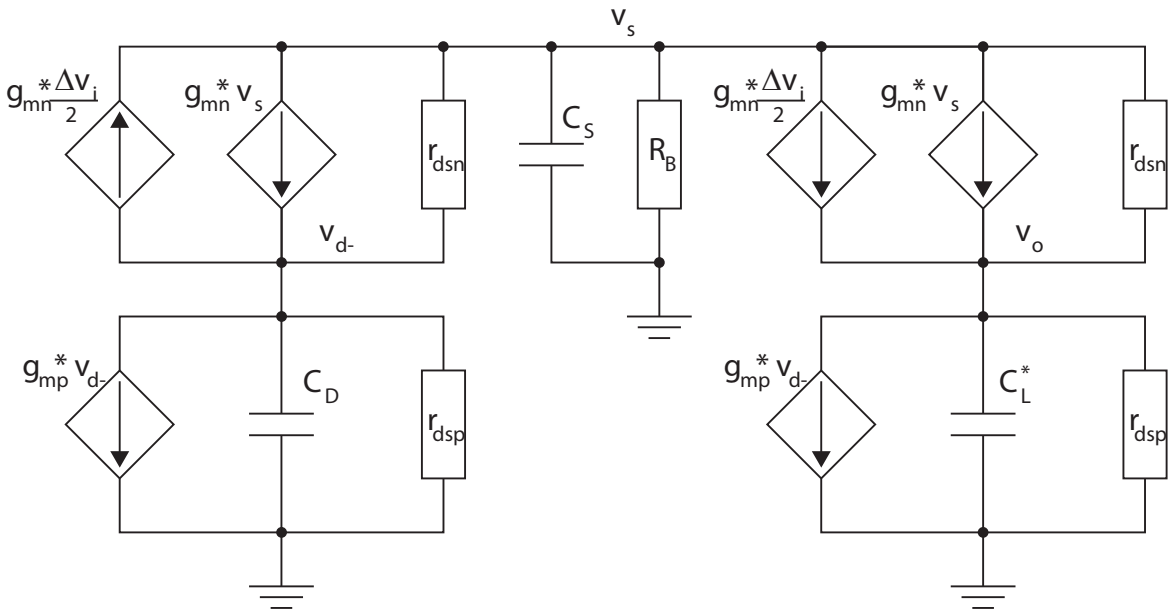
Frequency response: full solution



$$\frac{v_o}{\Delta v_i} = -G_M Z_O = g_{mn}(r_{dsn}||r_{dsp})\frac{1}{1+s(r_{dsn}||r_{dsp})C_L^*}$$

$$* \frac{\left(2g_{mp}+\frac{1}{r_{dsp}}+\frac{1}{r_{dsn}}+sC_D\right)\left(g_{mn}+\frac{1}{r_{dsn}}+\frac{1}{2R_B}+\frac{sC_S}{2}\right)-\left(g_{mn}+\frac{1}{r_{dsn}}\right)\frac{1}{r_{dsn}}}{\left(g_{mp}+\frac{1}{r_{dsp}}+\frac{1}{r_{dsn}}+sC_D\right)\left(2g_{mn}+2\frac{1}{r_{dsn}}+\frac{1}{R_B}+sC_S\right)-2\left(g_{mn}+\frac{1}{r_{dsn}}\right)\frac{1}{r_{dsn}}}$$

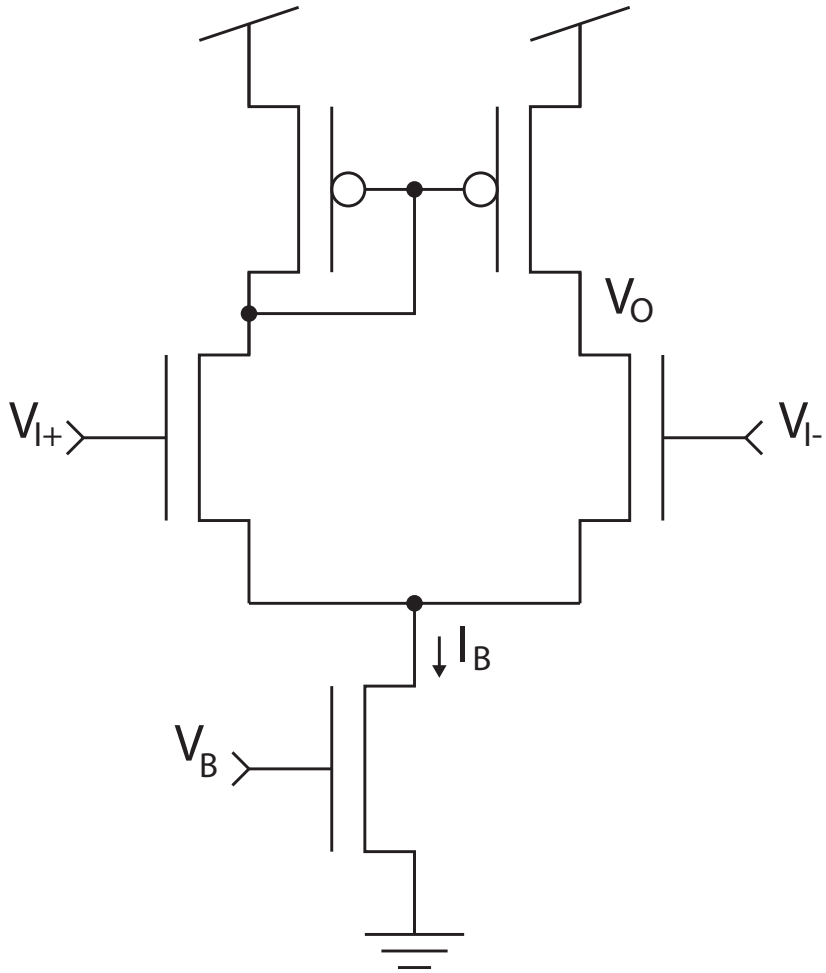
Frequency response: simplified solution



$$\frac{v_o}{\Delta v_i} = -G_M Z_O \approx g_{mn}(r_{dsn}||r_{dsp}) \frac{1}{1+s(r_{dsn}||r_{dsp})C_L^*} \frac{\left(1+s\frac{C_D}{2g_{mp}}\right)}{\left(1+s\frac{C_D}{g_{mp}}\right)}$$

Slew rate (1/2)

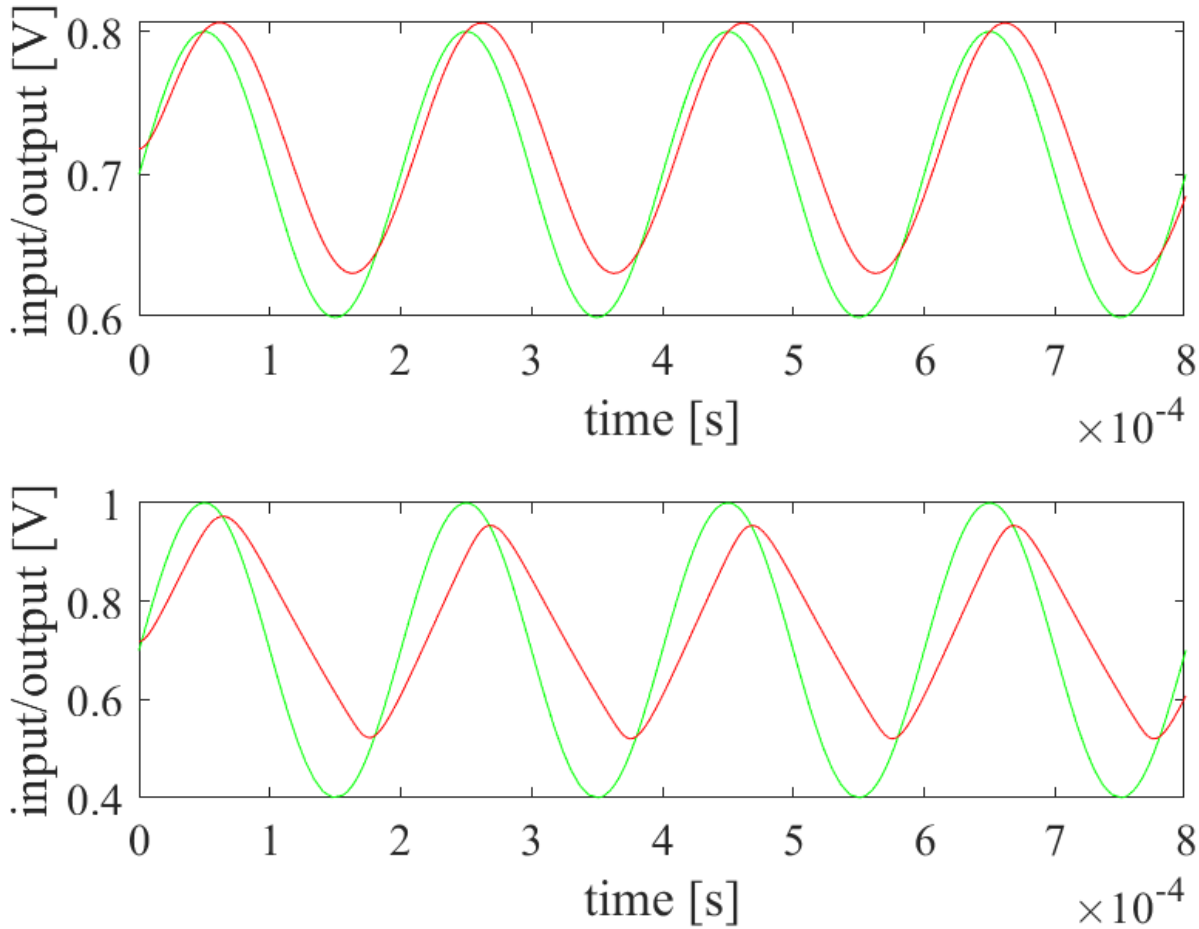
Maximum rate of change of output voltage.



Maximum output current is I_B, i.e. when input voltage difference is BIG and the small signal approximation does no longer hold.

Slew rate (2/2)

Can for example be problematic if transamp connected as follower.



Oftentimes though, it is due to an internal node in multi stage amplifiers and has nothing to do with the output capacitive load.

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