EE210: Microelectronics-I

Lecture-41 : MOS Amplifiers_1

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dc model parameters

$$\text{Linear}: \; I_{DS} = \beta_N \left\{ (\mathrm{V_{GS}} - V_{THN}) V_{DS} - \frac{{V_{DS}}^2}{2} \right\}$$

$$\text{Saturation}: \; I_{DS} = \frac{\beta_N}{2} (V_{GS} - V_{THN})^2 [1 + \lambda_n V_{DS}]$$

$$\beta_N = kP_N . \frac{W}{L}$$

$$\lambda_N$$

$$V_{THN} = V_{THN0} + \gamma \left(\sqrt{2\phi_F + V_{SB}} - \sqrt{2\phi_F} \right)$$

$$V_{THNO} = 1V; \gamma = 0.7\ V^{1/2}; 2\phi_F = 0.7V;$$

$$KP_N = 100\mu A / V^2; L = 1\mu m; \lambda = 0.01V^{-1}$$

L is usually fixed, W is determined by designer

Saturation

$$I_{D,sat} = \frac{K_n'}{2}\frac{W}{L}(V_{GS}-V_{TN})^2(1+\lambda V_{DS})$$

$$= \frac{K_n}{2}(V_{GS}-V_{TN})^2(1+\lambda V_{DS})$$

$$\frac{\partial I_{D,sat}}{\partial V_{DS}} = \lambda I_{D,sat} = \frac{1}{r_o} = g_{ds}$$

$$g_m = \left.\frac{\partial I_{DS}}{\partial V_{GS}}\right|_{V_{DS}\ \&\ V_{SB}\text{ constant}}$$

$$\Rightarrow \quad g_m = K_n(V_{GS}-V_T)(1+\lambda V_{DS}) \simeq K_n(V_{GS}-V_T)$$

$$= \sqrt{2K_n I_{DS}}$$

$$V_{TN} = V_{TN0} + \gamma\left(\sqrt{2\phi_F - V_{BS}}\right)$$

$$\frac{\partial V_{TN}}{\partial V_{BS}} = \gamma \cdot \frac{-1}{2\sqrt{2\phi_F - V_{BS}}}$$

$$g_{mb} = \left.\frac{\partial I_{DS}}{\partial V_{BS}}\right|_{V_{GS}\ \&\ V_{DS}\text{ constant}}$$

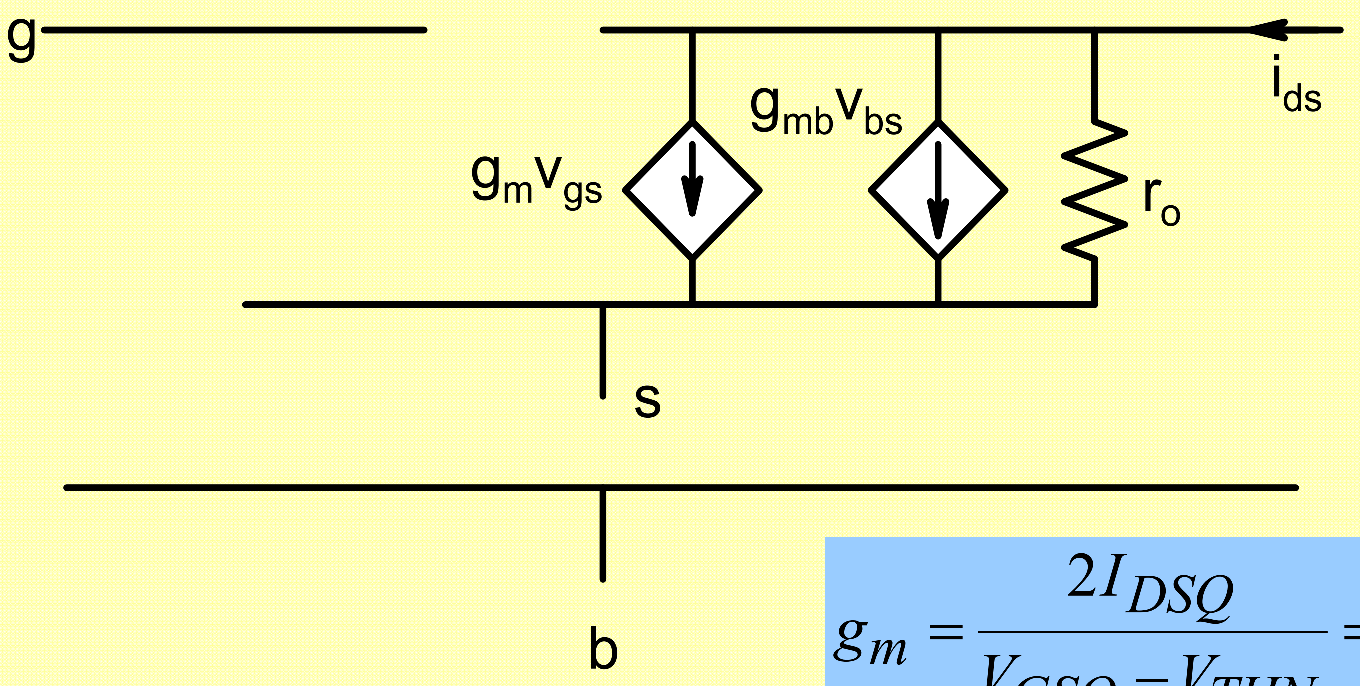
$$\Rightarrow \quad g_{mb} = \frac{K_n}{2}\left[2(V_{GS}-V_{TN})\cdot\left(-\frac{\partial V_{TN}}{\partial V_{BS}}\right)\right]$$

$$\Rightarrow \quad g_{mb} = g_m \times \frac{\gamma}{2\sqrt{2\phi_F - V_{BS}}} = g_m \cdot \eta \text{ or } = g_m \chi$$

where η or $\chi = \dfrac{\gamma}{2\sqrt{2\phi_F - V_{BS}}} = \dfrac{\gamma}{2\sqrt{2\phi_F + V_{SB}}}$

Low frequency Small Signal model

$$i_{ds} = g_m v_{gs} + g_{mb} v_{bs} + \frac{v_{ds}}{r_o}$$



$$g_m = \frac{2I_{DSQ}}{V_{GSQ} - V_{THN}} = \sqrt{2I_{DSQ}\beta}$$

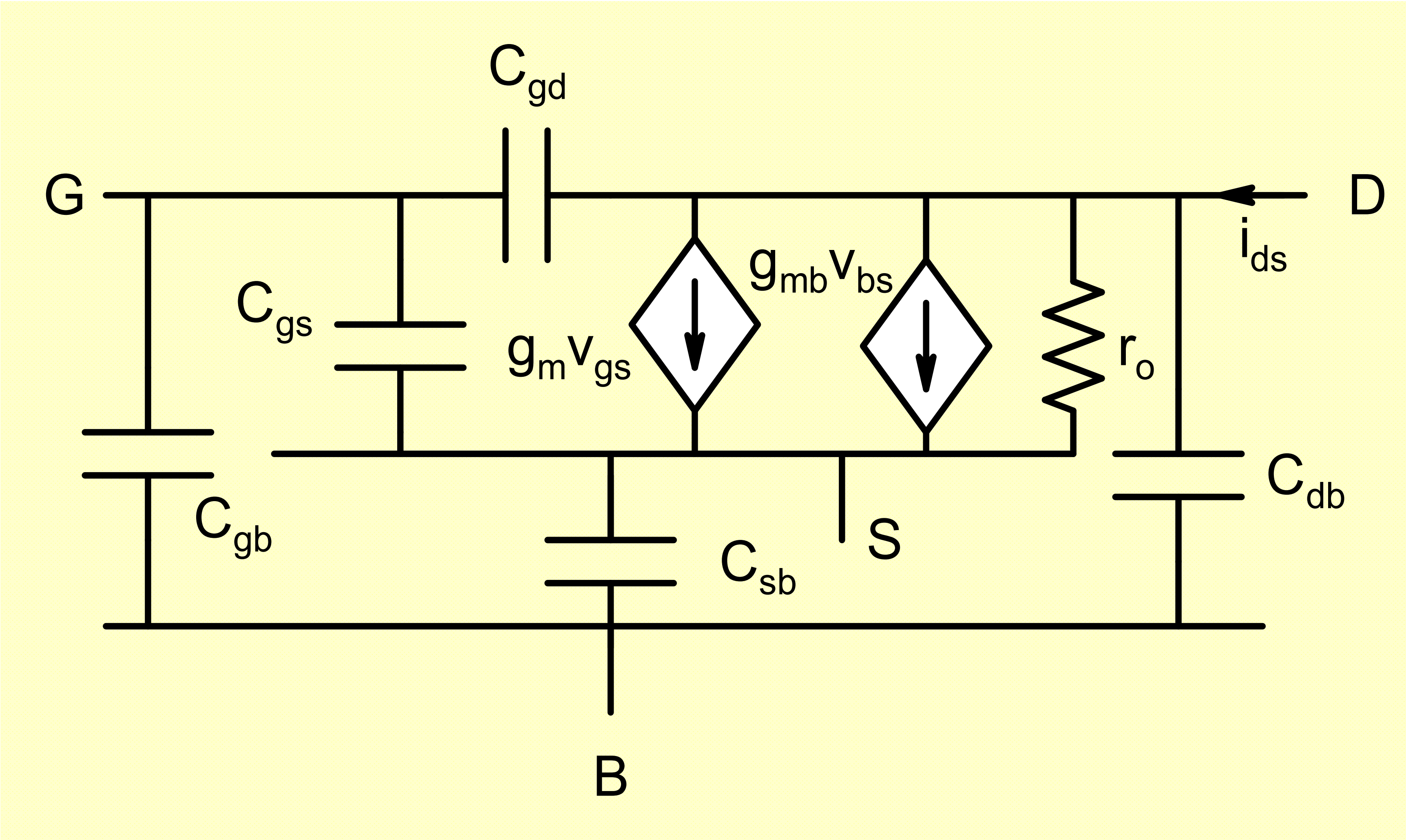
$$r_o = \frac{1}{\lambda_n I_{DSQ}}$$

$$g_{mb} = g_m . \eta$$

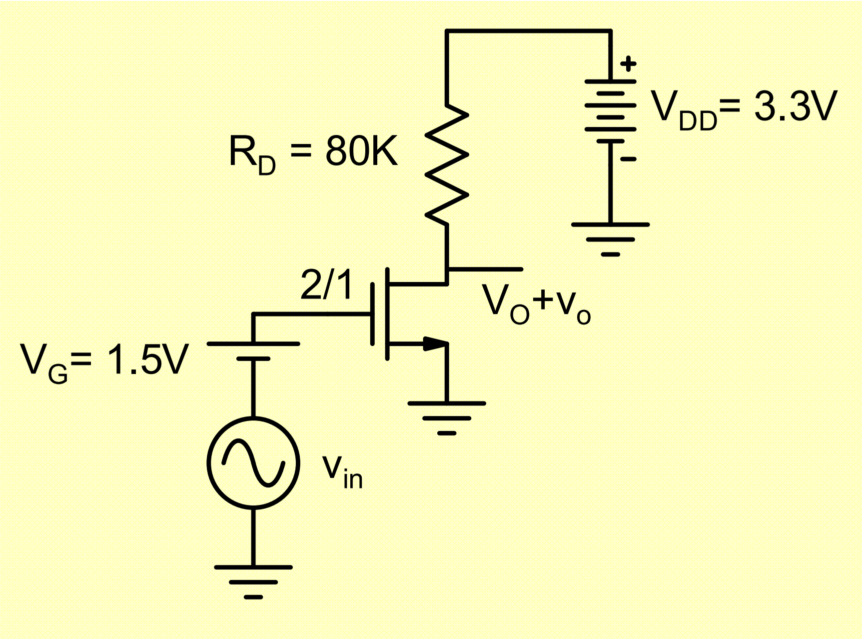
$$\eta = \frac{\gamma}{2\sqrt{2\Phi_F + V_{SBQ}}}$$

G-Number

High Frequency Small Signal Model



Common Source Amplifier with Resistive load



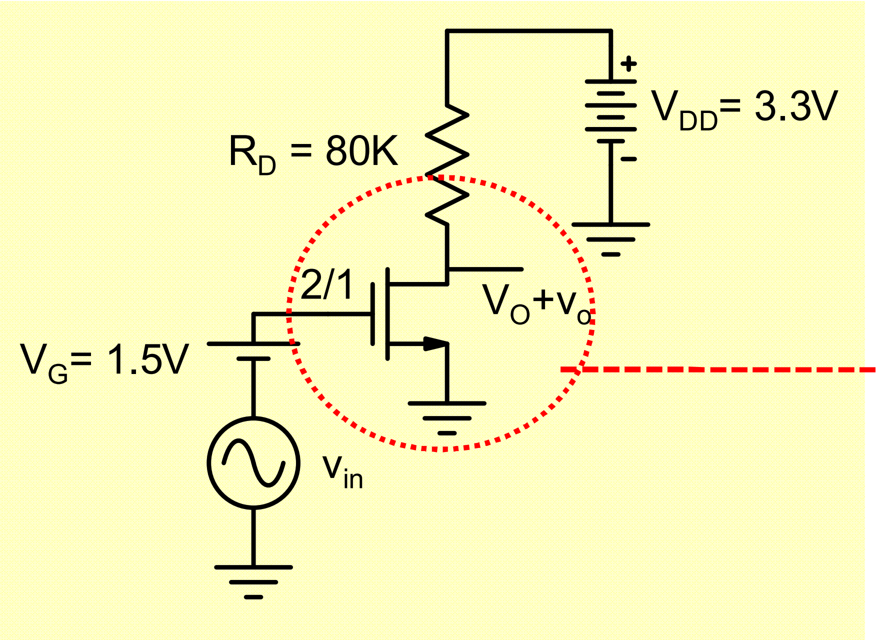
dc analysis

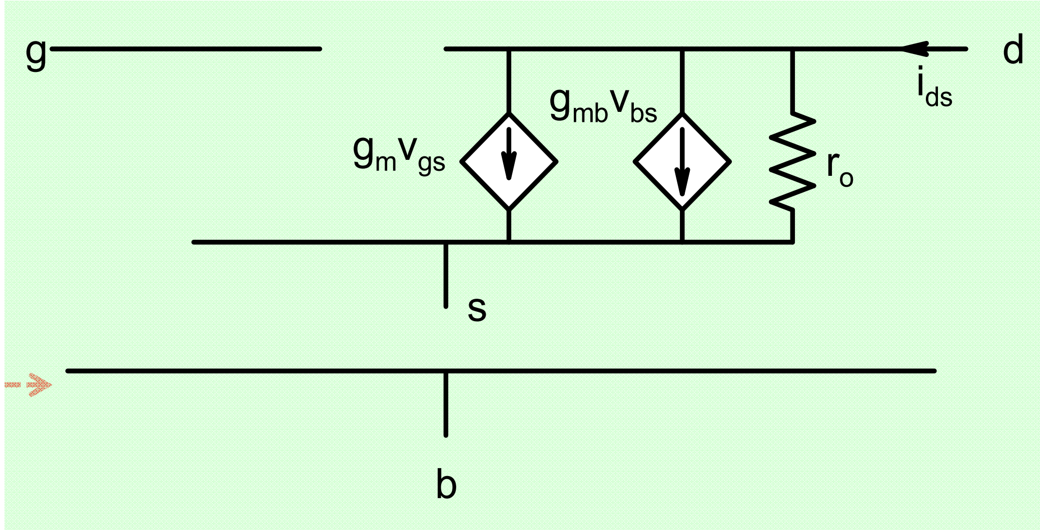
$$I_{DSQ} = \frac{\beta_n}{2}(V_{GS} - V_{TN})^2 = 25\mu A$$

$$V_{DSQ} = V_{DD} - I_{DSQ} \times R_D = 1.3V$$

$$V_{sat} = V_{GSQ} - V_{TN} = 0.51V \text{ so Tr. is in Saturation}$$

Small Signal Model





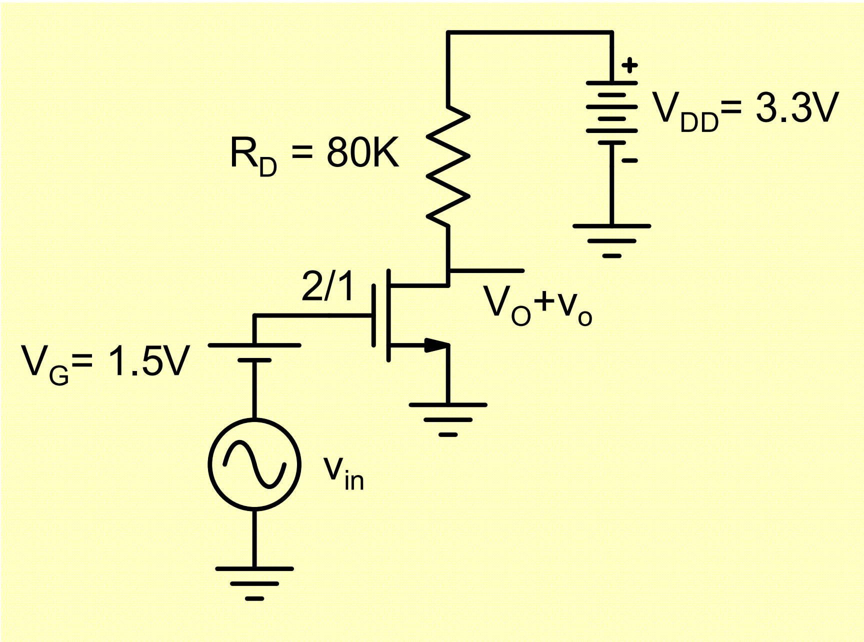
$$g_m = \frac{2I_{DSQ}}{V_{GSQ} - V_{THN}} = \sqrt{2I_{DSQ}\beta} = 100\mu A/V$$

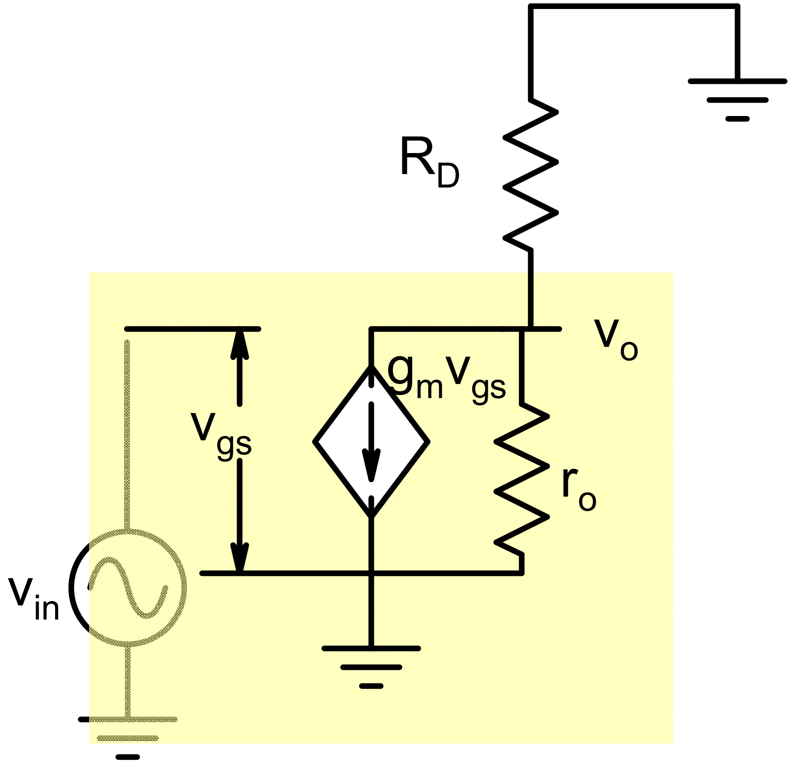
$$r_o = \frac{1}{\lambda_n I_{DSQ}} = 4M\Omega$$

$$g_{mb} = g_m.\eta = 41.83\mu A/V$$

$$\eta = \frac{\gamma}{2\sqrt{2\Phi_F + V_{SBQ}}}$$

1. Voltage Gain

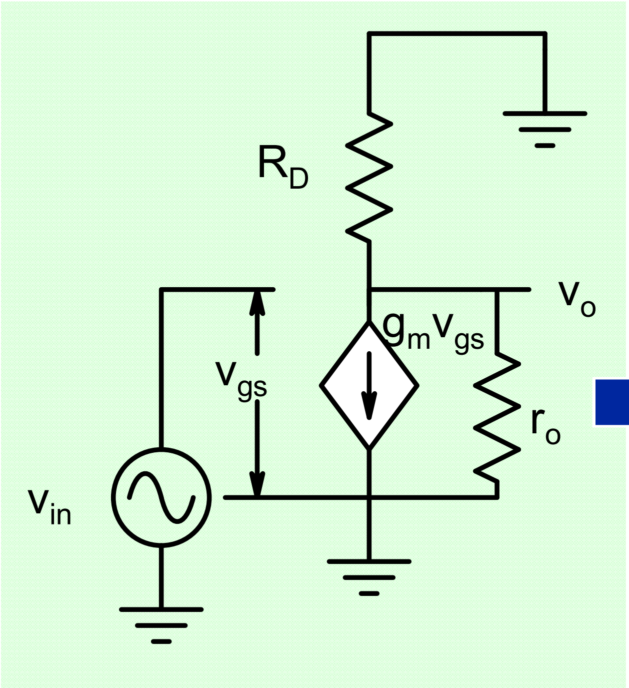


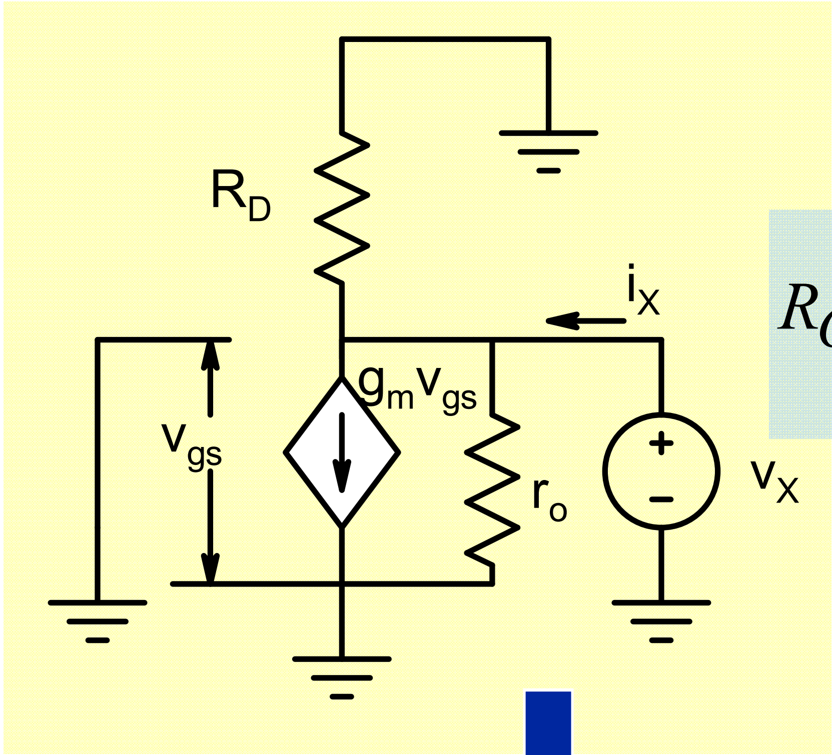


$$v_o = -g_m v_{in} \times R_D \| r_o$$

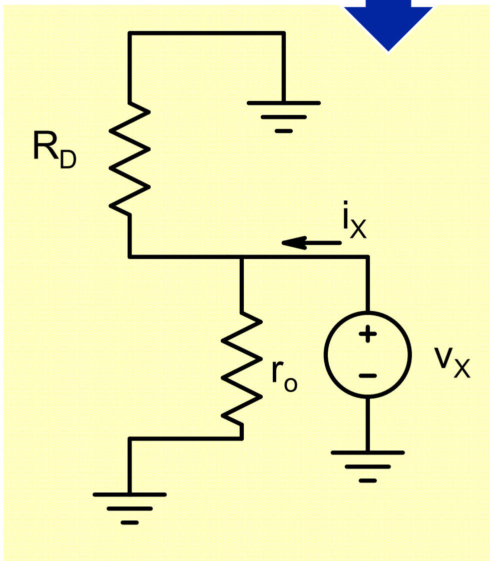
$$A_V = \frac{v_o}{v_{in}} = -g_m \times R_D \| r_o \cong -g_m R_D = -8$$

2. Output Resistance



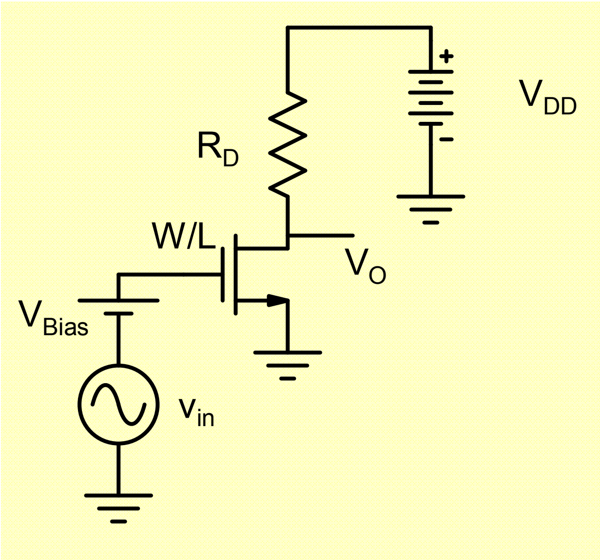


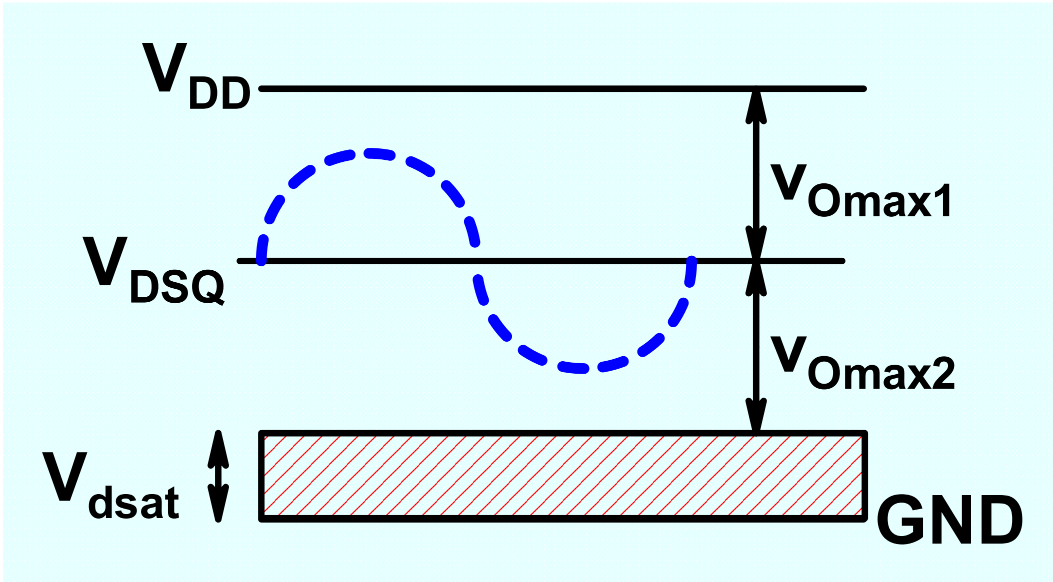
$$R_O = \left.\frac{v_x}{i_x}\right|_{v_{in}=0}$$



$$R_O = R_D \| r_o \cong R_D = 80K$$

3. Output Voltage Swing





$$v_{o\,\max 1} = V_{DD} - V_{DSQ} \qquad v_{o\,\max 2} = V_{DSQ} - V_{dsat}$$

$$v_{o\,\max} = Min\{v_{o\,\max 1}; v_{o\,\max 2}\}$$

Voltage swing limited by harmonic distortion

- Harmonic distortion in CS amplifier occurs because the relationship between drain current and gate voltage is nonlinear.

$$I_{DSQ} + i_{ds} = \frac{\beta}{2} \times (V_{GSQ} + v_{gs} - V_T)^2 \quad i_{ds} = g_m v_{gs} + \left(\frac{0.5}{V_{GSQ} - V_T}\right) \times g_m v_{gs}^2$$

$$v_{gs} = a_o \sin(2\pi f_o t)$$

$$i_{ds} = g_m a_o \sin(2\pi f_o t) + \left(\frac{a_o^2 g_m}{4V_{GSQ} - V_T}\right) - \left(\frac{a_o^2 g_m}{4(V_{GSQ} - V_T)}\right) \cos(2\pi 2 f_o t)$$

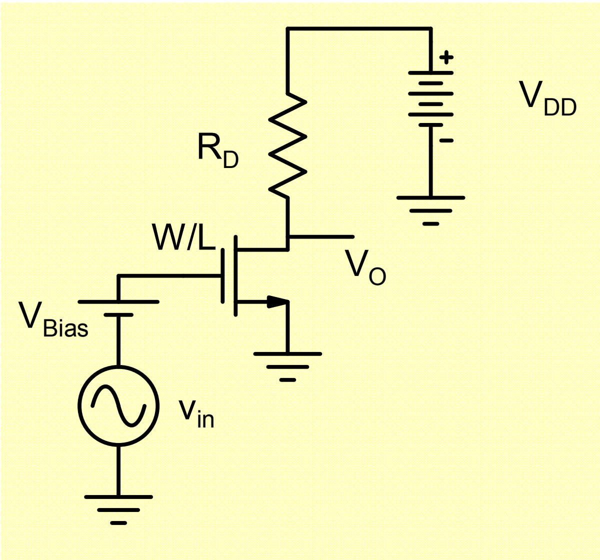
$$HD_2(\%) = \frac{a_o / 4}{V_{GSQ} - V_T} \times 100$$

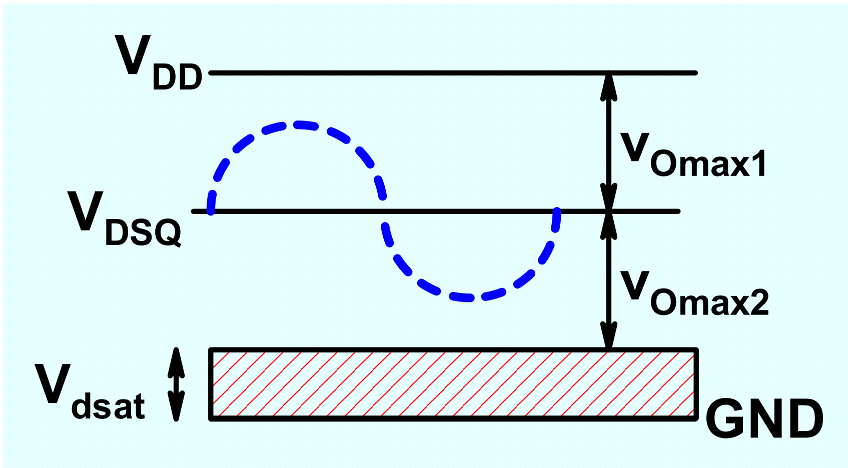
Output Voltage Swing

$$v_{in} = a_o \sin(2\pi f_o t) \qquad HD_2(\%) = \frac{a_o/4}{V_{GSQ} - V_T} \times 100 = \frac{v_{in}}{V_{dsat}} \times 25$$

$$v_{in} = \frac{HD_2}{25} \times V_{dsat} \qquad v_o = A_v \times v_{in} \cong g_m R_D \parallel R_L \times v_{in} \qquad g_m = \frac{2I_{DSQ}}{V_{GSQ} - V_T}$$

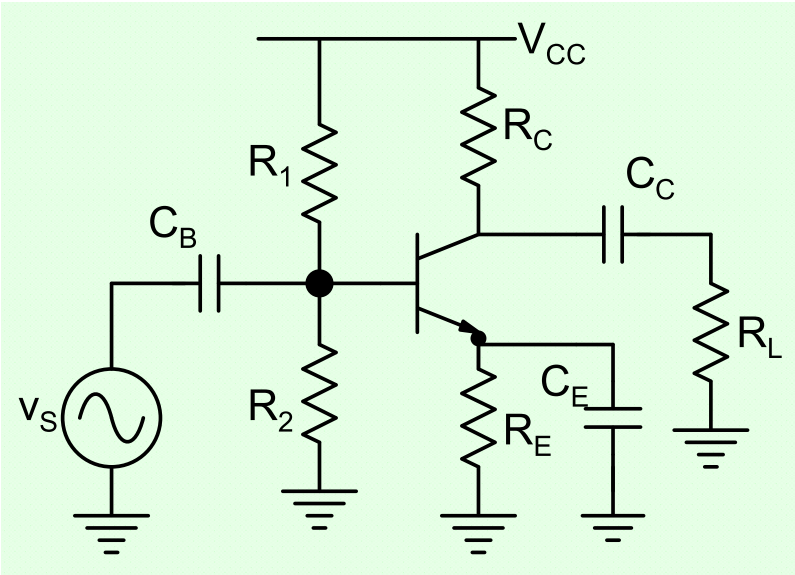
$$v_{o\max 3} \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5}$$





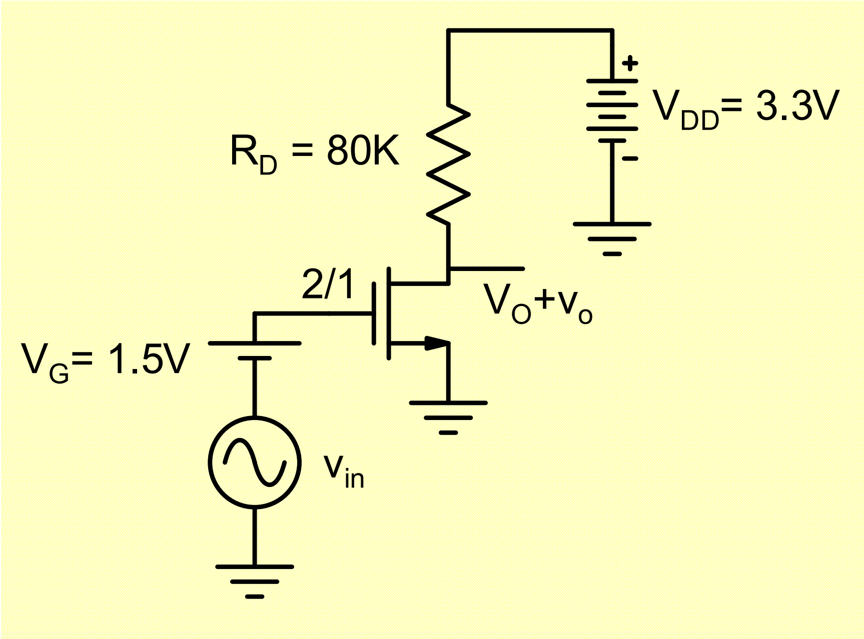
$$v_{o\max 1} = V_{DD} - V_{DSQ} = V_{DD} - I_{DSQ} R_D \qquad v_{o\max 3} < v_{o\max 1}$$

CE Amplifier: Limits on Output voltage swing



$$v_{om} < V_{CEQ} - 0.2$$

$$v_{om} \le (I_{CQ} R_C \| R_L) \times \frac{HD_2(\%)}{25}$$

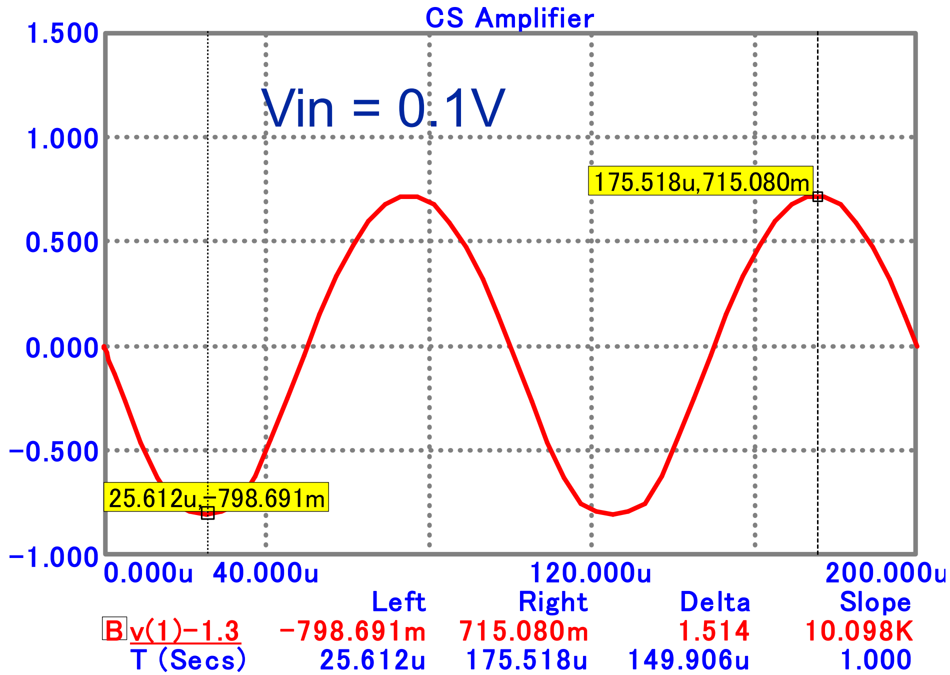


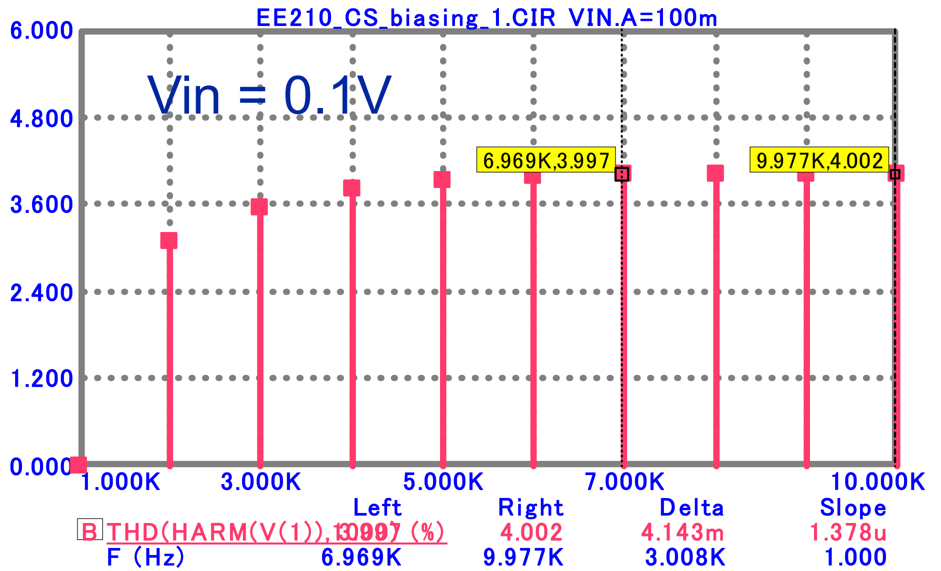
$$V_{DSQ} = 1.3V ; V_{sat} = 0.5V$$

$$v_{o\max} \le V_{DSQ} - V_{sat} = 0.8V$$

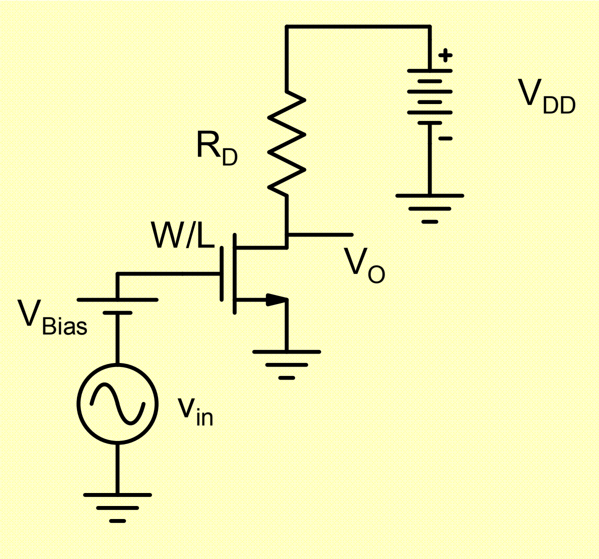
$$v_{o\max 3} \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5}$$

$$= 0.8V \text{ for } \mathrm{HD}_2 = 5\%$$





Distortion is a little smaller due to clipping



$$v_o \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5}$$

$$v_o \le V_{DSQ} - V_{sat}$$

Optimum drain-source bias

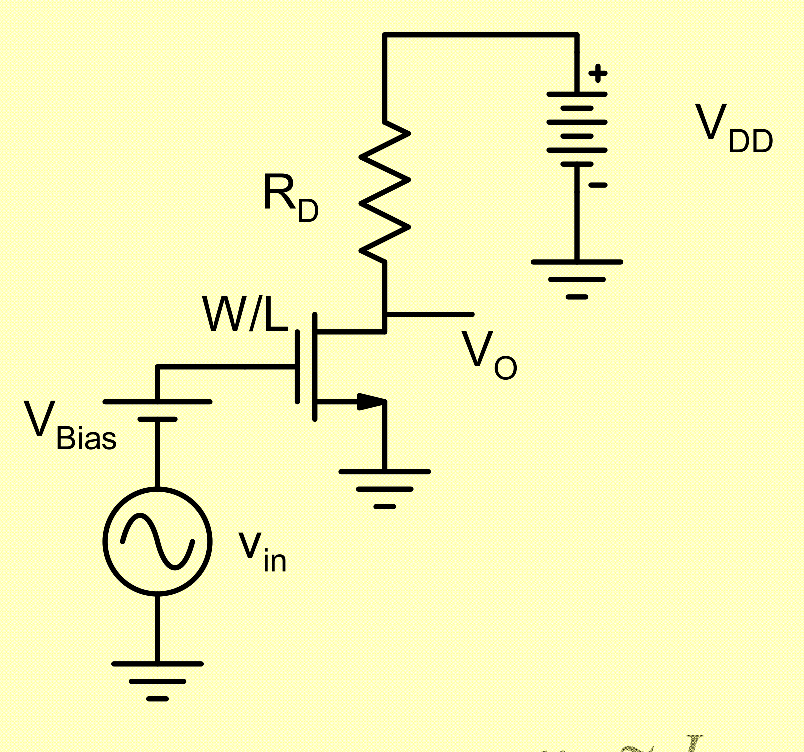
$$I_{DSQ} R_D \times \frac{HD_2}{12.5} \times \frac{1}{1 + R_D / R_L} = V_{DSQ} - V_{sat}$$

$$V_{DSQ} = \frac{V_{DD} \times \dfrac{HD_2}{\chi} + V_{sat}}{1 + \dfrac{HD_2}{\chi}}$$

$$v_{o\,\max} = \frac{V_{DD} - V_{sat}}{1 + \dfrac{\chi}{HD_2}}$$

$$\chi = 12.5 \times (1 + R_D / R_L)$$

Voltage Gain



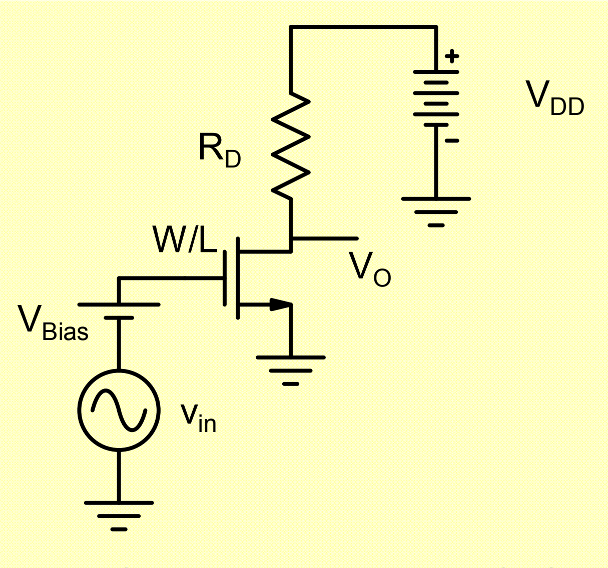
$$A_V = -g_m R_D \qquad g_m = \sqrt{2 I_{DSQ} \beta}$$

$$R_D = \frac{V_{DD} - V_{DSQ}}{I_{DSQ}}$$

$$A_v = \sqrt{\frac{2\beta}{I_{DSQ}}} \times (V_{DD} - V_{DSQ})$$

$$v_o \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5} \qquad v_o \le V_{DSQ} - V_{sat}$$

$$A_v \le \sqrt{\frac{2\beta}{I_{DSQ}}} \times (V_{DD} - v_{om} - \sqrt{\frac{2 I_{DSQ}}{\beta}})$$



$$A_v \le \sqrt{\frac{2\beta}{I_{DSQ}}} \times (V_{DD} - v_{om} - \sqrt{\frac{2I_{DSQ}}{\beta}})$$

W=2μm and v_{om} = 0.1V

I_{DS} (μA)	AV	RD (MΩ)	V_{GS} (V)
100	4.4	0.022	2
50	7.05	0.05	1.707
25	10.8	0.108	1.5
10	18.2	0.29	1.316
5	26.6	0.59	1.224
2	43.2	1.53	1.14
1	62	3.1	1.1

I_{DS}=25μA and v_{om} = 0.1V, R_D =108k

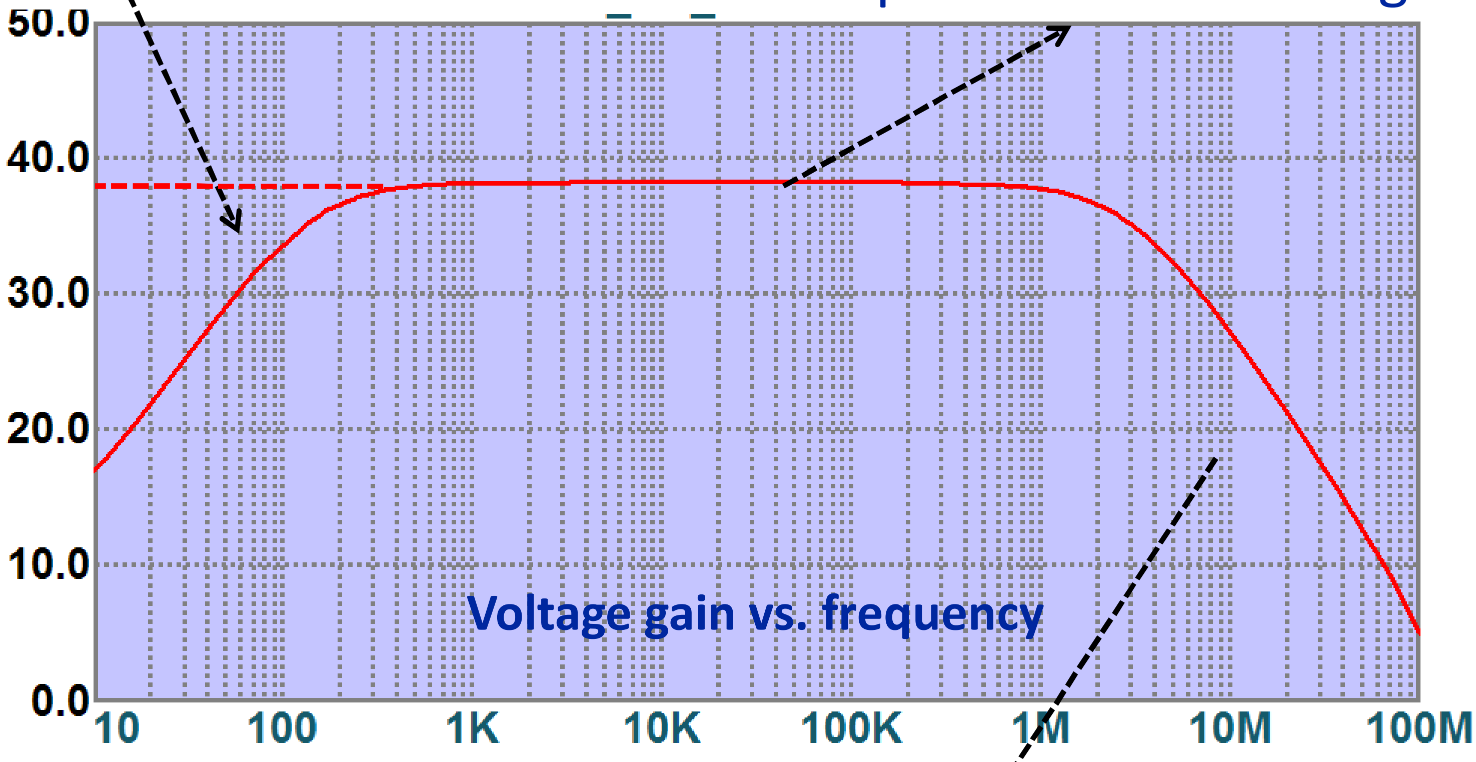
W (μm)	AV	V_{GS} (V)
2	10.8	1.5
5	18.24	1.316
10	26.6	1.22
20	38.47	1.158
50	62	1.1
100	88.5	1.07

- Low current or large size is needed to obtain large gain.
- Gain saturates as transistor gets closer to threshold voltage

4. Frequency Response

Low frequency behavior is caused by external capacitances

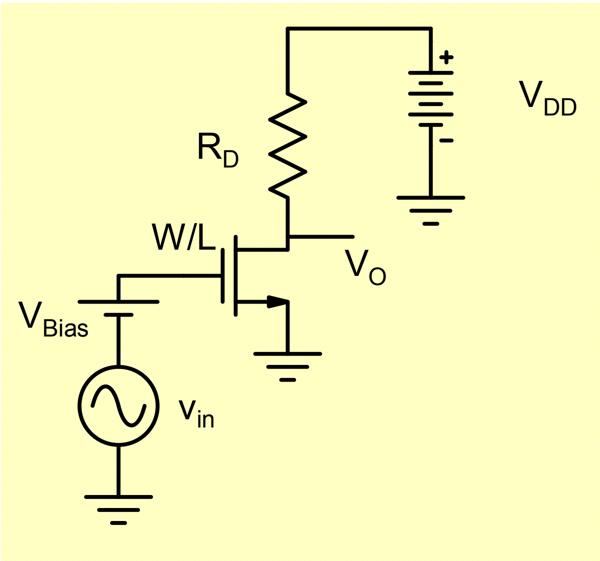
Mid-frequency region: all capacitances can be ignored

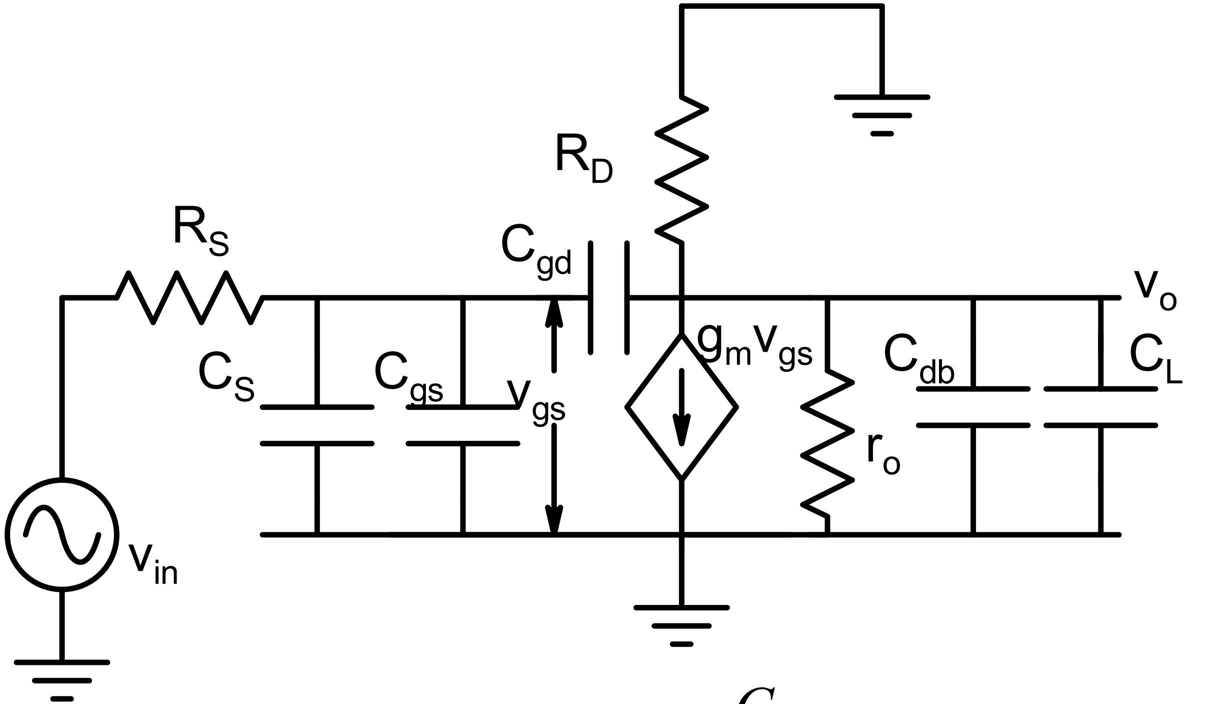


High frequency behavior is caused by internal transistor capacitances

G-Number

3-dB Upper Cutoff Frequency





•The general method for determining frequency response of an amplifier is to carry out analysis of the circuit to obtain the transfer function and then obtain the 3dB frequency there from.

$$A(s) = -g_m R_D \frac{1 - s\dfrac{C_{gd}}{g_m}}{1 + sg + s^2 h}$$

$$g = R_S \times (C_{gs} + C_{gd}(1 + g_m R_D)) + R_D \times (C_{gd} + C_{db})$$

$$h = R_S R_D \times \{C_{gd} + C_{db}) \times C_{gs} + C_{gd} C_{db}\}$$

• The determination of poles from the transfer function becomes easier if one of the poles (say p_1) is dominant (much lower frequency) than the other pole p_2.

$$A(s) = -g_m R_D \frac{1 - s\dfrac{C_{gd}}{g_m}}{1 + sg + s^2 h}$$

$$(1 + gs + hs^2) = \left(1 - \frac{s}{p_1}\right)\left(1 - \frac{s}{p_2}\right) = 1 - s\left(\frac{1}{p_1} + \frac{1}{p_2}\right) + \frac{s^2}{p_1 p_2} \cong 1 - \frac{s}{p_1} + \frac{s^2}{p_1 p_2}$$

$$\Rightarrow \quad p_1 \cong -\frac{1}{g} \qquad p_1 p_2 \cong h^{-1}; \quad \Rightarrow p_2 \cong -\frac{g}{h} \qquad \omega_{3dB} \cong -p_1$$

$$f_{3dB} \cong \frac{1}{2\pi} \times \frac{1}{R_S(C_{gs} + C_{gd}(1 + g_m R_D)) + R_D(C_{gd} + C_{db})}$$

$$\omega_{p2} \cong \frac{g_m C_{gd}}{C_{gs}C_{gd} + C_{gs}C_{db} + C_{gd}C_{db}}$$

Open Circuit Time Constant Approach

• A simpler technique which gives approximate answer and is also based on dominant pole approximation is :

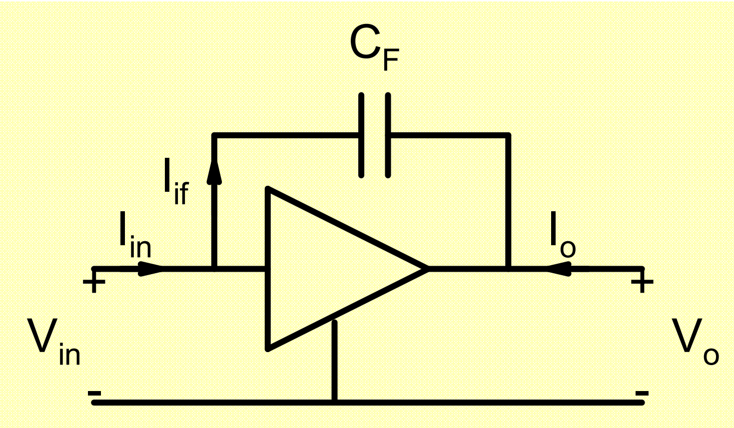
$$f_{3dB} = \frac{1}{2\pi \sum \tau_j}$$

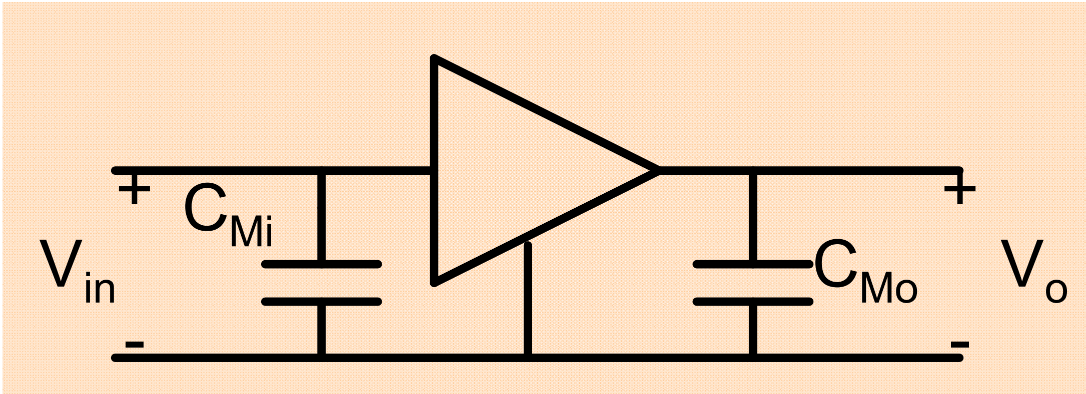
τ_j is the time constant associated with capacitor C_j

$$\tau_j = R_j C_j$$

where R_j is the effective resistance seen by the capacitor when all other capacitors are removed from the circuit

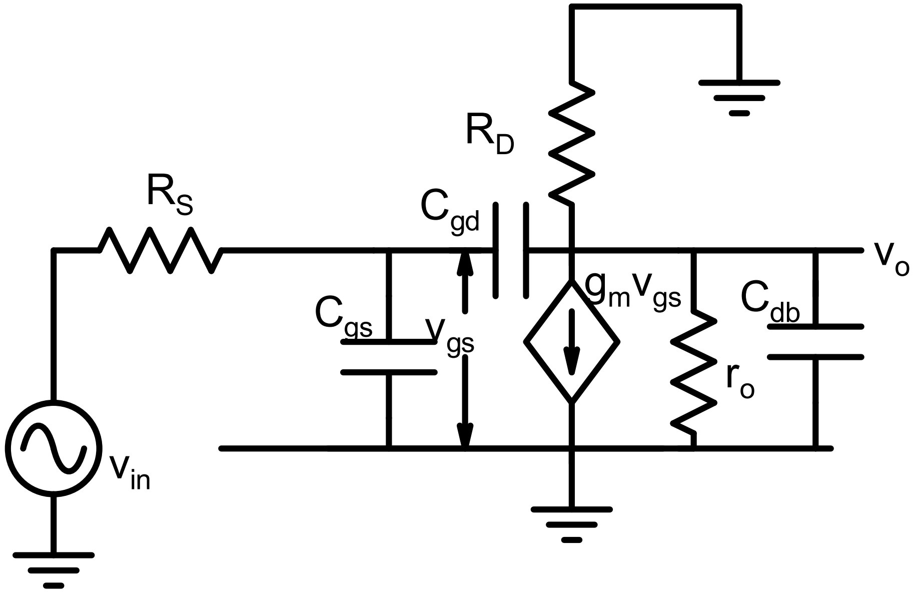
Millers Theorem





$$C_{Mi} = C_F \times (1 - A_V)$$

$$C_{Mo} = C_F \times (1 - \frac{1}{A_V})$$



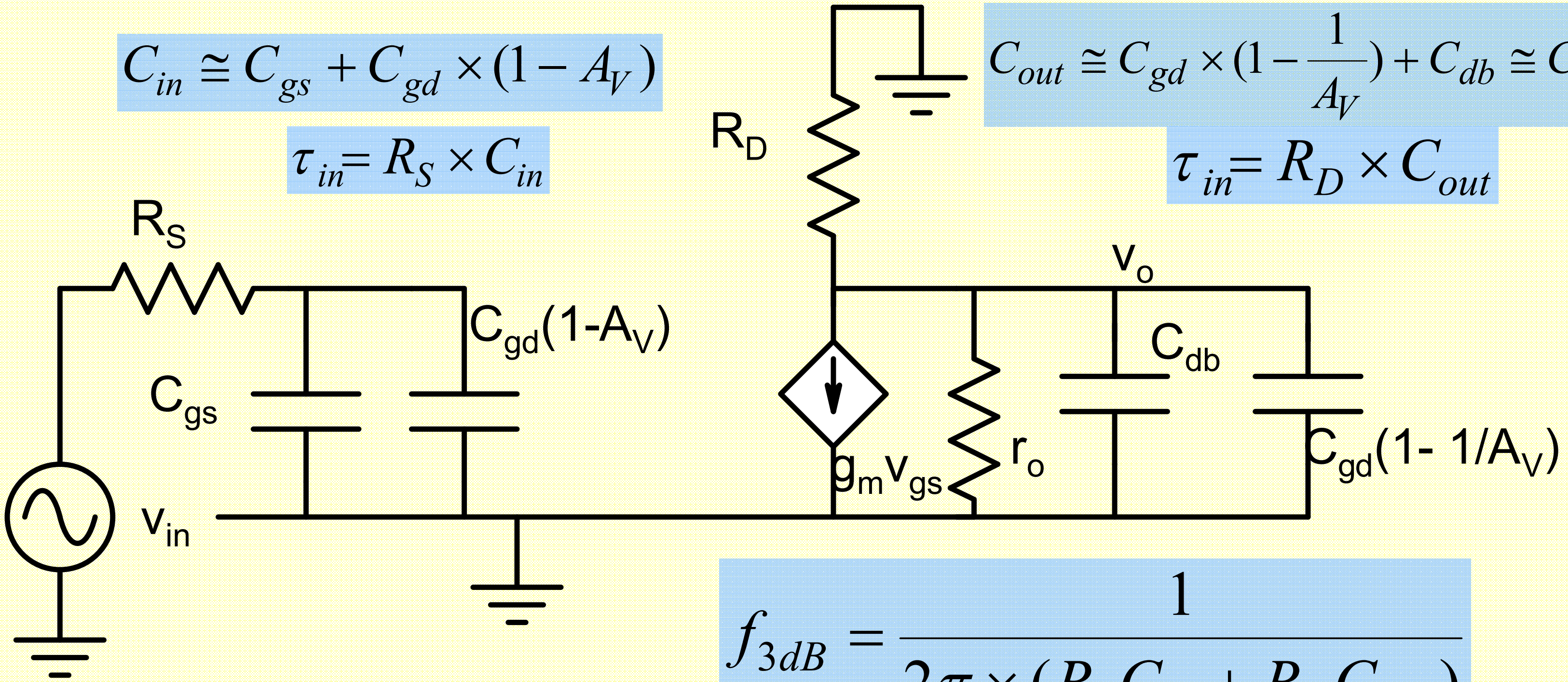
•The estimation of time constants becomes simpler through use of Miller's theorem which allows the capacitance C_{gd} to be split into two capacitances, one at the input and the other at the output.

$$C_{in} \cong C_{gs} + C_{gd} \times (1 - A_V)$$

$$\tau_{in} = R_S \times C_{in}$$

$$C_{out} \cong C_{gd} \times (1 - \frac{1}{A_V}) + C_{db} \cong C_{gd} + C_{db}$$

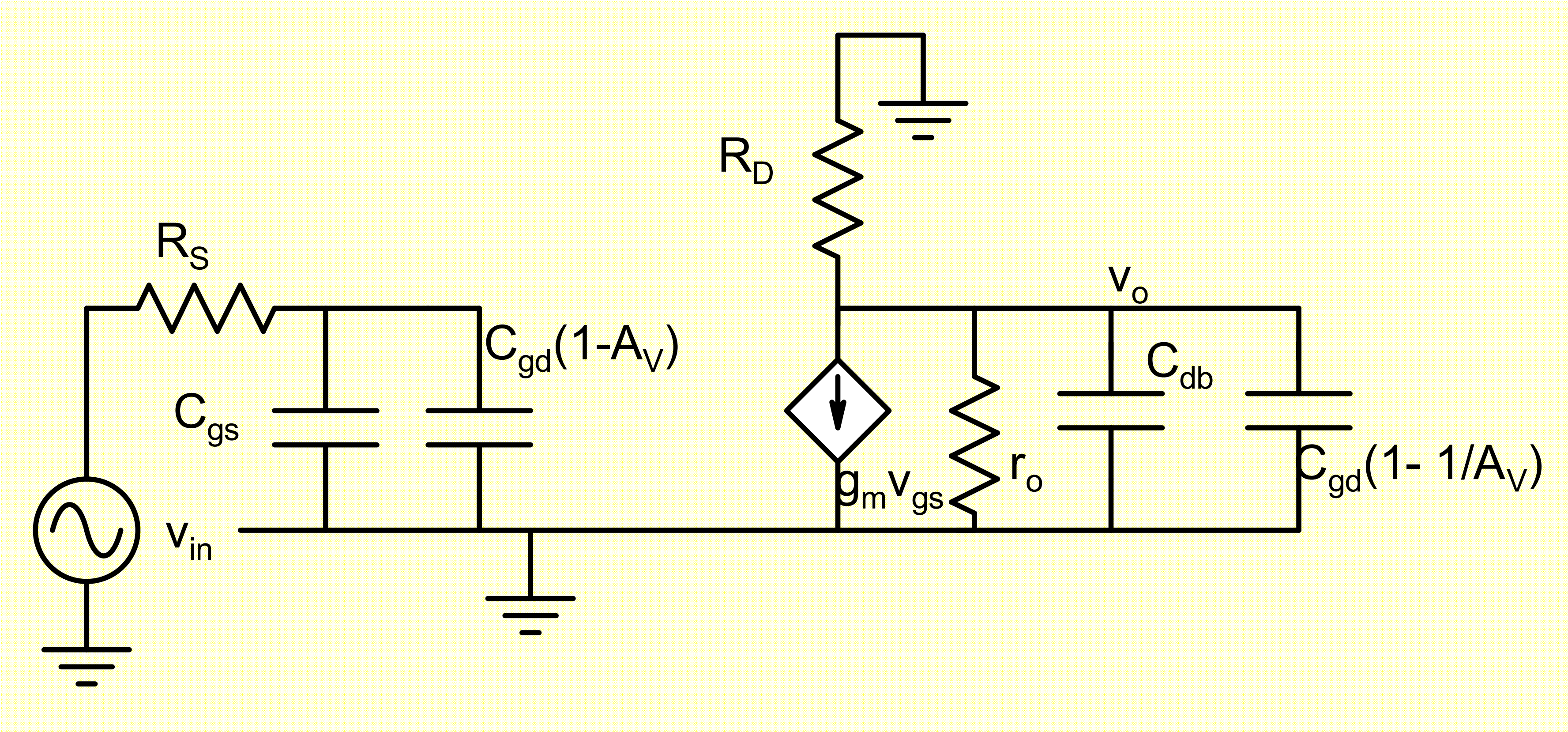
$$\tau_{in} = R_D \times C_{out}$$



$$f_{3dB} = \frac{1}{2\pi \times (R_S C_{in} + R_D C_{out})}$$

G-Number

Key approximation in Miller's theorem



What is A_V ?

For the estimation of capacitance, normally low frequency value for $A_V = -g_m R_D$ is used

MOS Capacitances

$$C_{gs} \cong \frac{2}{3} C_{ox'}.W.L + C_{gso}.W \qquad C_{gd} = C_{GDO}.W$$

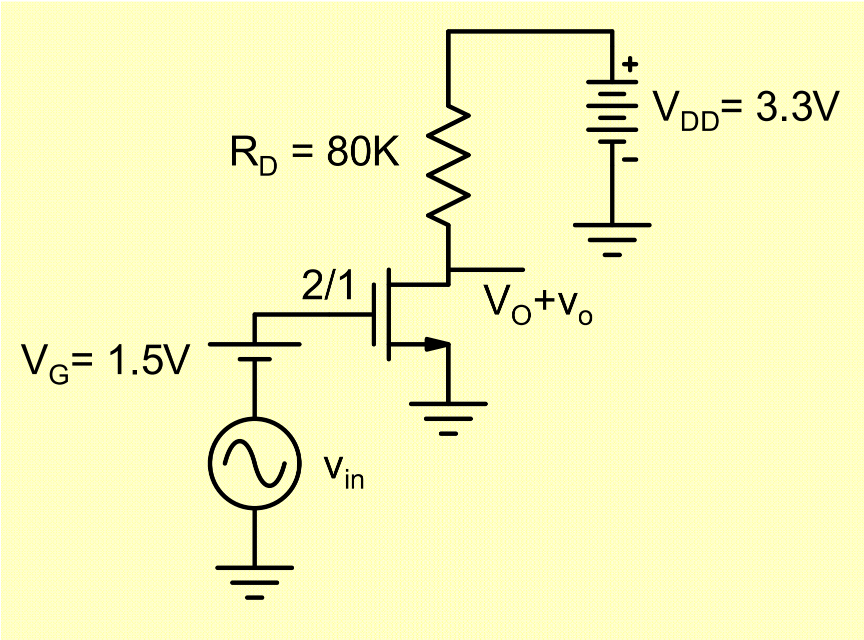
$$C_{sb} = \frac{C_j.A_s}{\left(1 + \frac{V_{SB}}{P_B}\right)^{M_j}} + \frac{C_{jsw}.P_S}{\left(1 + \frac{V_{SB}}{P_{BSW}}\right)^{M_{jsw}}}, \qquad P_S = 2L_S + W \quad , \quad A_s = W.L_S$$

$$C_{db} = \frac{C_{jsw}.P_D}{\left(1 + \frac{V_{DB}}{P_{BSW}}\right)^{M_{jsw}}} + \frac{C_j.A_D}{\left(1 + \frac{V_{DB}}{P_B}\right)^{M_j}} \qquad P_D = 2L_D + W$$

The capacitance model presented herein requires 10 parameters:

$$C_{GSO}, C_{GDO}, C_{GBO}, C'_{OX}, C_J, PB, M_J, C_{JSW}, P_{BSW}, M_{JSW}$$

3dB Upper Cutoff Frequency



$$C_{gs} == 4\,fF \quad C_{gd} = 0.4\,fF$$

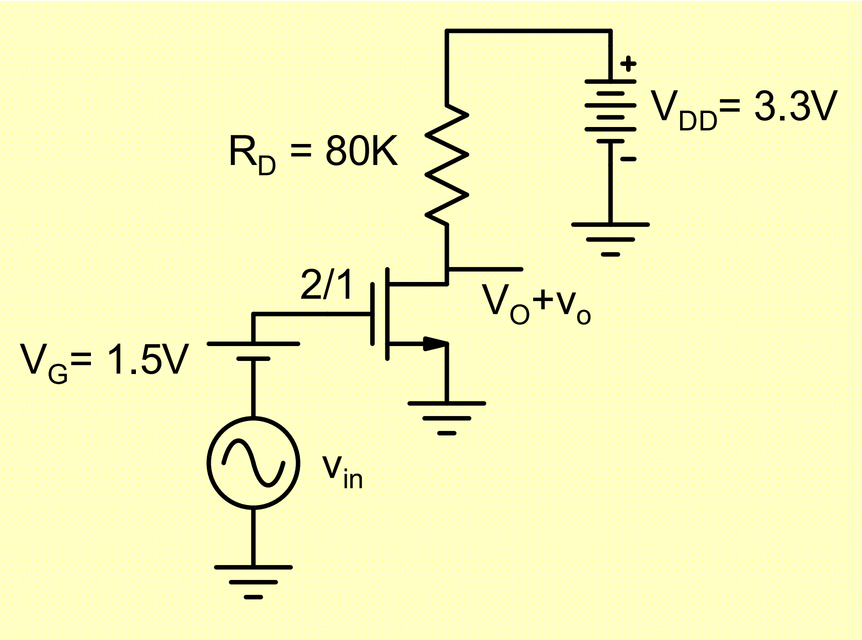
$$C_{db} = 4\,fF$$

$$f_{3dB} = \frac{1}{2\pi} \times \frac{1}{R_S(C_{gs} + C_{gd}(1 + g_m R_D)) + R_D(C_{gd} + C_{db})}$$

$$= 0.45GHz$$

From simulation $f_{3dB} = 0.44GHz$

Example



$$I_{DSQ} = 25\mu A$$

$$V_{DSQ} = 1.3V ; V_{sat} = 0.5V$$

$$A_V \cong -g_m R_D = -8 \qquad R_O \cong R_D = 80K$$

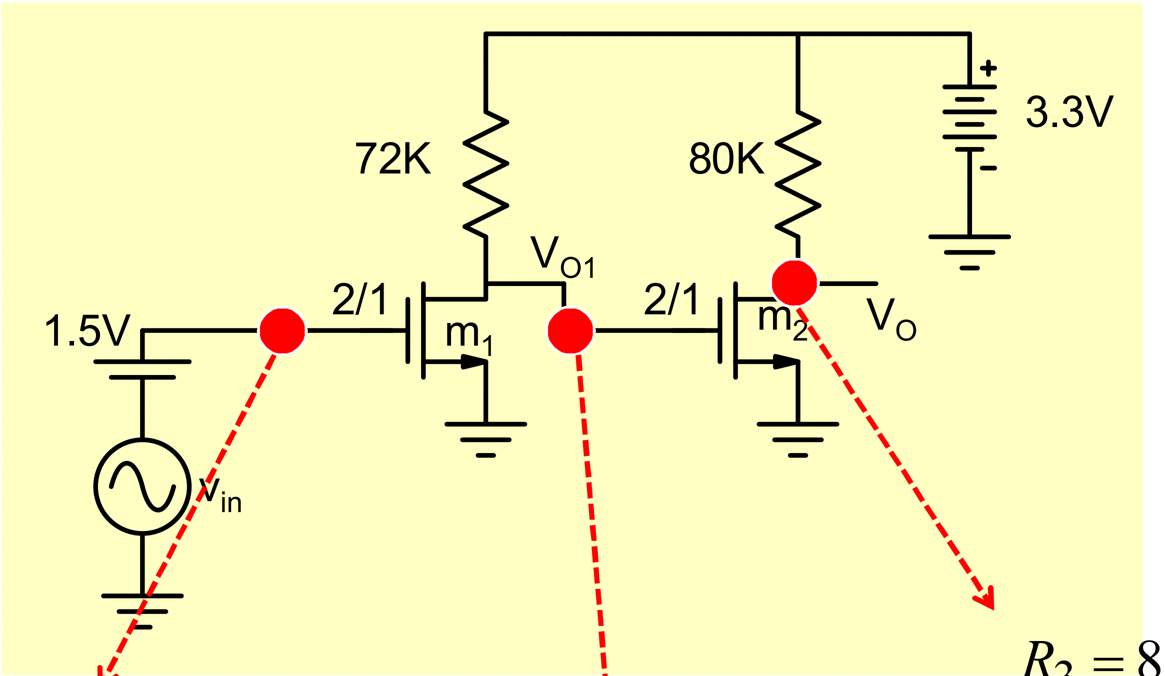
$$v_{o\max} \le V_{DSQ} - V_{sat} = 0.8V$$

$$v_{o\max 3} \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5}$$

$$= 0.8V \text{ for } \mathrm{HD}_2 = 5\%$$

$$f_{3dB} = \frac{1}{2\pi} \times \frac{1}{R_S(C_{gs} + C_{gd}(1 + g_m R_D)) + R_D(C_{gd} + C_{db})}$$

$$= 0.45GHz$$



$$f_{3dB} = \frac{1}{2\pi \sum \tau_j}$$

$$\tau_j = R_j C_j$$

$$R_3 = 80k$$

$$R_1 = 0$$

$$C_1 = C_{gs1} + C_{gd1} \times (1 - A_{V1})$$

$$C_3 = C_{db2} + C_{gd2} \times (1 - \frac{1}{A_{V2}})$$

$$R_2 = 72k$$

$$C_2 = C_{gs2} + C_{gd2} \times (1 - A_{V2}) + C_{db1} + C_{gd1} \times (1 - \frac{1}{A_{V1}})$$

EE210: Microelectronics-I

Lecture-42 : MOS Amplifiers_2

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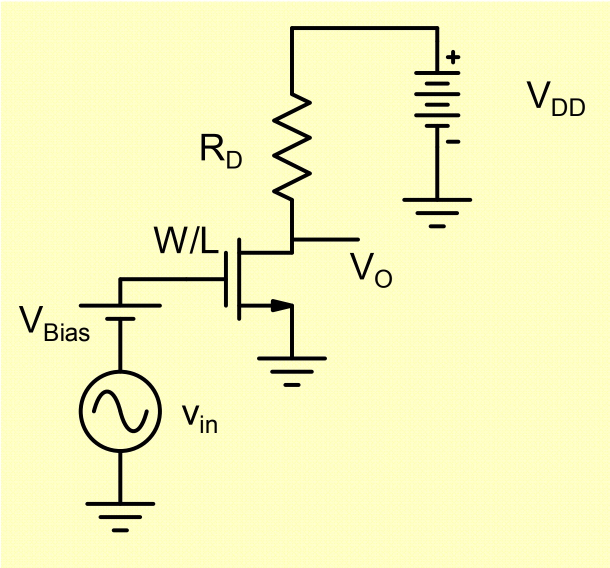


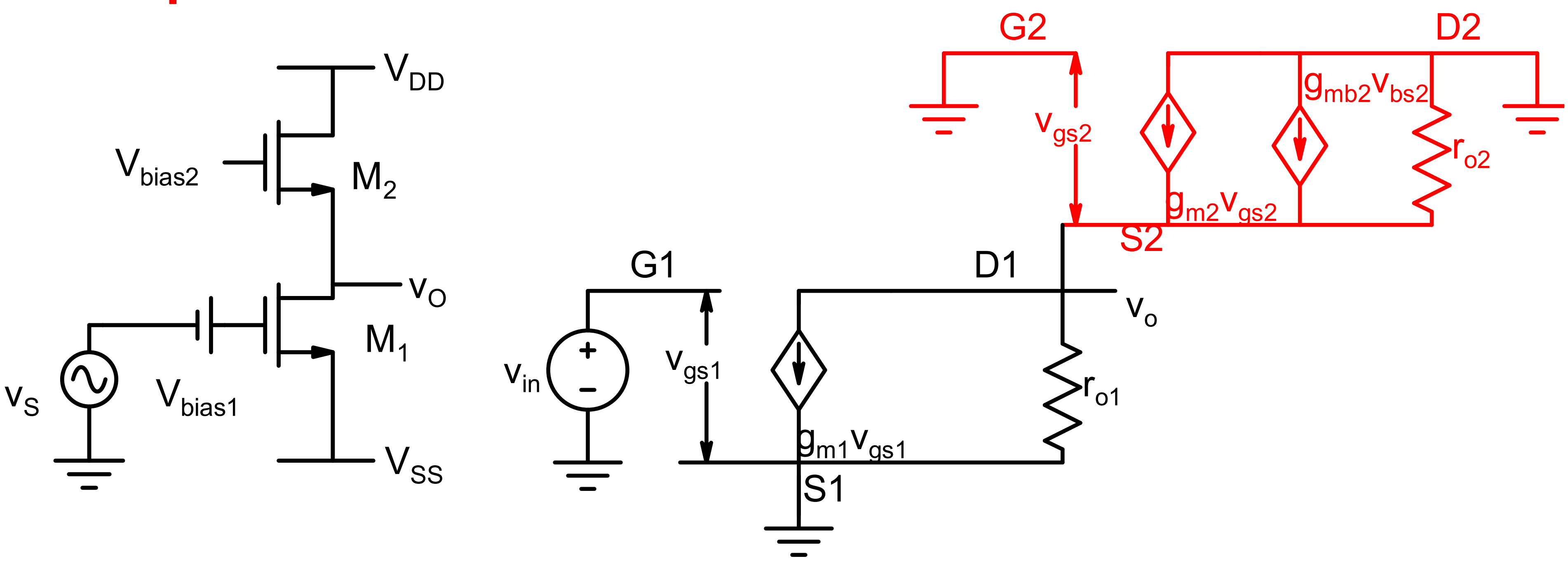
$$A_v \le \sqrt{\frac{2\beta}{I_{DSQ}}} \times (V_{DD} - v_{om} - \sqrt{\frac{2I_{DSQ}}{\beta}})$$

W=2µm and v_{om} = 0.1V

I_{DS} (µA)	AV	RD (MΩ)	V_{GS} (V)
100	4.4	0.022	2
50	7.05	0.05	1.707
25	10.8	0.108	1.5
10	18.2	0.29	1.316
5	26.6	0.59	1.224
2	43.2	1.53	1.14
1	62	3.1	1.1

CS amplifier with Active Load

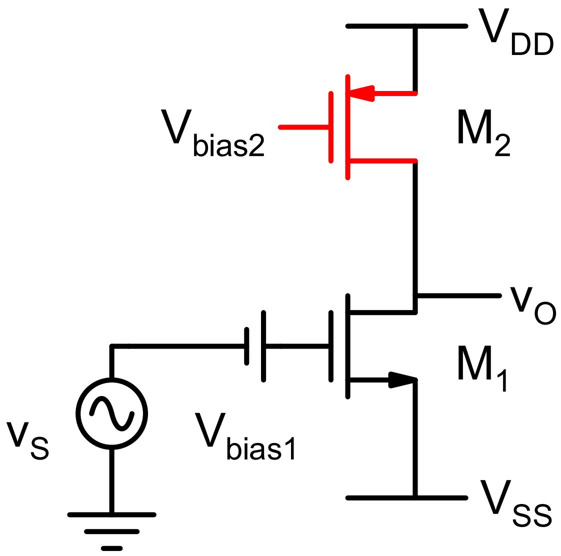
CS Amplifier with Active Load

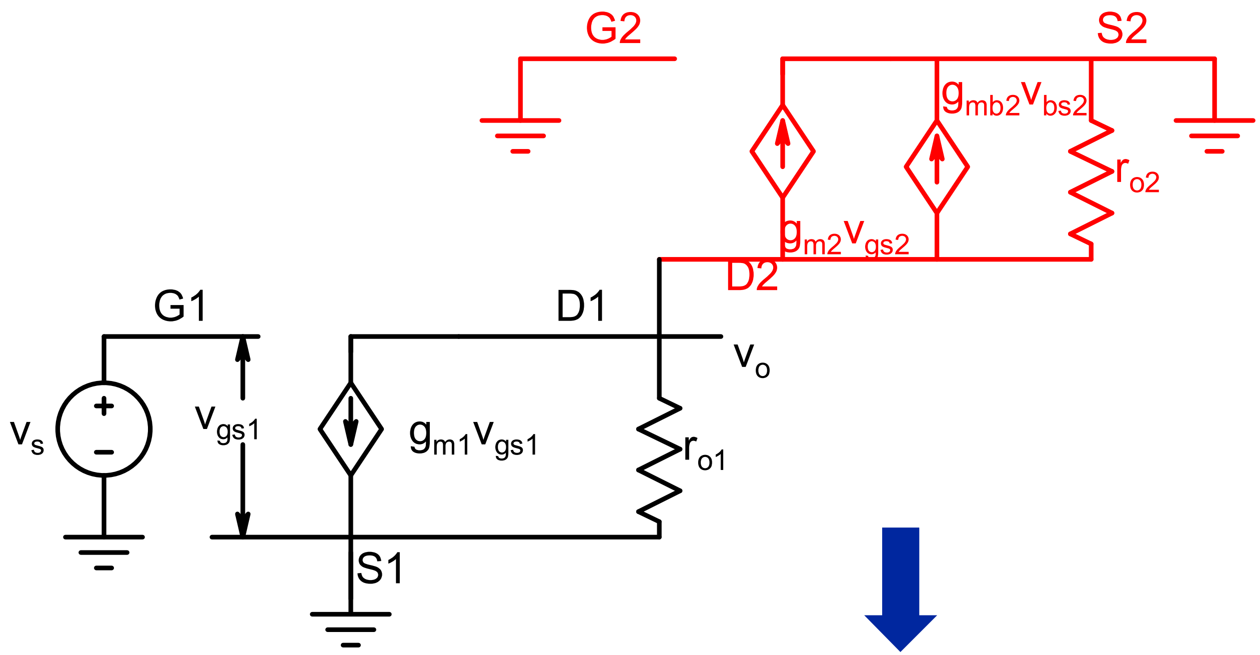


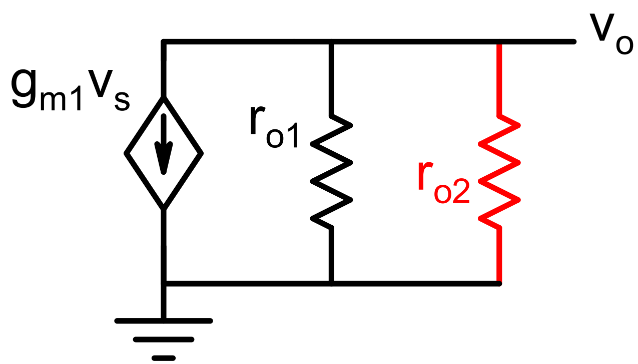
$$v_o = -g_{m1} v_{in} \times \left\{ r_{o1} \,\|\, r_{o2} \,\|\, \frac{1}{g_{m2}(1+\eta)} \right\}$$

$$A_V \cong -\frac{g_{m1}}{g_{m2}(1+\eta)}$$

$$= \frac{\sqrt{2KP \times (W/L)_1 \times I_{DSQ1}}}{\sqrt{2KP \times (W/L)_2 \times I_{DSQ1}}} \times \frac{1}{1+\eta} = \sqrt{\frac{(W/L)_1}{(W/L)_2}} \times \frac{1}{1+\eta}$$

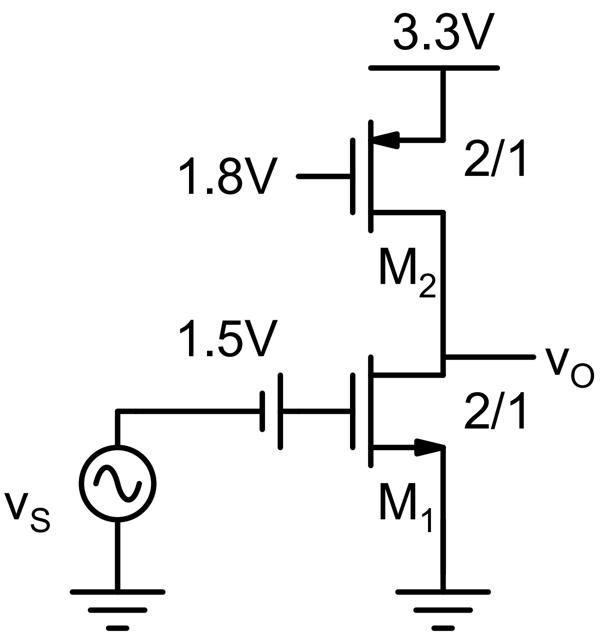






$$A_V = -g_{m1} \times r_{o1} \| r_{o2}$$

Example



$$I_{DSQ} = 25\mu A; V_{DSQ} = 1.65(?)$$

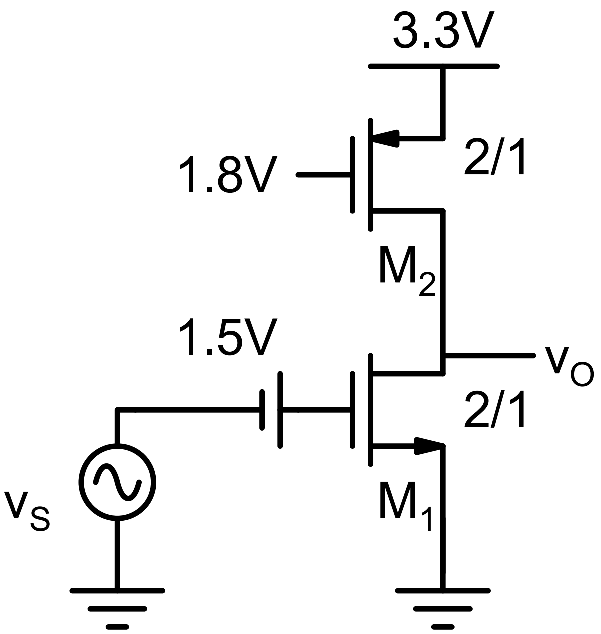
$$g_{mn} = g_{mp} = 100\mu A/V \; ; \; r_{on} = r_{op} = 4M\Omega$$

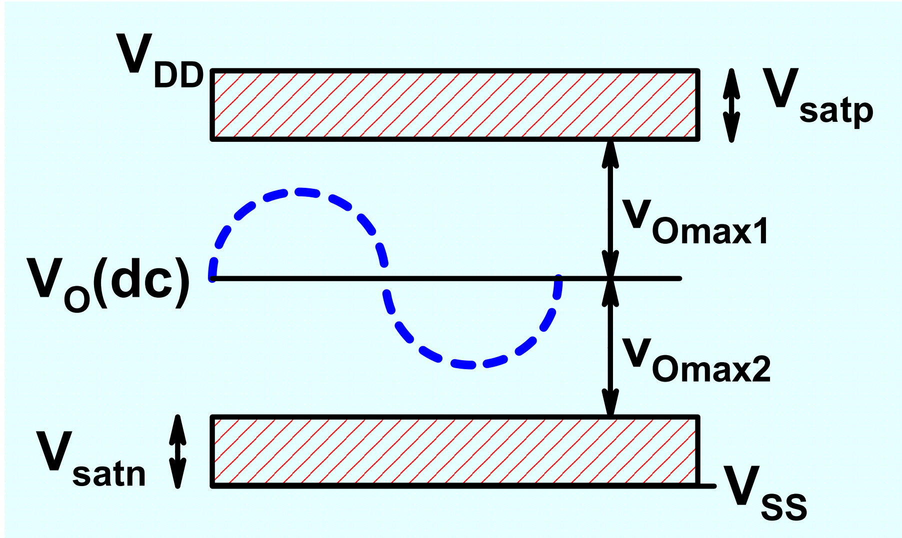
$$\left|A_V\right| = g_{mn} \times r_{on} \| r_{op}$$
$$= -200$$

$$R_O = r_{on} \| r_{op} = 2M\Omega$$

Large gain is easily obtained without requiring large resistors which are difficult to fabricate

Output Voltage Swing





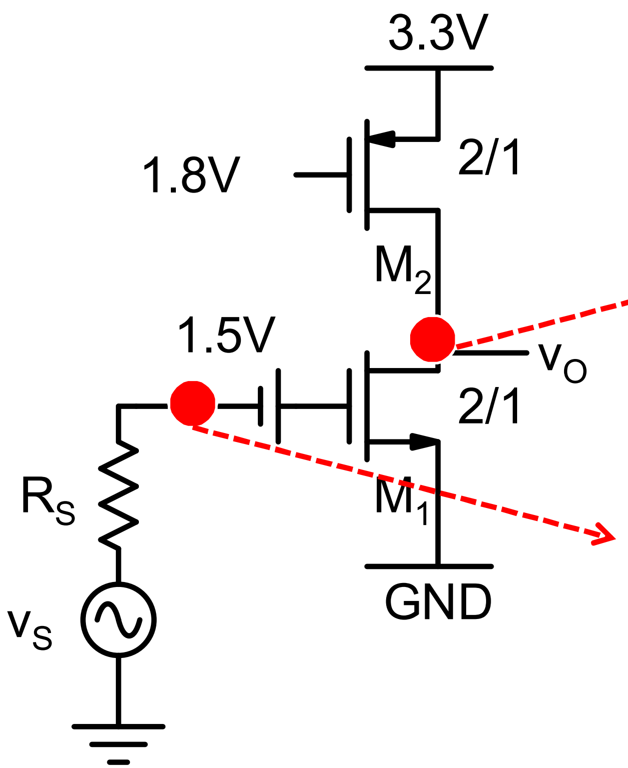
$$v_{o\max 1} = V_{DD} - V_O(dc) - V_{satp}$$ 3.3-1.65-0.5=1.15V

$$v_{o\max 2} = V_O(dc) - V_{ss} - V_{satn}$$ 3.3-1.65-0.5=1.15V

$$v_{o\max 3} \cong I_{DSQ}\, r_{op} \| r_{on} \times \frac{HD_2}{12.5}$$ 20V for HD_2 = 5%

Harmonic distortion is less of a problem here

Frequency Response



$$f_{3dB} = \frac{1}{2\pi \sum \tau_j} \qquad \tau_j = R_j C_j$$

$$R_2 = r_{on} \| r_{op} = 2M\Omega$$

$$C_2 = C_{dbn} + C_{gdn} \times (1 - \frac{1}{A_V}) + C_{gdp} + Cdbp$$

$$R_1 = R_S$$

$$C_1 = C_{gsn} + C_{gdn} \times (1 - A_V)$$

$$C_{gsn} = 4.0\,fF \qquad C_{gdn} = 0.4\,fF \qquad C_{dbn} = 4\,fF$$

$$C_{gdp} = 0.4\,fF \qquad C_{dbp} = 4\,fF$$

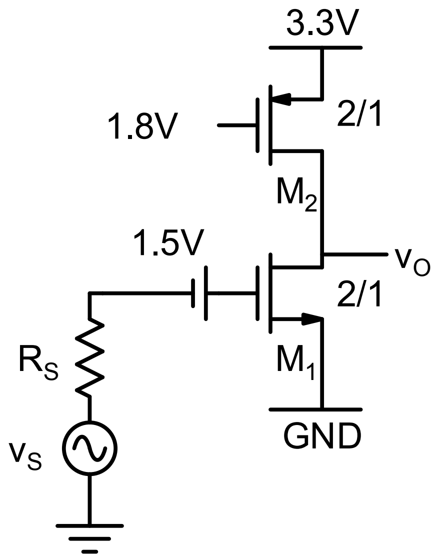
$$C_1 = 84.4\,fF; C_2 = 8.8\,fF$$

$$f_{3dB} = \frac{1}{2\pi \sum \tau_j} = 9MHz \text{ for } R_S = 0$$

$$= 1.56MHz \text{ for } R_S = 1M\Omega$$

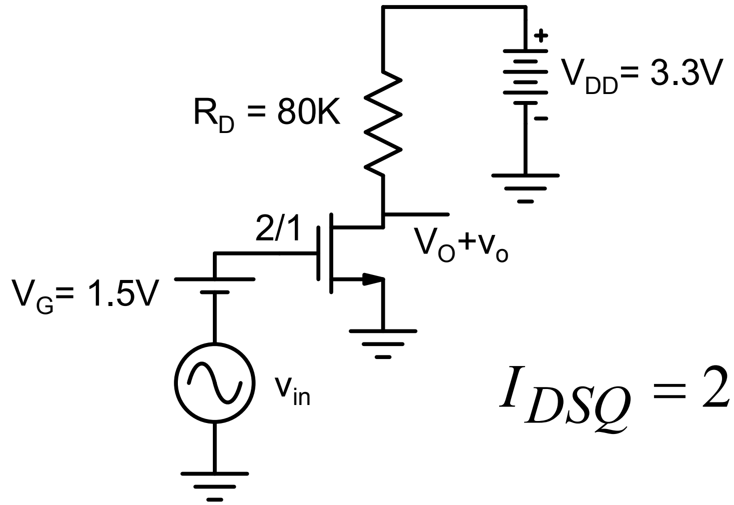
G-Number

Analysis: Summary



$$I_{DSQ} = 25\mu A$$

$$V_O(dc) = 1.65V$$



$$I_{DSQ} = 25\mu A$$

$$V_{DSQ} = 1.3V; V_{sat} = 0.5V$$

↑ $|A_V| = -200$ $R_O = 2M\Omega$ ↓

↓ $A_V \cong -g_m R_D = -8$ $R_O \cong R_D = 80K$ ↑

$$v_{o\max 1} = 1.15V$$

$$v_{o\max 2} = 1.15V$$

$$v_{o\mathrm{m}1} \leq 0.8V$$

↑ $v_{o\max 3} \cong 20V$ for $\mathrm{HD}_2 = 5\%$

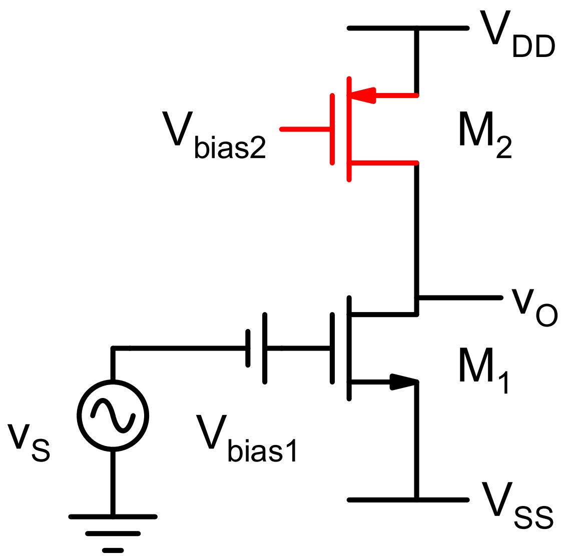
$$v_{o\max 3} \cong I_{DSQ} R_D \| R_L \times \frac{HD_2}{12.5}$$

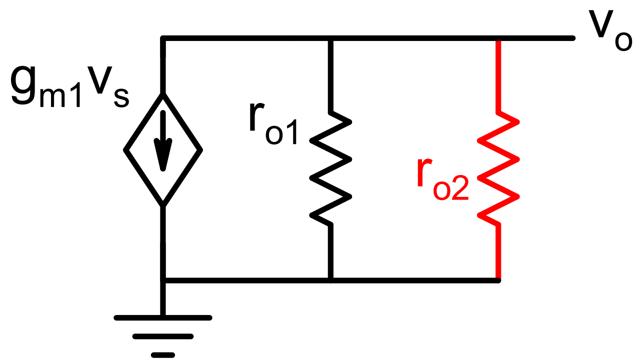
↓ $= 0.8V$ for $\mathrm{HD}_2 = 5\%$

$$f_{3dB} = 9MHz \text{ for } R_S = 0$$
$$= 1.56MHz \text{ for } R_S = 1M\Omega$$

↑ $f_{3dB} = 0.45GHz$ for $R_S = 0$

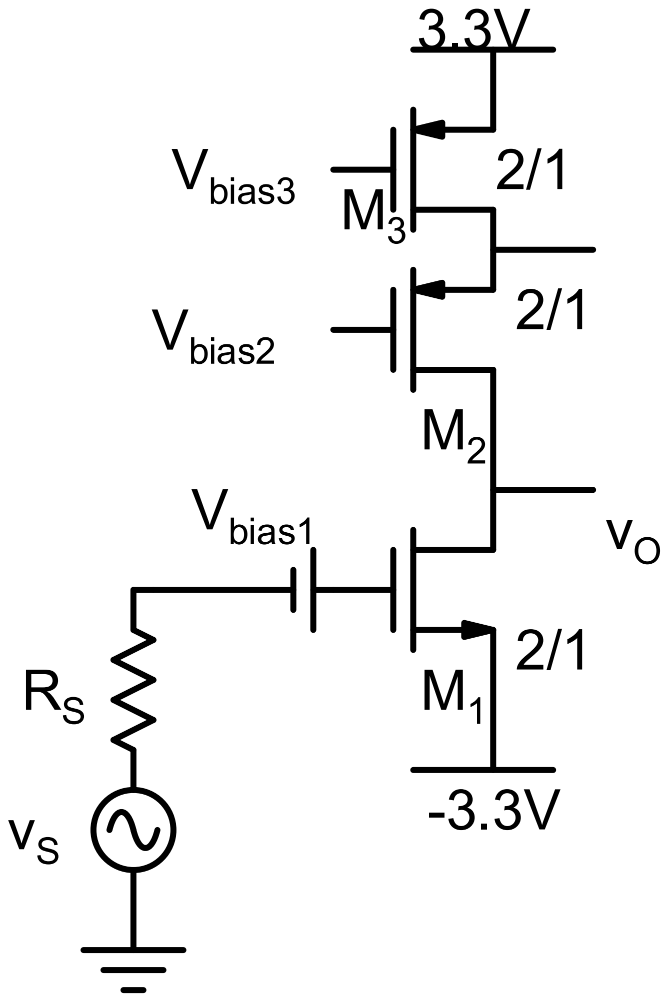
G-Number





$$A_V = -g_{m1} \times r_{o1} \| r_{o2}$$

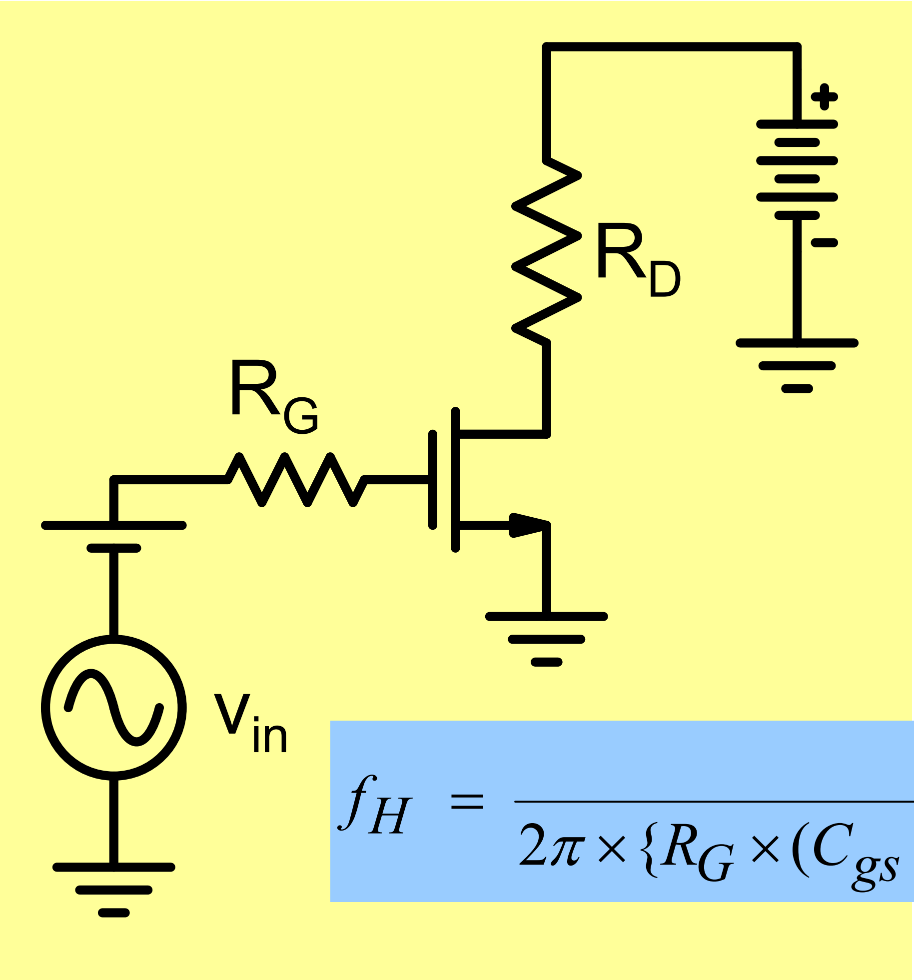
Higher Voltage gain can be obtained by stacking transistors vertically.



Common Gate Amplifier

Why do we want another amplifier configuration?

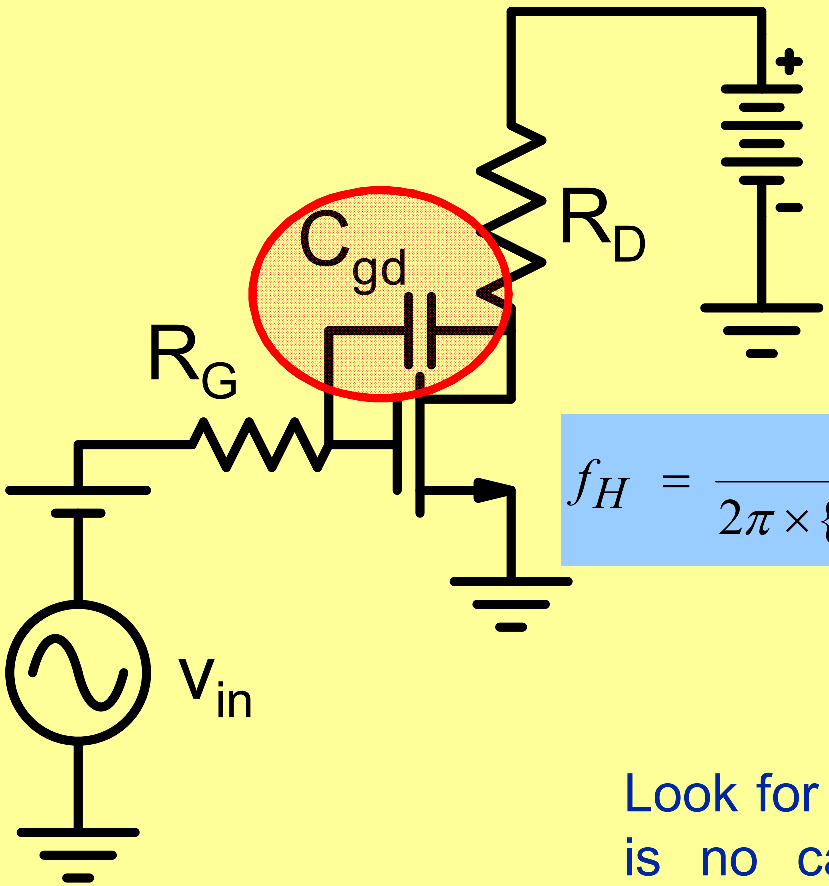
Problem with CS amplifier:



$$f_H = \frac{1}{2\pi \times \{R_G \times (C_{gs} + C_{gd}(1 - A_V)) + R_D \times (C_{gd} + C_{db})\}}$$

Increase in voltage gain reduces bandwidth

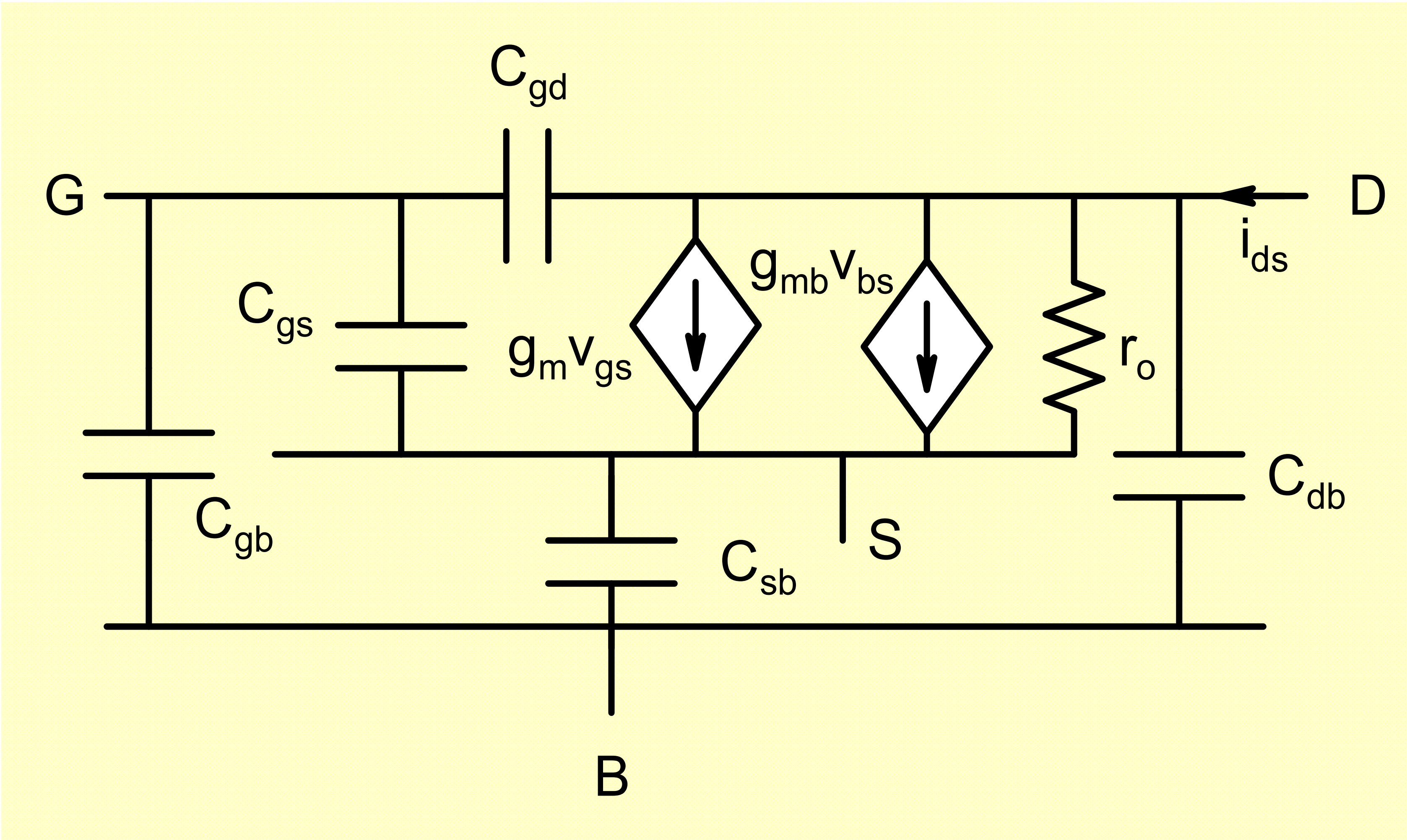
How do we eliminate the Miller Capacitance?



$$f_H = \frac{1}{2\pi \times \{R_G \times (C_{gs} + C_{gd}(1 - A_V)) + R_D \times (C_{gd} + C_{db})\}}$$

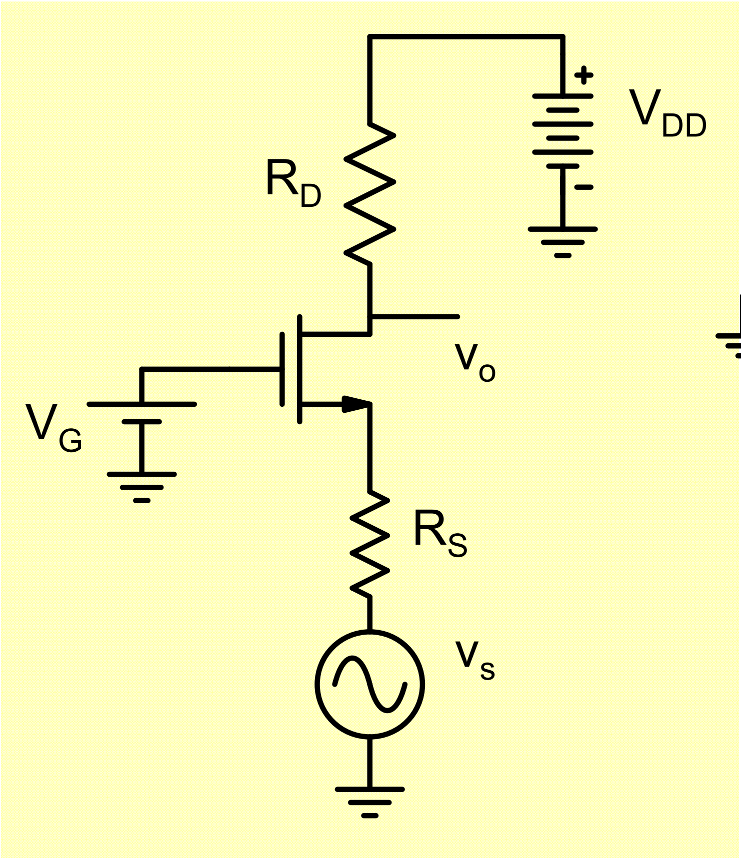
Look for an amplifier configuration in which there is no capacitance between input and output nodes

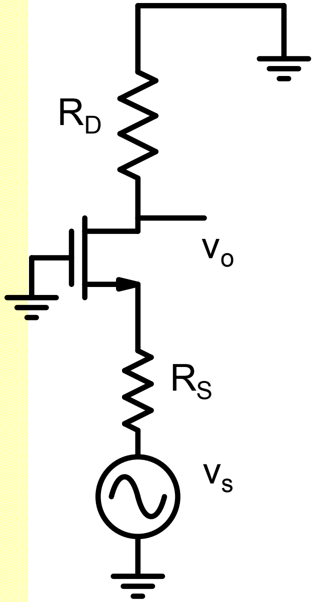
MOS high frequency small signal model

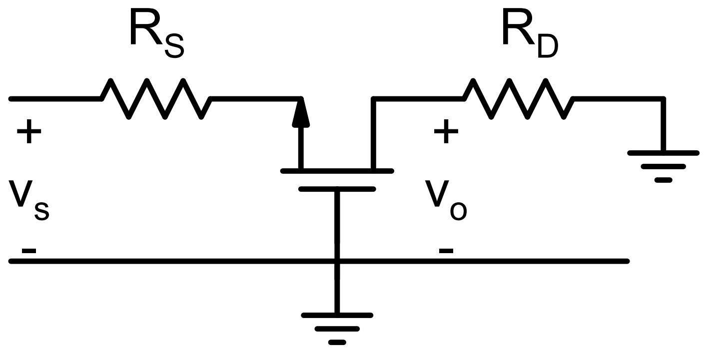


Apply input at source and take output at drain

CG amplifier



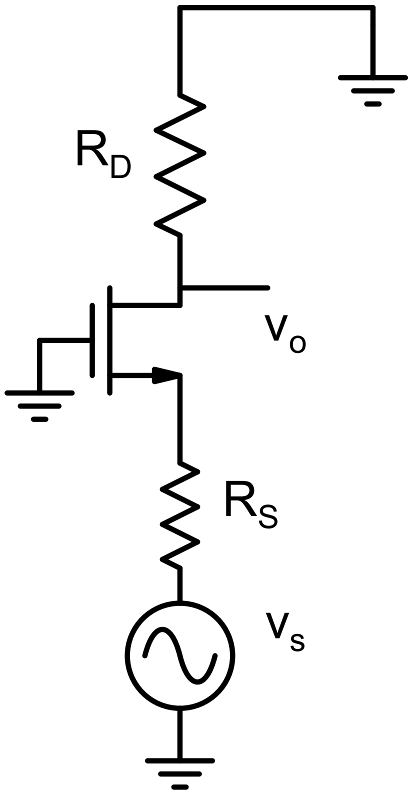


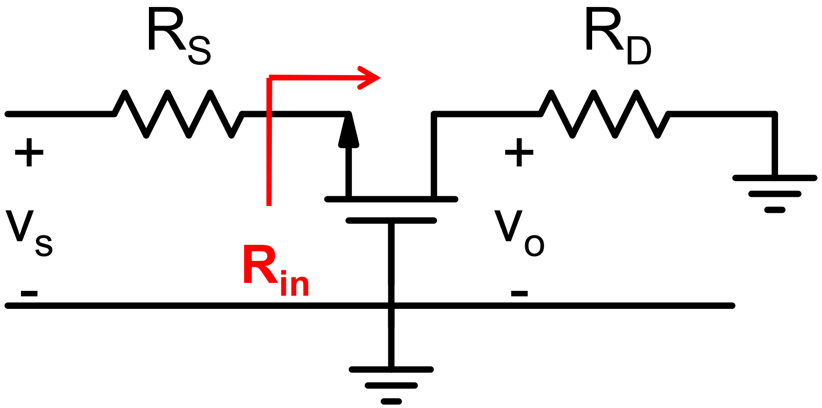


Gate is common to both input and output ports and hence the name Common Gate Amplifier

$$A_V \cong \frac{(g_m + g_{mb})}{1 + (g_m + g_{mb})R_S} \times R_D \qquad R_O \cong R_D$$

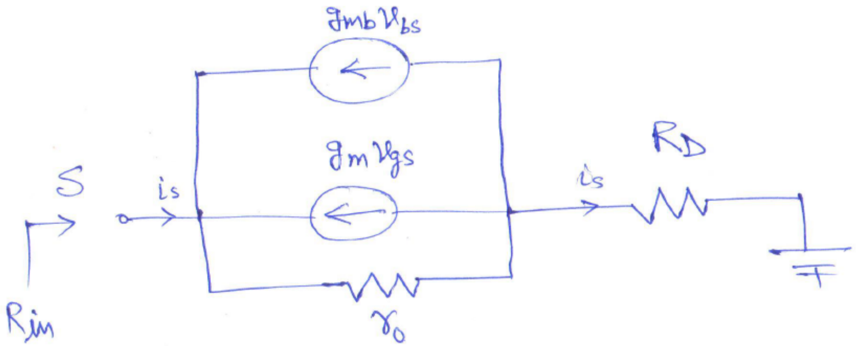
$$g_m = 100\mu A/V; g_{mb} = 42\ \mu A/V \ \text{ for } I_{DSQ} = 25\mu A$$





$$R_{in} \cong \frac{1 + R_D/r_o}{g_m + g_{mb}}$$

~7k



$$R_{in} = \frac{v_s}{i_s} \simeq \frac{1 + \frac{R_D}{r_o}}{g_m + g_{mb}}$$

$$v_s = (i_s + g_m v_{gs} + g_{mb} v_{bs}) r_o + R_D \cdot i_s$$

$$= (i_s - g_m v_s - g_{mb} v_s) r_o + R_D \cdot i_s$$

$$v_s \left(\frac{1}{r_o} + g_m + g_{mb}\right) = i_s \left(1 + \frac{R_D}{r_o}\right)$$

$$\frac{v_s}{i_s} = \frac{1 + \frac{R_D}{r_o}}{\frac{1}{r_o} + g_m + g_{mb}}$$

$v_{bs} = -v_s$

$v_{gs} = -v_s$

$g_{mb} = g_m \cdot \eta$

G-Number

CG

$$A_V \cong \frac{(g_m + g_{mb})}{1+(g_m + g_{mb})R_S} \times R_D$$

$$R_O \cong R_D$$

$$R_{in} \cong \frac{1+R_D/r_o}{g_m + g_{mb}}$$

CS

$$A_V \cong -g_m R_D$$

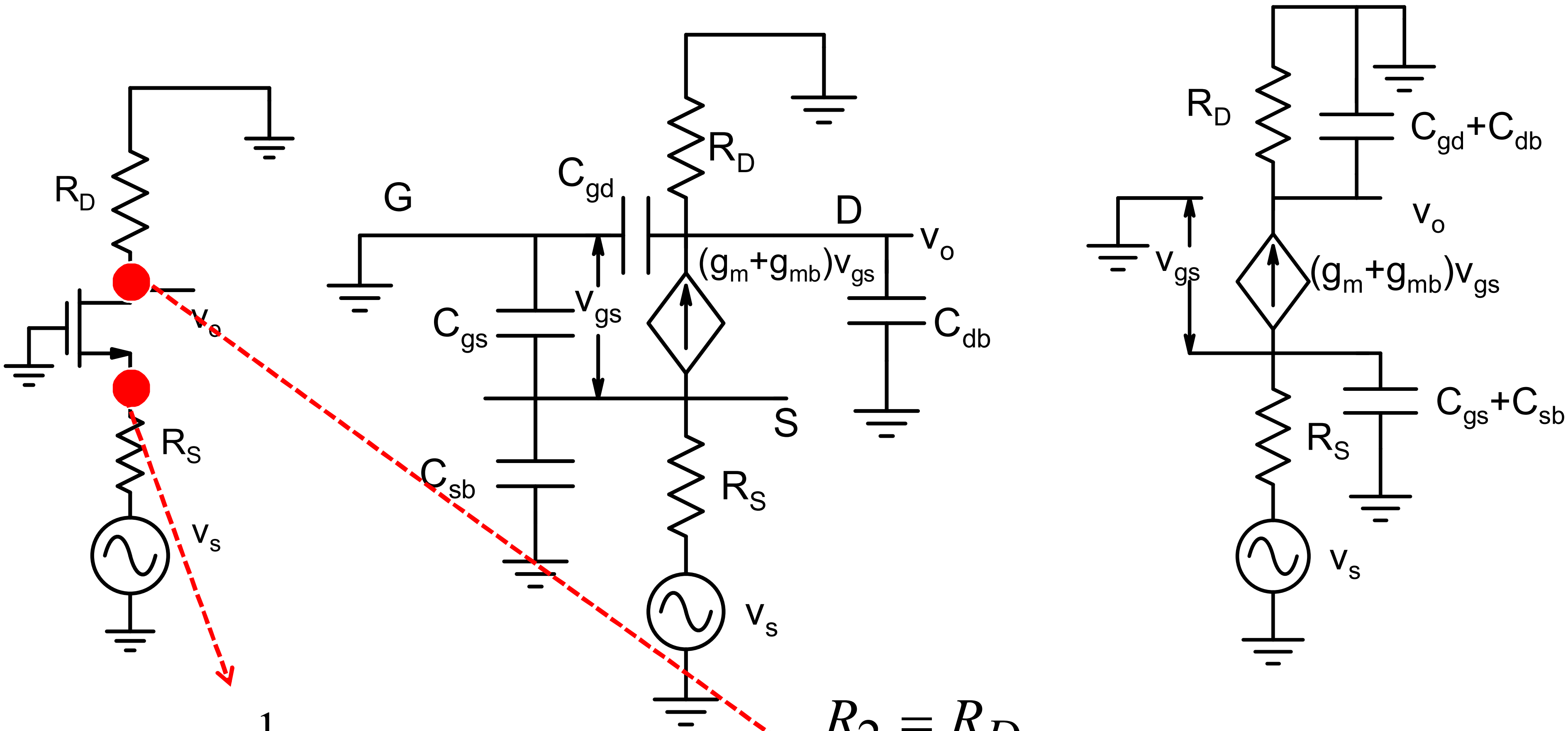
$$R_O \cong R_D$$

$$R_{in} \to \infty$$

- A CG amplifier thus has similar voltage gain and output resistance as a CS amplifier but has a very low input impedance

$$\frac{A_V \times R_{in}}{R_o} = 1 \quad \text{for negligible } R_S$$

Frequency Response



$$R_1 = \frac{1}{g_m + g_{mb}} \| R_S$$

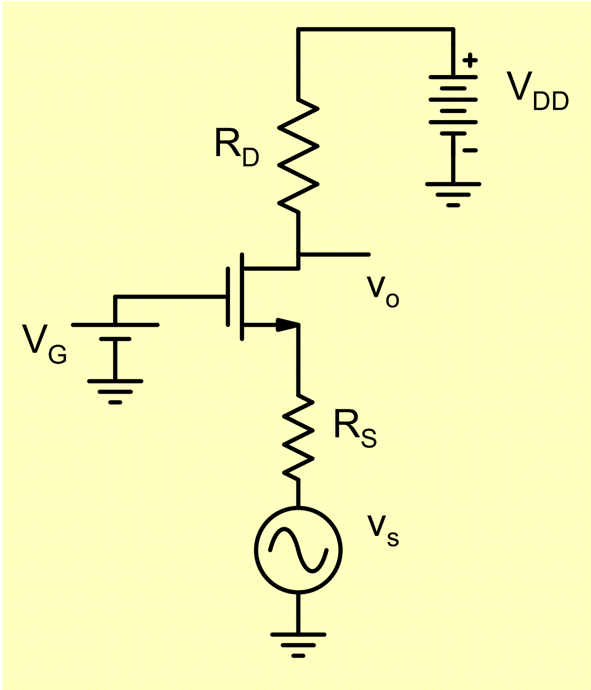
$$C_1 = C_{gs} + C_{sb}$$

$$R_2 = R_D$$

$$C_2 = C_{gd} + C_{db}$$

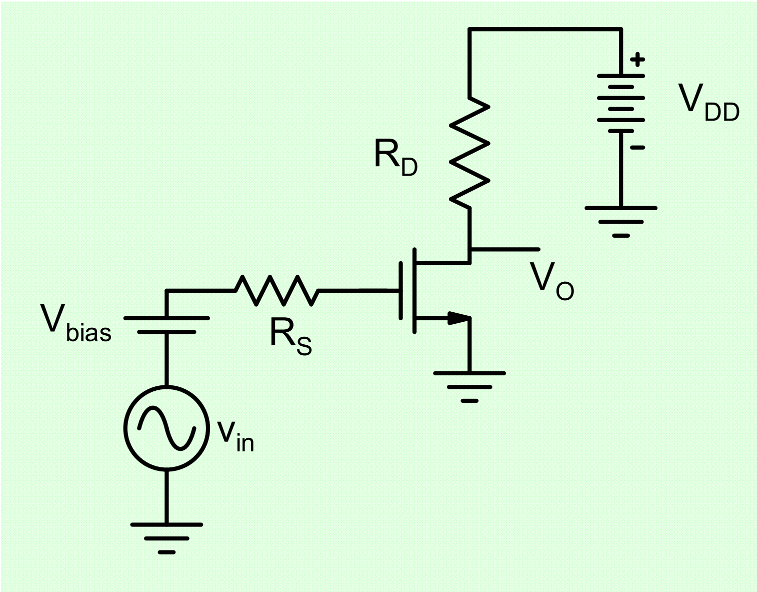
$$f_{3dB} \cong \frac{1}{2\pi R_D (C_{gd} + C_{db})}$$

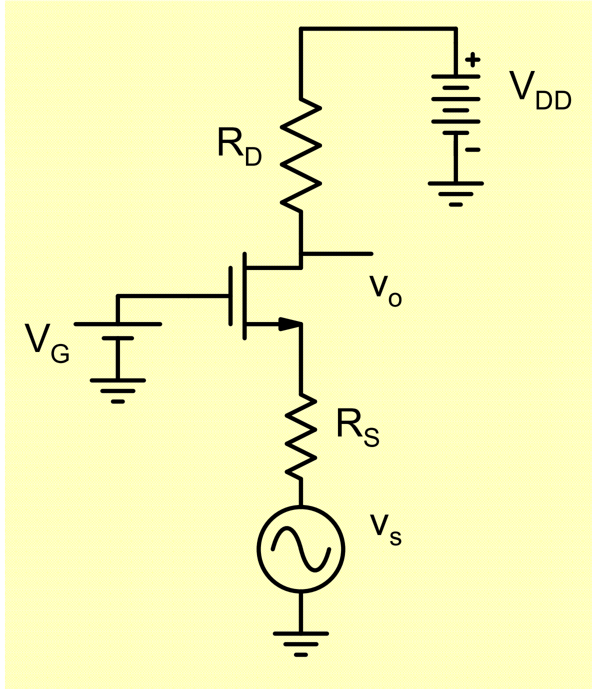
Output Swing



Output swing is similar to CS amplifier determined by transistor entering linear region and harmonic distortion

CS-CG Comparison





$$A_V \cong -g_m R_D$$

$$R_O \cong R_D$$

$$R_{in} \cong \text{very high}$$

$$f_{3dB} = \frac{1}{2\pi R_S (C_{gs} + C_{gd}(1 - A_V)) + R_D (C_{gd} + C_{db})}$$

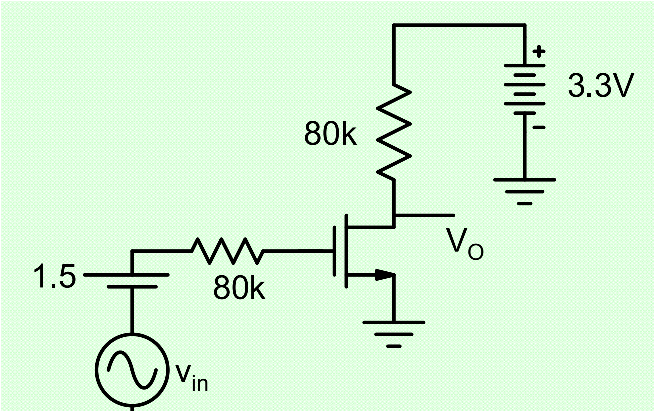
$$A_V \cong \frac{(g_m + g_{mb})}{1 + (g_m + g_{mb}) R_S} \times R_D$$

$$R_O \cong R_D$$

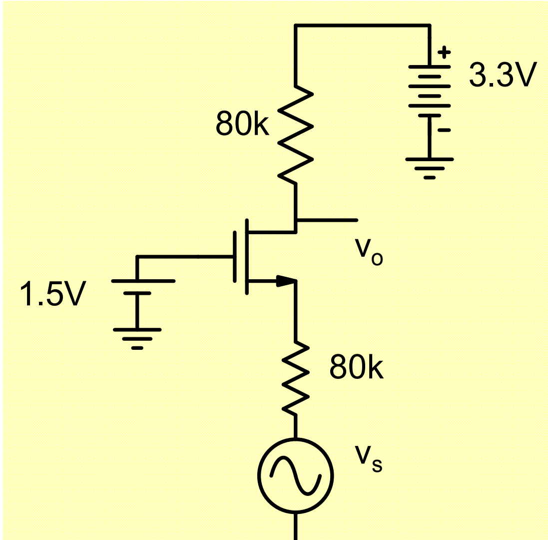
$$R_{in} \cong \frac{1 + R_D / r_o}{g_m + g_{mb}}$$

$$f_{3dB} \cong \frac{1}{2\pi R_D (C_{gd} + C_{db})}$$

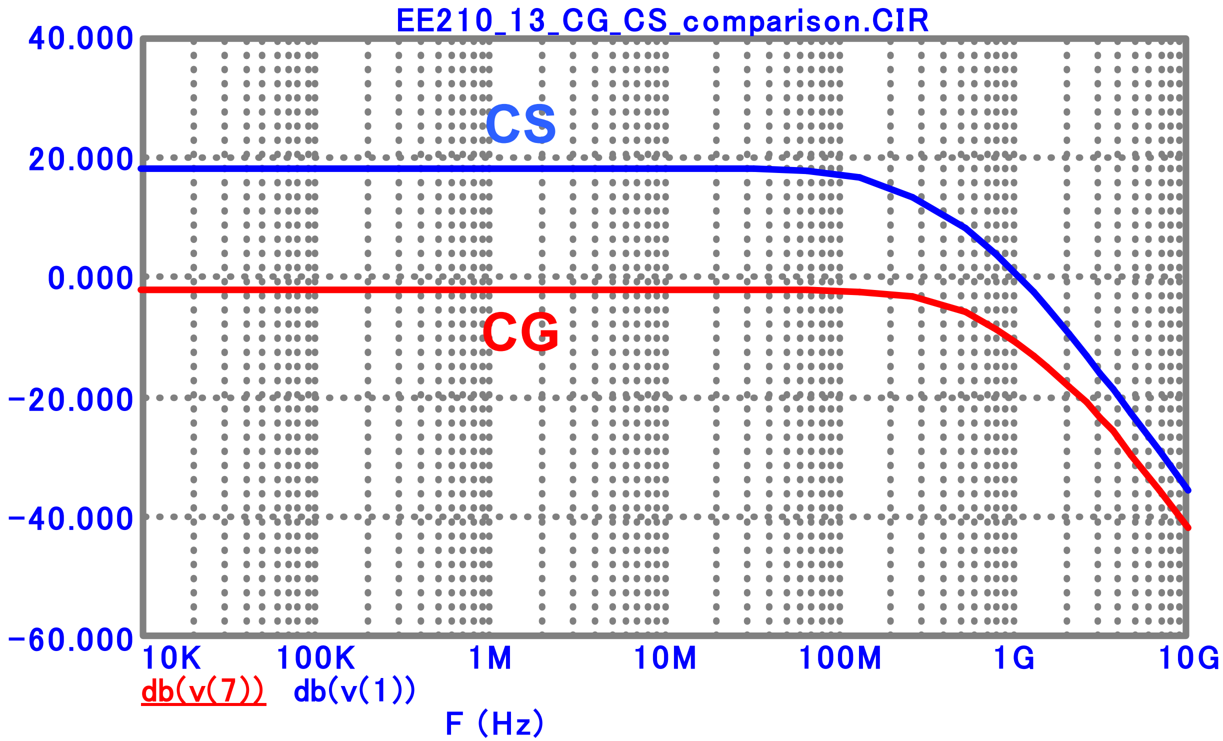
CS-CG Comparison



$$f_{3dB} \sim 180MHz$$



$$f_{3dB} \sim 411MHz$$



EE210: Microelectronics-I

Lecture-43 : MOS Amplifiers_3

http://youtu.be/RswZAEPcefg

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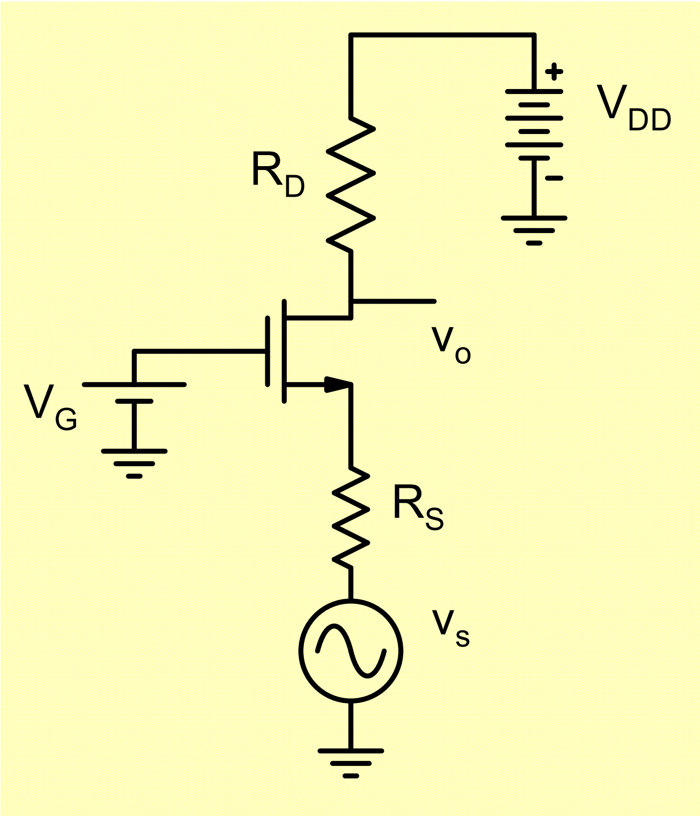
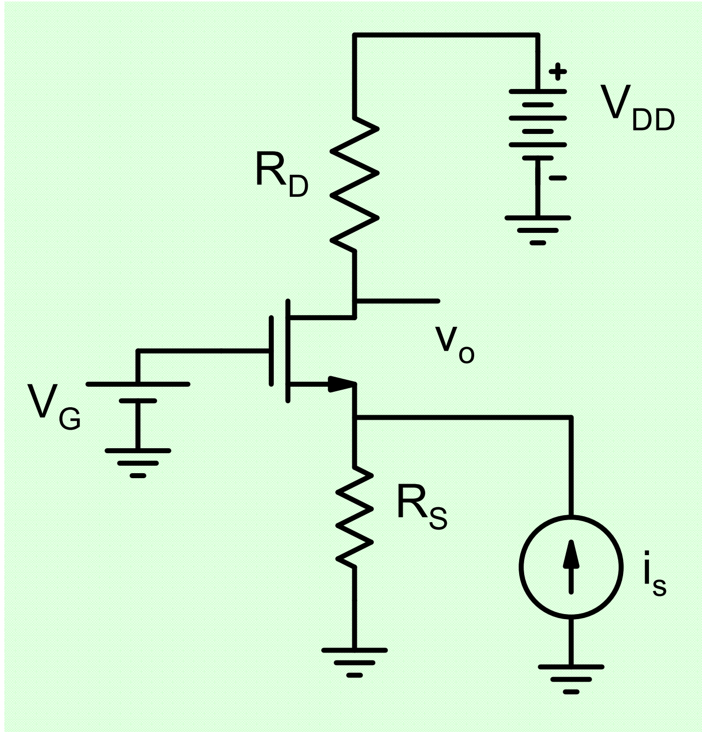
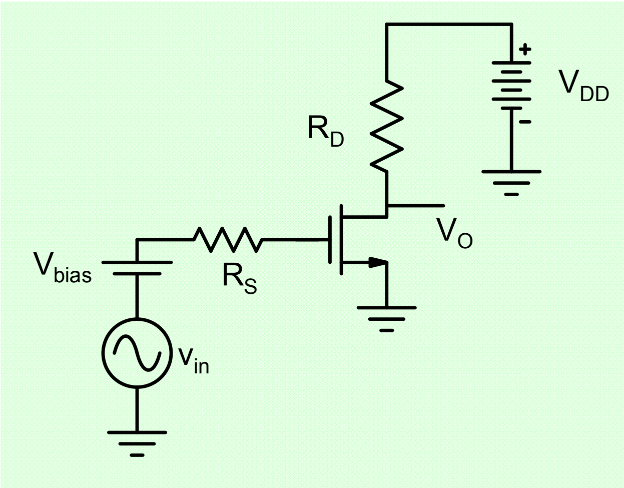
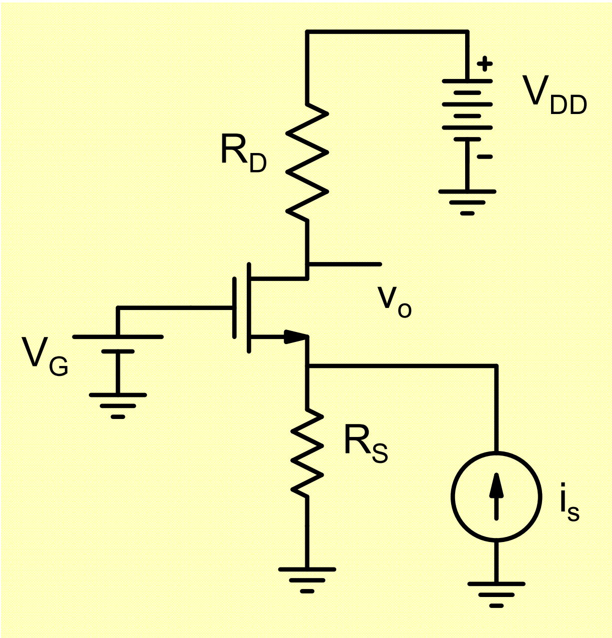


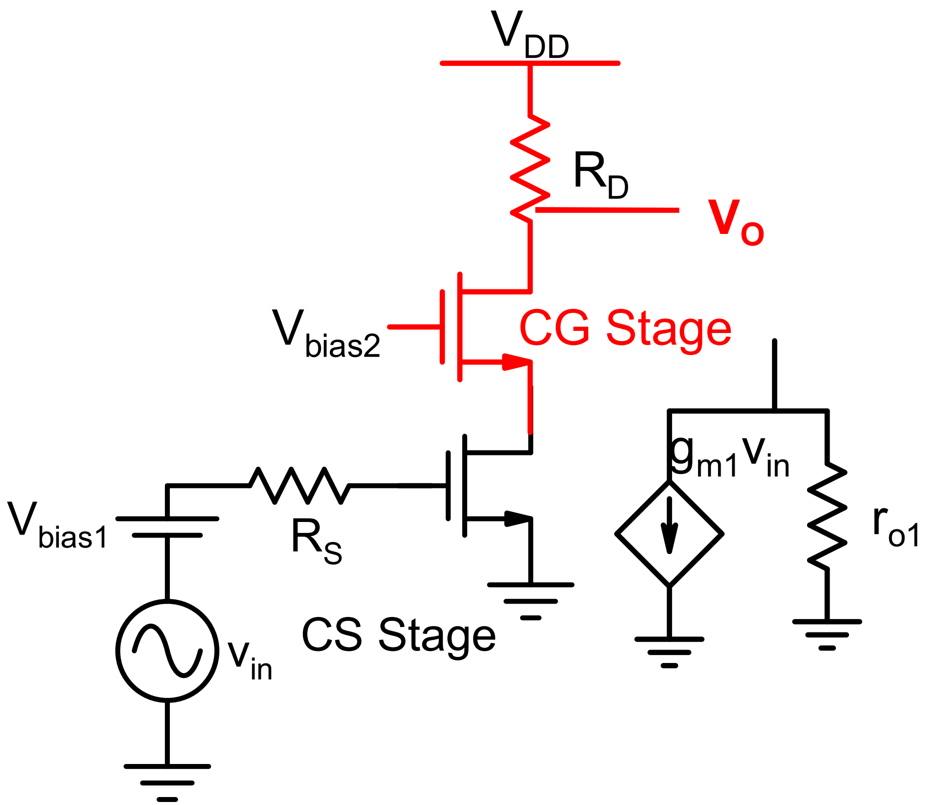


Transconductance Amplifier

Can one combine the advantages of both the configurations?



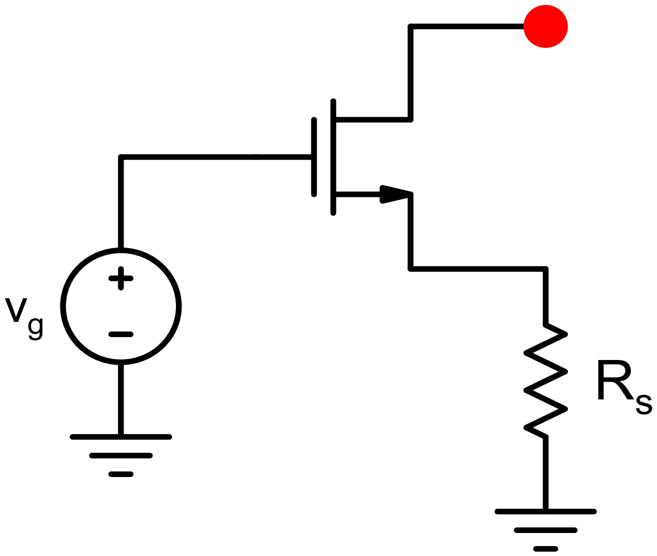


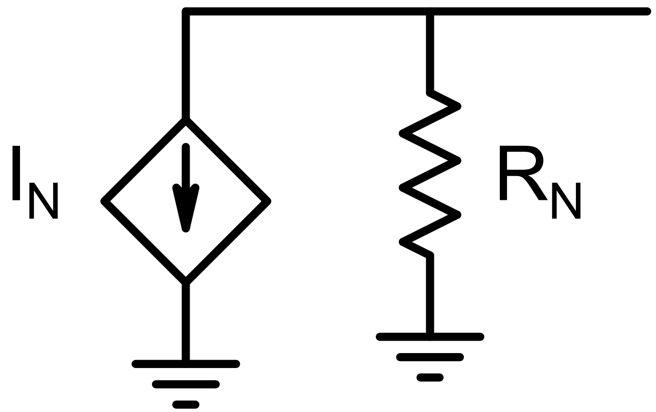


This CS-CG combination is called **CASCODE** Amplifier

When drain of one Tr. Is connected to source of next Tr., the connection is called a **CASCODE**

Useful Results

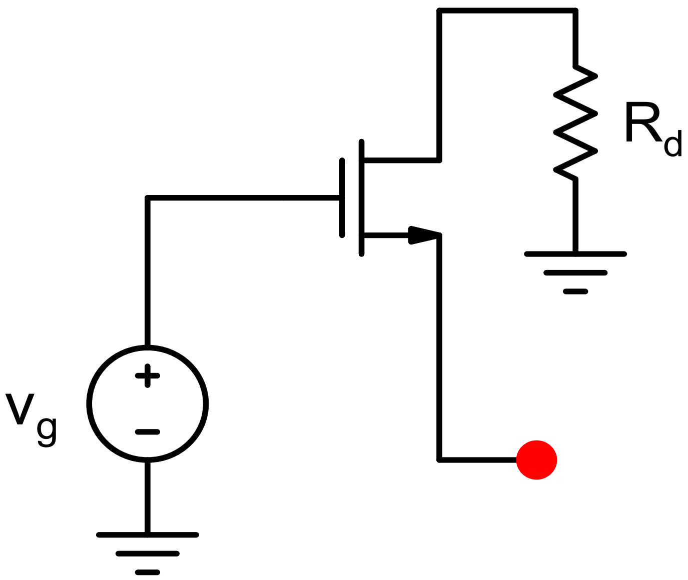


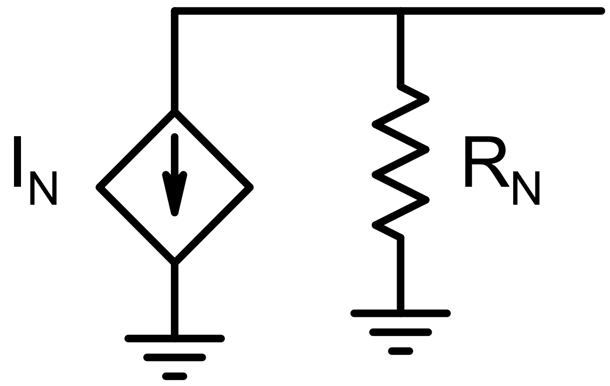


$$i_n \cong \frac{g_m v_g}{1+(g_m+g_{mb})R_S}$$

$$R_n \cong r_o \times \{1+(g_m+g_{mb})R_S\}$$

Useful Results



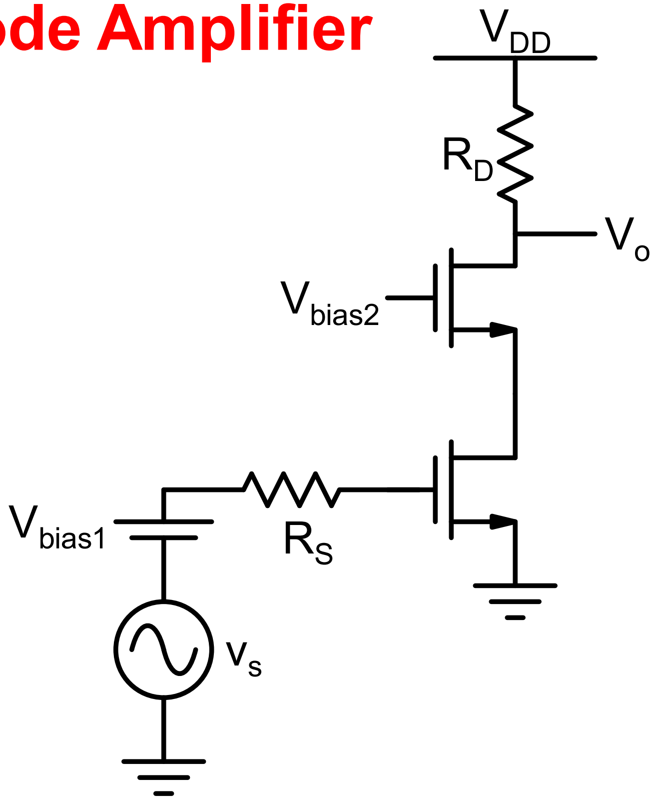


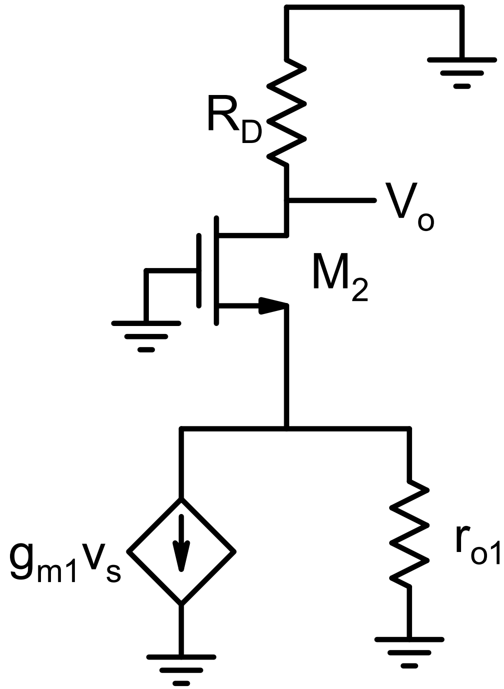
$$i_n = -\frac{g_m v_g}{1 + R_d / r_o}$$

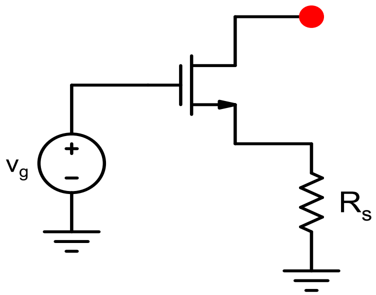
$$R_n \cong \frac{1 + R_d / r_o}{g_m + g_{mb}}$$

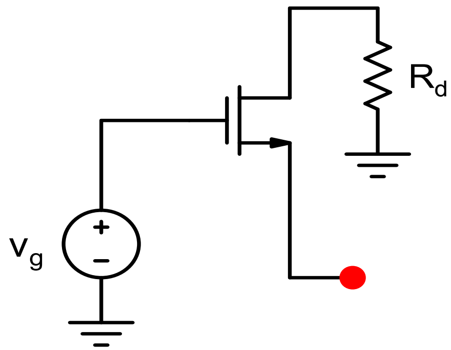
G-Number

Cascode Amplifier









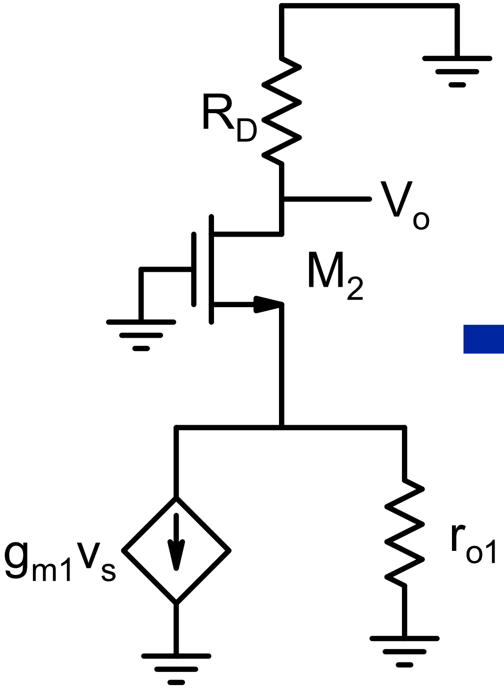
$$i_n \cong \frac{g_m v_g}{1+(g_m+g_{mb})R_S}$$

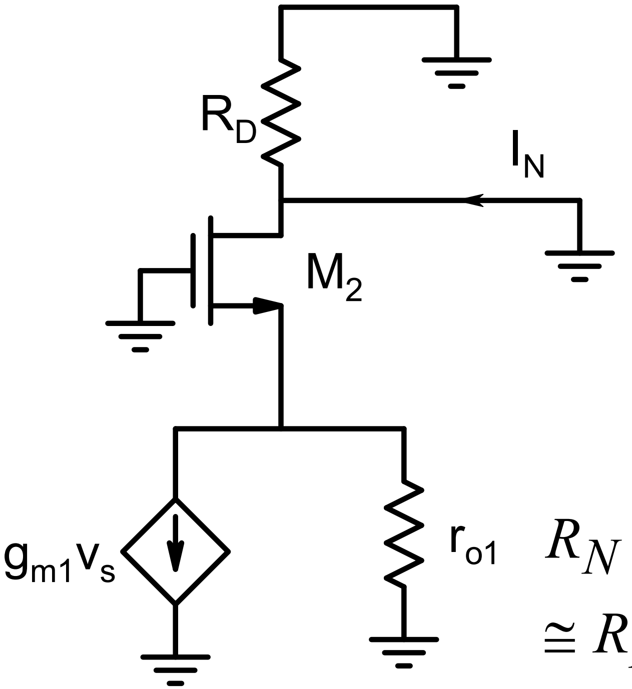
$$R_n \cong r_o \times \{1+(g_m+g_{mb})R_S\}$$

$$i_n = -\frac{g_m v_g}{1+R_d/r_o} \qquad R_n \cong \frac{1+R_d/r_o}{g_m+g_{mb}}$$

G-Number

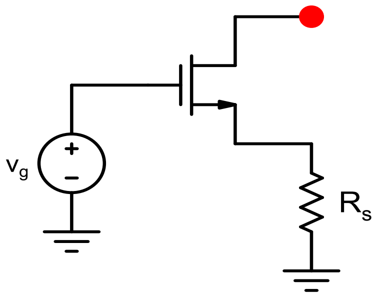
Cascode Amplifier





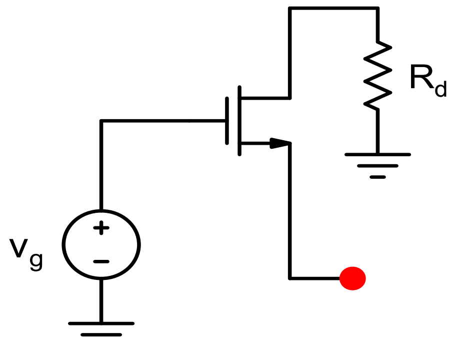
$$i_N \cong g_{m1} v_s$$

$$R_N = R_D \| r_{o2}(1 + g_{m2} r_{o1})$$
$$\cong R_D$$



$$i_n \cong \frac{g_m v_g}{1 + (g_m + g_{mb}) R_S}$$

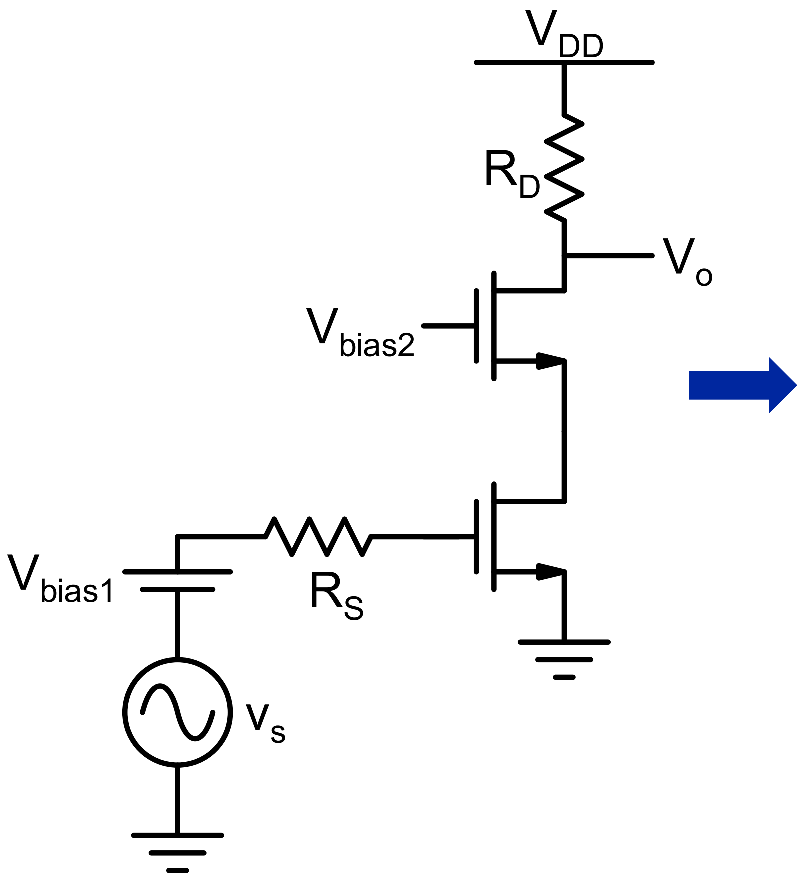
$$R_n \cong r_o \times \{1 + (g_m + g_{mb}) R_S\}$$

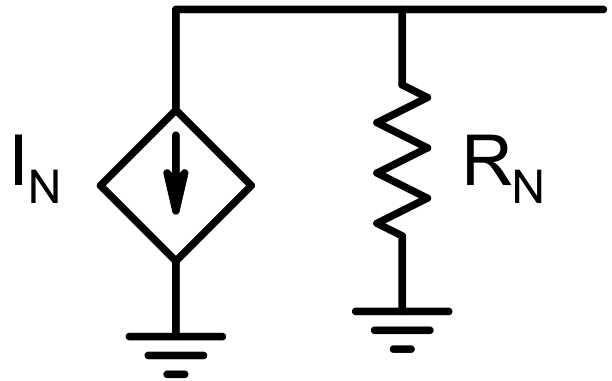


$$i_n = -\frac{g_m v_g}{1 + R_d / r_o} \qquad R_n \cong \frac{1 + R_d / r_o}{g_m + g_{mb}}$$

G-Number

Cascode Amplifier

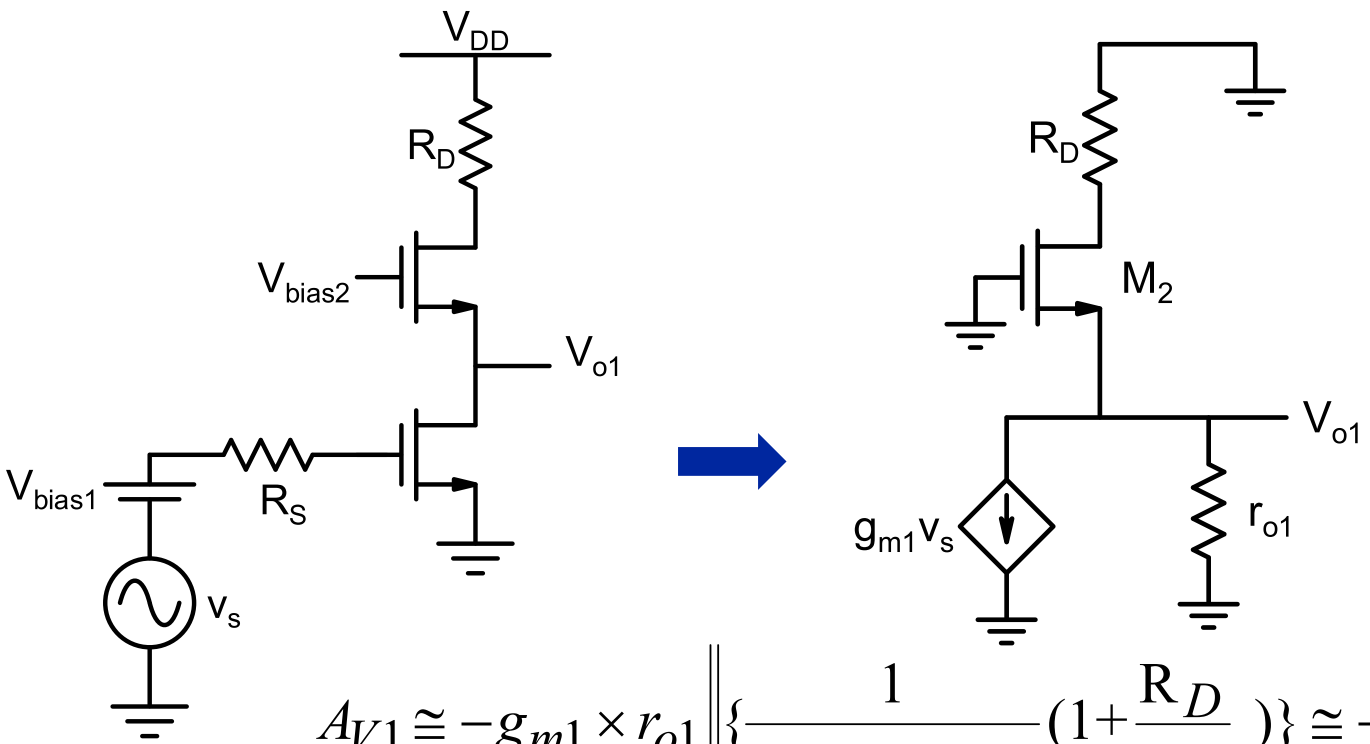




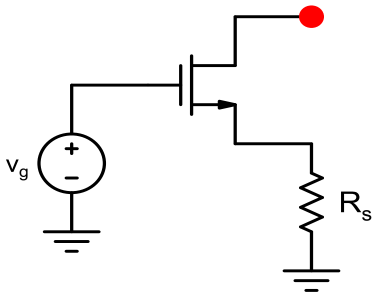
$$i_N \cong g_{m1} v_s \qquad R_N \cong R_D$$

$$A_V \cong -g_{m1} R_D \text{ just like a CS amplifier}$$

$$R_O \cong R_D$$

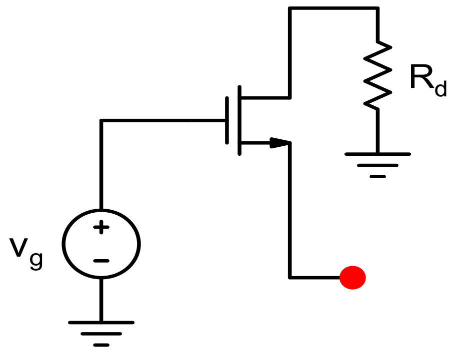


$$A_{V1} \cong -g_{m1} \times r_{o1} \,\Big\|\, \left\{ \frac{1}{g_{m2}+g_{mb2}} \left(1+\frac{R_D}{r_{o2}}\right) \right\} \cong -\frac{g_{m1}}{g_{m2}+g_{mb2}}$$

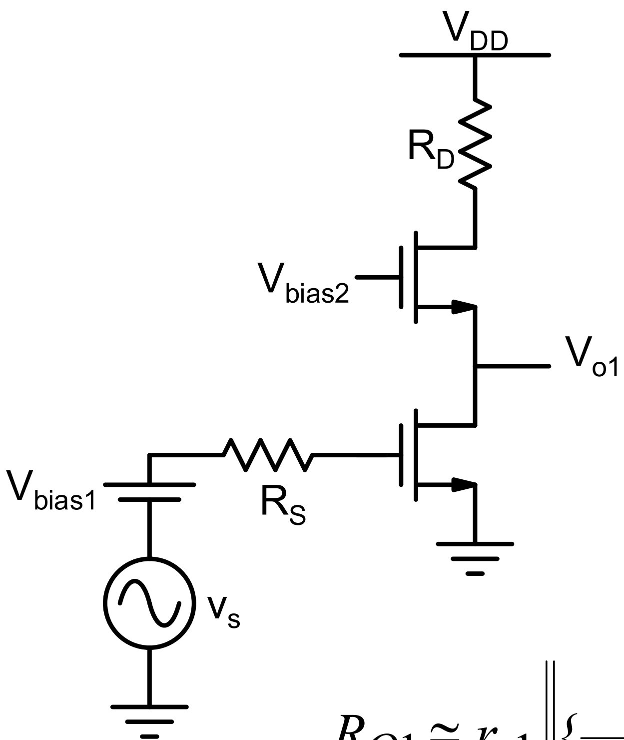


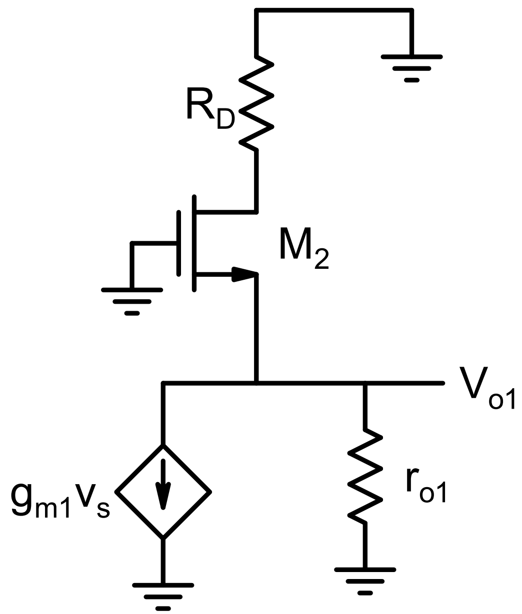
$$i_n \cong \frac{g_m v_g}{1+(g_m+g_{mb})R_S}$$

$$R_n \cong r_o \times \{1+(g_m+g_{mb})R_S\}$$

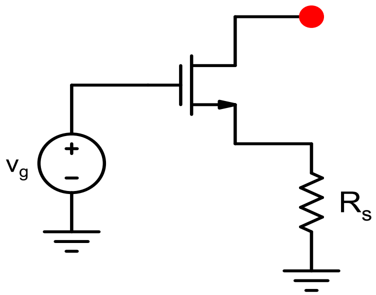


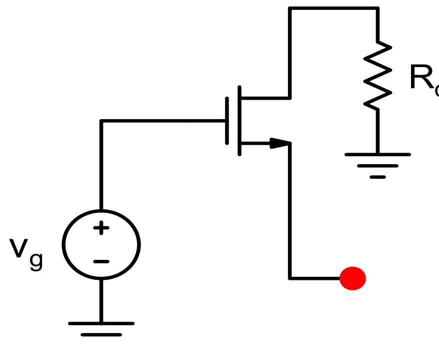
$$i_n = -\frac{g_m v_g}{1+R_d/r_o} \qquad R_n \cong \frac{1+R_d/r_o}{g_m+g_{mb}}$$





$$R_{O1} \cong r_{o1} \left\| \left\{ \frac{1}{g_{m2}+g_{mb2}} \left(1+\frac{R_D}{r_{o2}}\right) \right\} \right. \cong \frac{1}{g_{m2}+g_{mb2}}$$





$$i_n \cong \frac{g_m v_g}{1+(g_m+g_{mb})R_S}$$

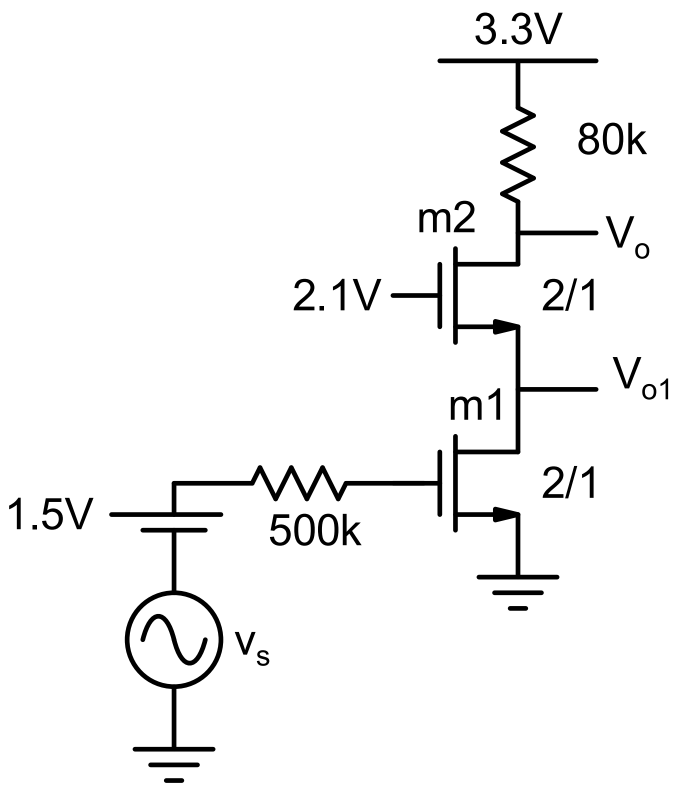
$$R_n \cong r_o \times \{1+(g_m+g_{mb})R_S\}$$

$$i_n = -\frac{g_m v_g}{1+R_d/r_o}$$

$$R_n \cong \frac{1+R_d/r_o}{g_m+g_{mb}}$$

G-Number

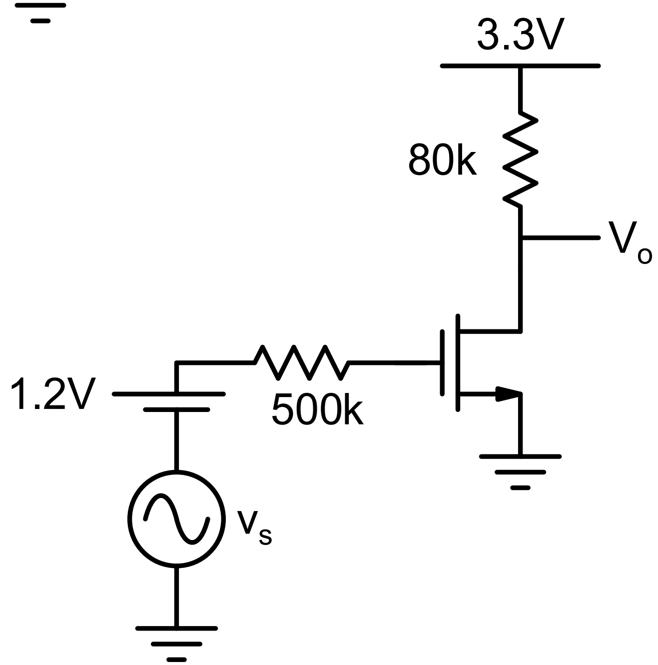
Example



$$I_{DSQ} = 25\mu A \qquad V_{sat1} = 0.5V$$

$$V_{o1}(dc) = 0.6V \qquad V_o(dc) = 1.3V$$

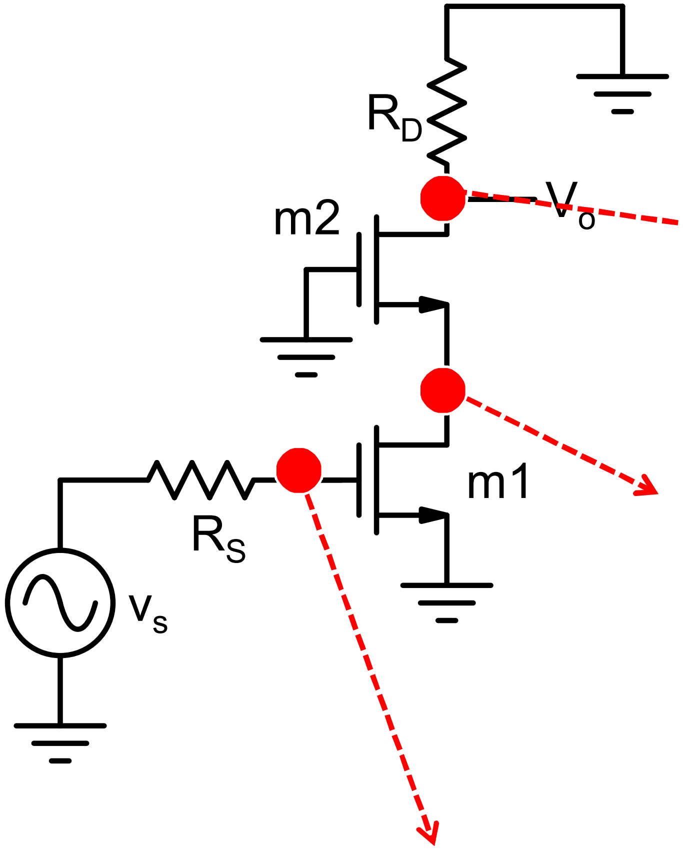
$$A_V = \frac{v_o}{v_s} = -8 \qquad A_{V1} = \frac{v_{o1}}{v_s} = -0.77$$



$$A_V = \frac{v_o}{v_s} = -8$$

G-Number

Cascode Amplifier: Frequency Response



$$R_3 = R_D$$
$$C_3 = C_{gd} + C_{db}$$

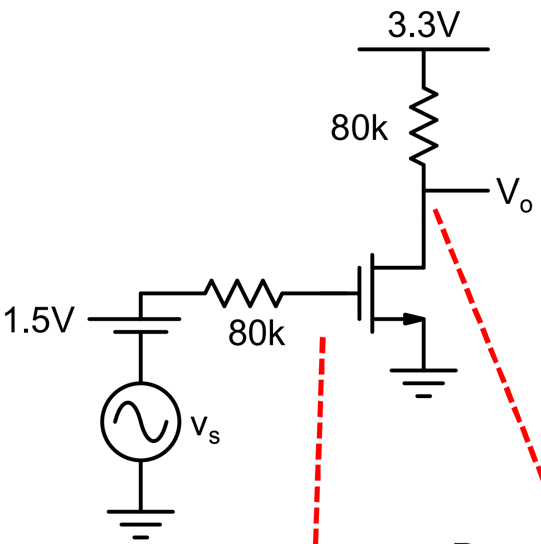
$$R_2 = r_{o1} \| \{ \frac{1}{g_{m2} + g_{mb2}} (1 + \frac{R_D}{r_{o2}}) \}$$

$$C_2 = C_{gd1}(1 - \frac{1}{A_{v1}}) + C_{gs2} + C_{db1} + C_{sb2}$$

$$R_1 = R_S$$
$$C_1 = C_{gs1} + C_{gd1}(1 - A_{v1})$$

$$\boxed{f_{3dB} \cong \frac{1}{2\pi \sum R_j C_j}}$$

Comparison-1

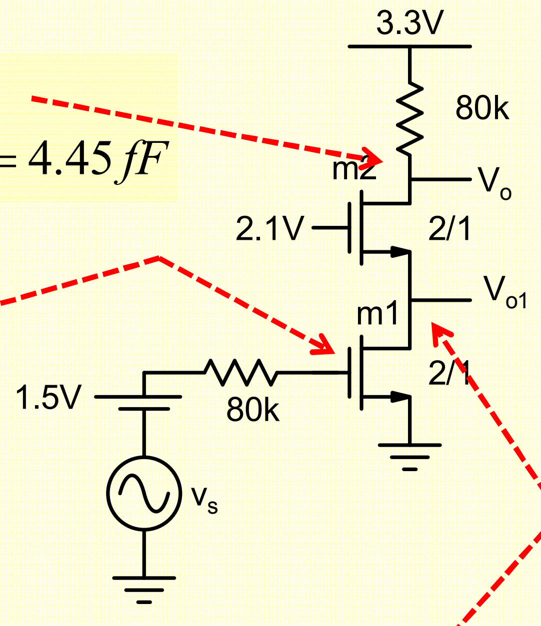


$$R_3 = R_D = 80k$$
$$C_3 = C_{gd} + C_{db} = 4.45\,fF$$

$$R_1 = R_S$$
$$C_1 = C_{gs1} + C_{gd1}(1 - A_v)$$
$$= 4 + 3.6 = 7.6\,fF$$

$$f_{3dB} \cong \frac{1}{2\pi \sum R_j C_j} = 166MHz$$

$$R_3 = R_D = 80k$$
$$C_3 = C_{gd} + C_{db} = 4.45\,fF$$



$$R_1 = R_S$$
$$C_1 = C_{gs1} + C_{gd1}(1 - A_{v1})$$
$$= 4 + 0.8 = 4.8\,fF$$

$$R_2 = r_{o1} \Big\| \{\frac{1}{g_{m2} + g_{mb2}}(1 + \frac{R_D}{r_{o2}})\} = 10k$$

$$C_2 = C_{gd1}(1 - \frac{1}{A_{v1}}) + C_{gs2} + C_{db1} = 12.8\,fF$$

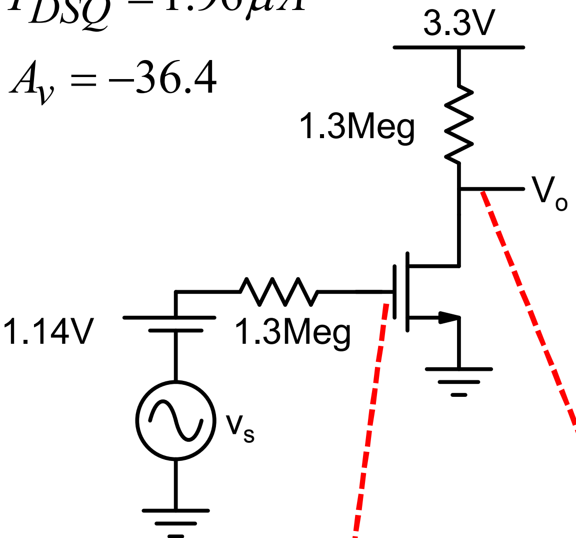
$$f_{3dB} \cong \frac{1}{2\pi \sum R_j C_j} = 183MHz$$

Role of C_{gd} is less due to small gain and small value of R_S

Comparison-2

$$I_{DSQ} = 1.96\mu A$$

$$A_v = -36.4$$



$$R_3 = R_D = 1300k$$

$$C_3 = C_{gd} + C_{db} = 4.4\,fF$$

$$R_1 = R_S$$

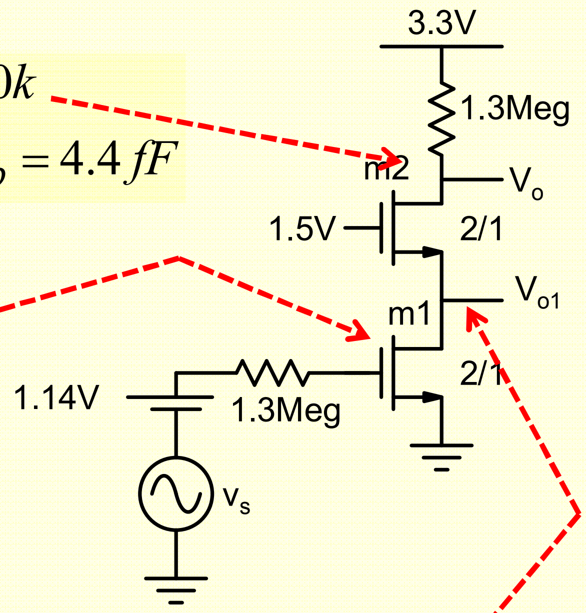
$$C_1 = C_{gs1} + C_{gd1}(1 - A_v)$$

$$= 4 + 15 = 19\,fF$$

$$f_{3dB} \cong \frac{1}{2\pi \sum R_j C_j} = 5.24MHz$$

$$R_3 = R_D = 1300k$$

$$C_3 = C_{gd} + C_{db} = 4.4\,fF$$



$$R_1 = R_S$$

$$C_1 = C_{gs1} + C_{gd1}(1 - A_{v1})$$

$$= 4 + 0.8 = 4.8\,fF$$

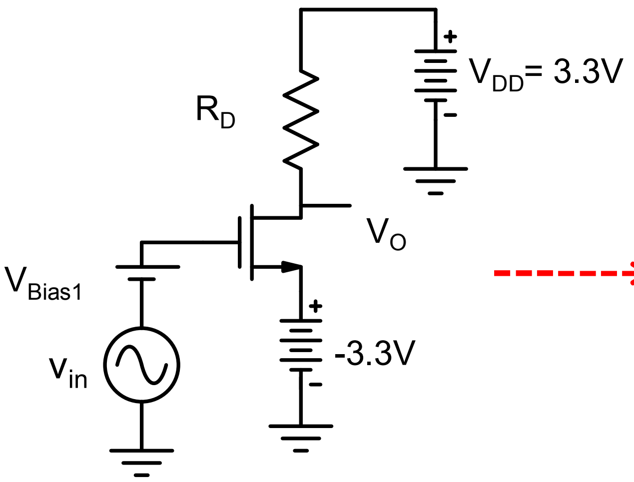
$$R_2 = r_{o1} \Big\| \{\frac{1}{g_{m2} + g_{mb2}}(1 + \frac{R_D}{r_{o2}})\} = 35.7\text{k}$$

$$C_2 = C_{gd1}(1 - \frac{1}{A_{v1}}) + C_{gs2} + C_{db1} + C_{sb2} = 12.8\,fF$$

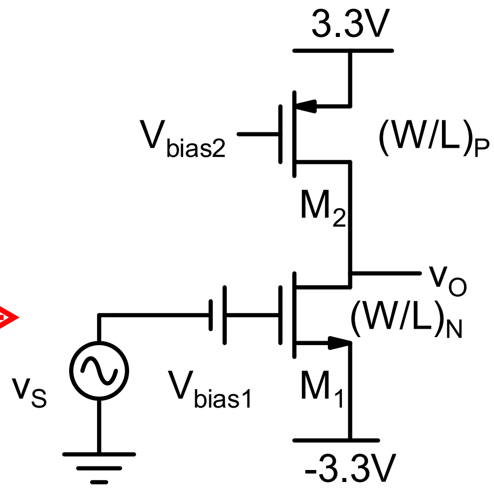
$$f_{3dB} \cong \frac{1}{2\pi \sum R_j C_j} = 12.8MHz$$

Body effect is ignored

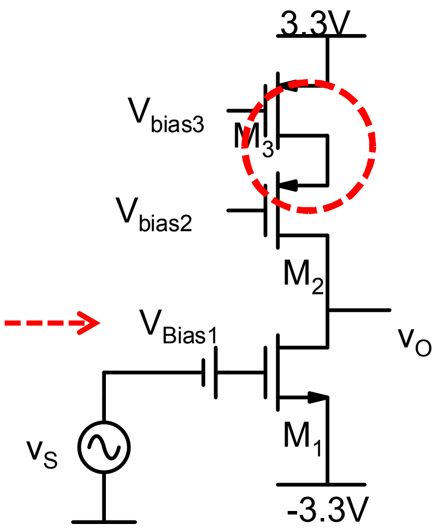
G-Number



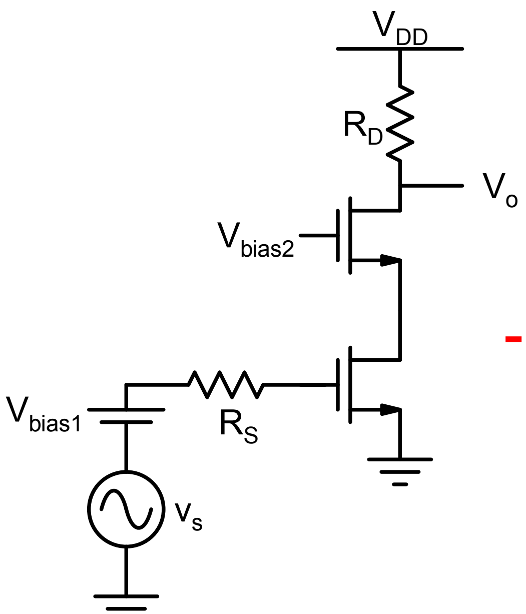
CS amplifier



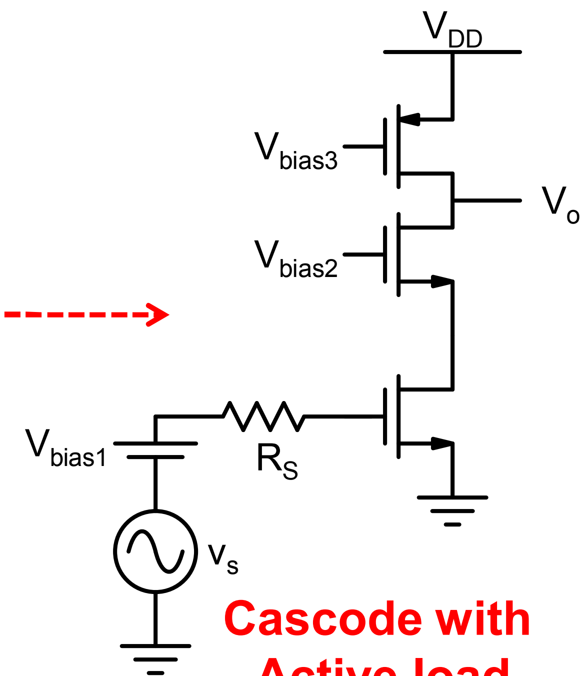
CS with Active Load



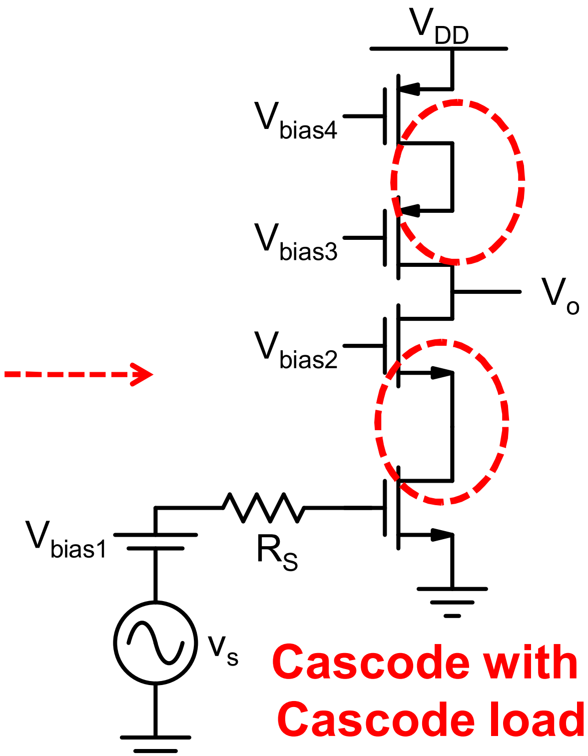
CS with Cascode Load



Cascode with Resistive load



Cascode with Active load



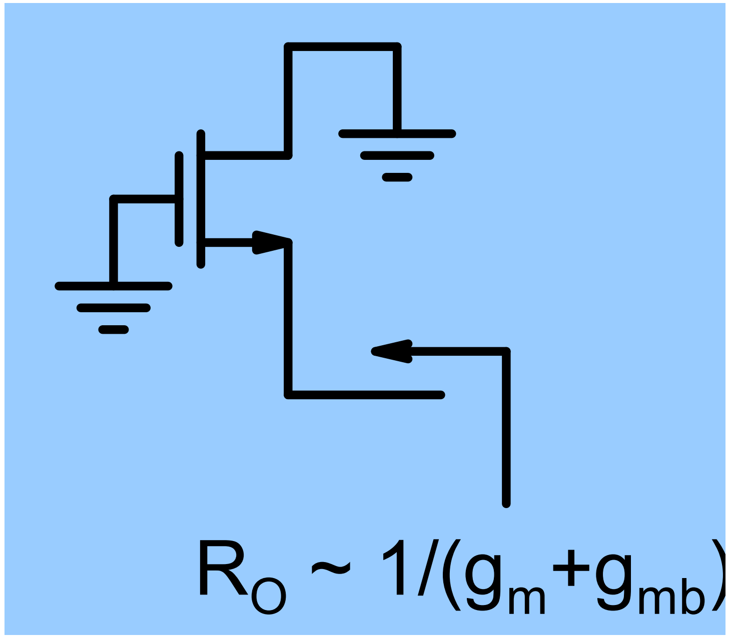
Cascode with Cascode load

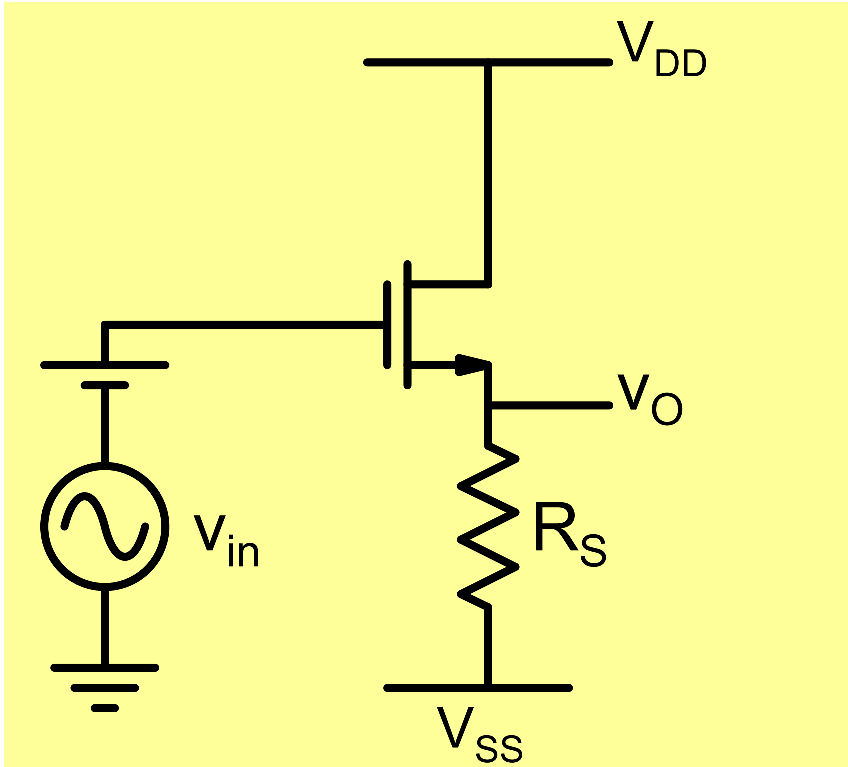
G-Number

Common Drain Amplifier

Amplifier with Low output resistance

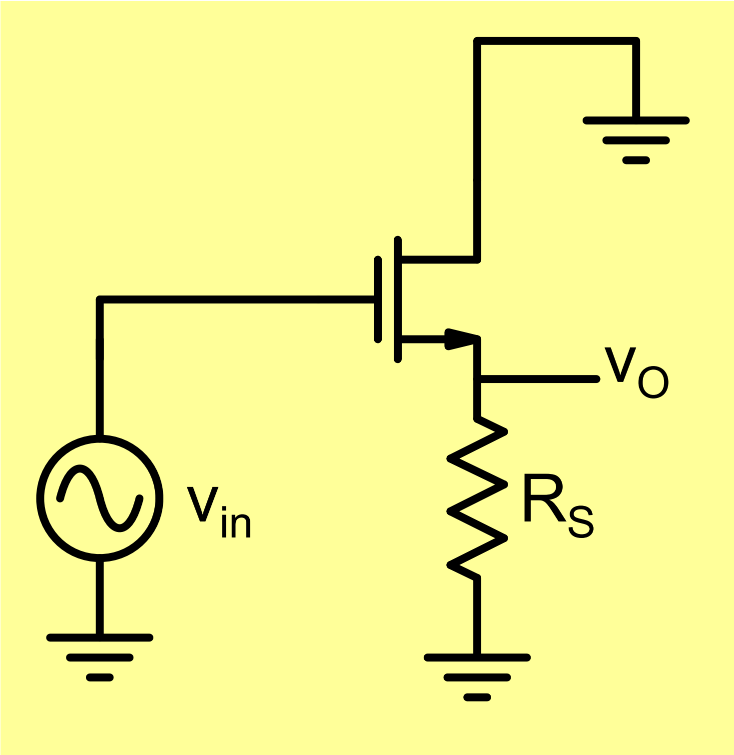
Resistance looking into source is small !

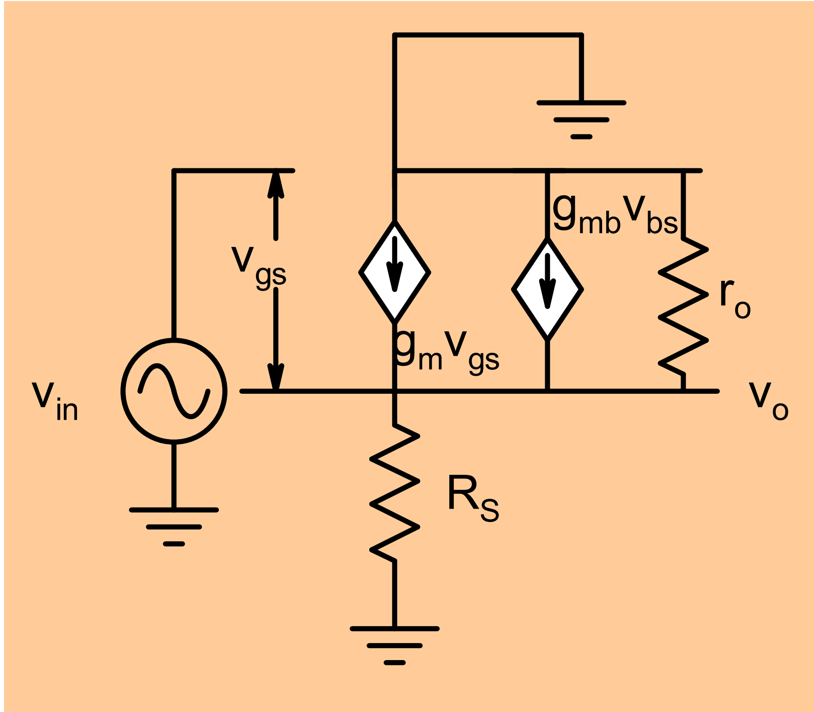




Apply input at gate, take output at Source

Small Signal Analysis





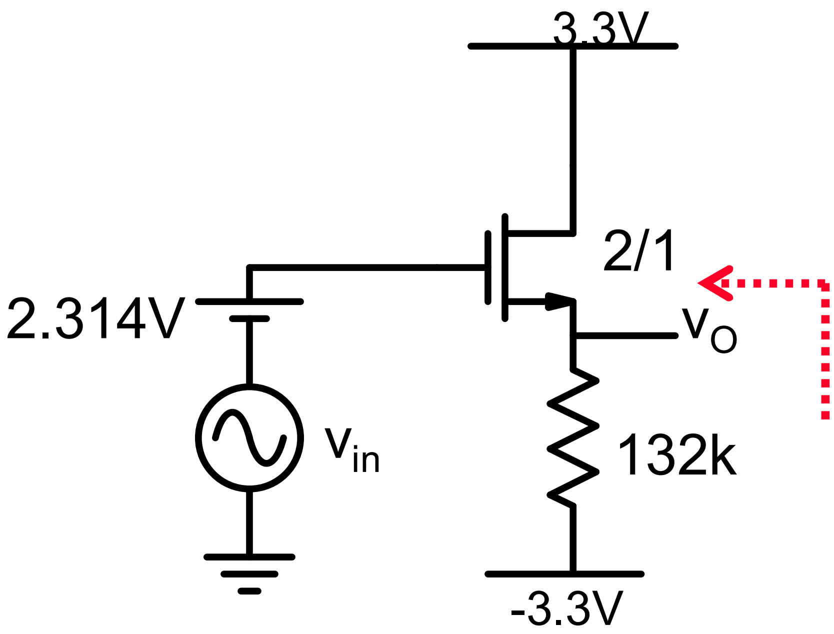
$$A_v = \frac{g_m R_S \| r_o}{1 + (g_m + g_{mb}) R_S \| r_o}$$

$$R_o = R_S \left\| \frac{1}{g_m + g_{mb}} \cong \frac{1}{g_m + g_{mb}}\right.$$

Gain is less than unity !

Output resistance is low !

Example:



$$V_{Bias} = V_{GSQ} + I_{DSQ}R_S + V_{SS}$$

$$I_{DSQ} = \frac{\beta_N}{2}(V_{GSQ} - V_{THN})^2[1 + \lambda_n V_{DSQ}]$$

$$V_{THN} = V_{THN0} + \gamma(\sqrt{2\phi_F + V_{SBQ}} - \sqrt{2\phi_F})$$

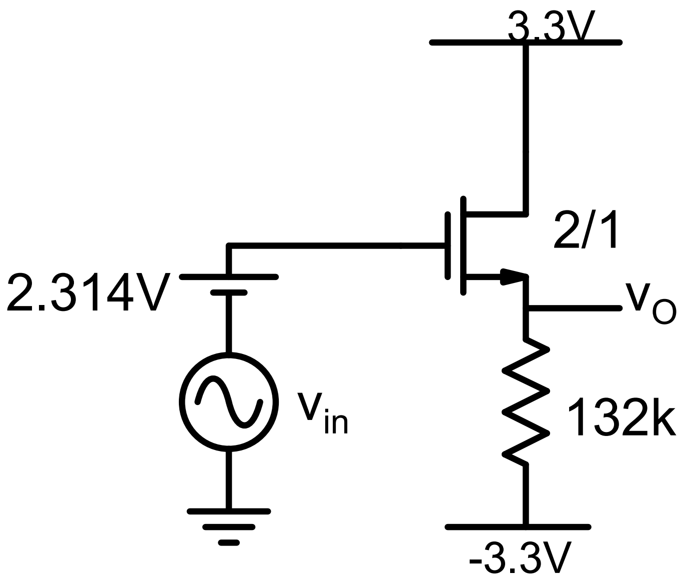
$$V_{SBQ} = I_{DSQ}R_S \ ; \ V_{DSQ} = V_{DD} - I_{DSQ}R_S - V_{SS}$$

$I_{DSQ} = 25\mu A$ $\quad V_{SBQ} = 3.3V \Rightarrow V_{THN} = 1.8V ; V_{GSQ} = 2.314V$ $\quad V_{DSQ} = 3.3V$

$$g_m = 100\mu A/V \ ; \ g_{mb} = 17.5\mu A/V ; \ r_o = 4\text{M}\Omega$$

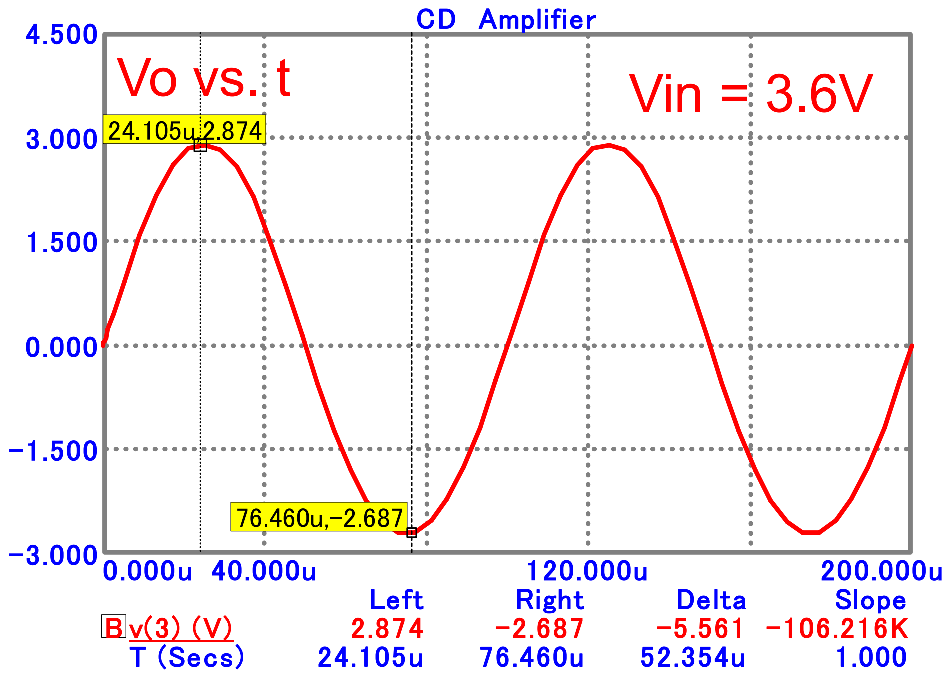
$$A_v = \frac{g_m R_S}{1 + (g_m + g_{mb})R_S} = 0.8 \qquad R_O = R_S \left\| \frac{1}{g_m + g_{mb}} \right. \sim 8k$$

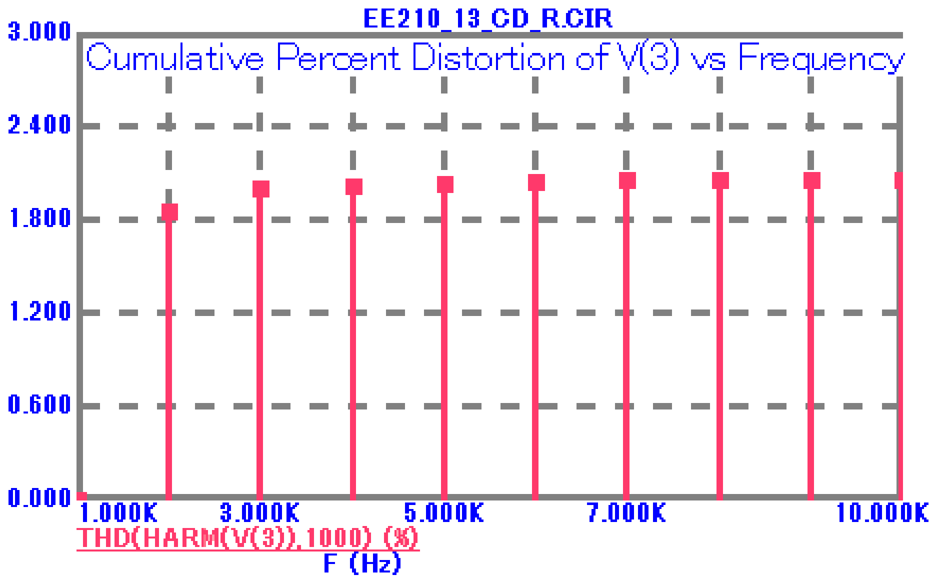
G-Number



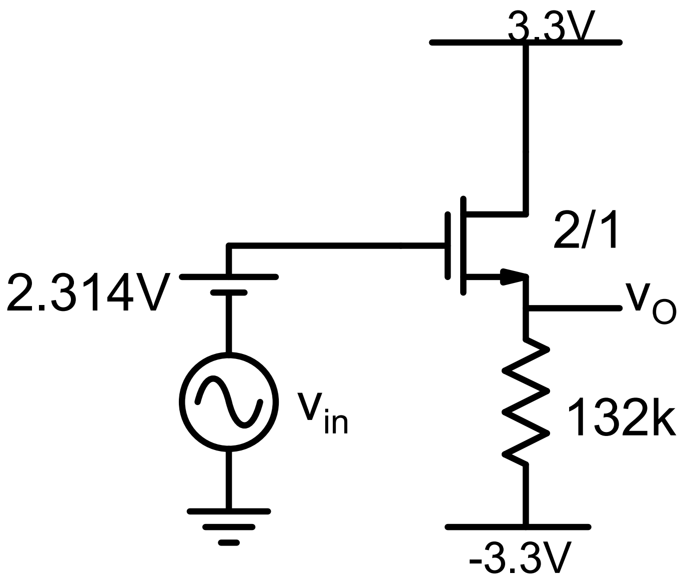
CD amplifier has good linearity and thus less prone to harmonic distortion

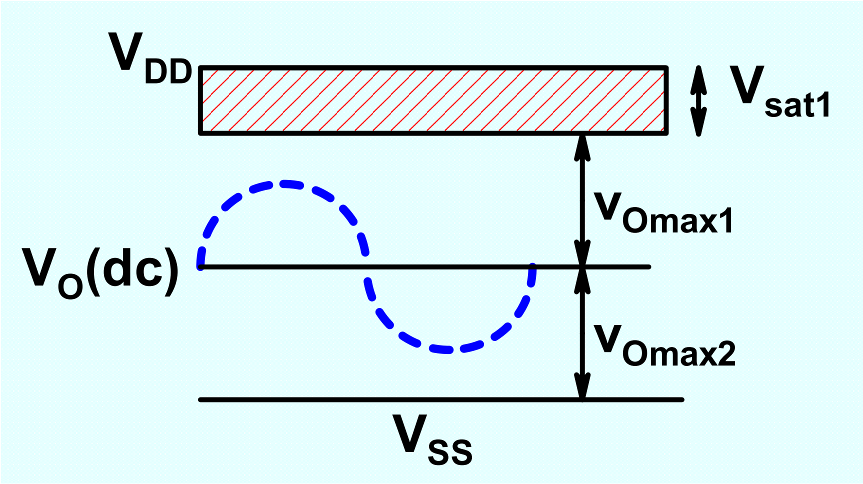
Vin = 3.6V

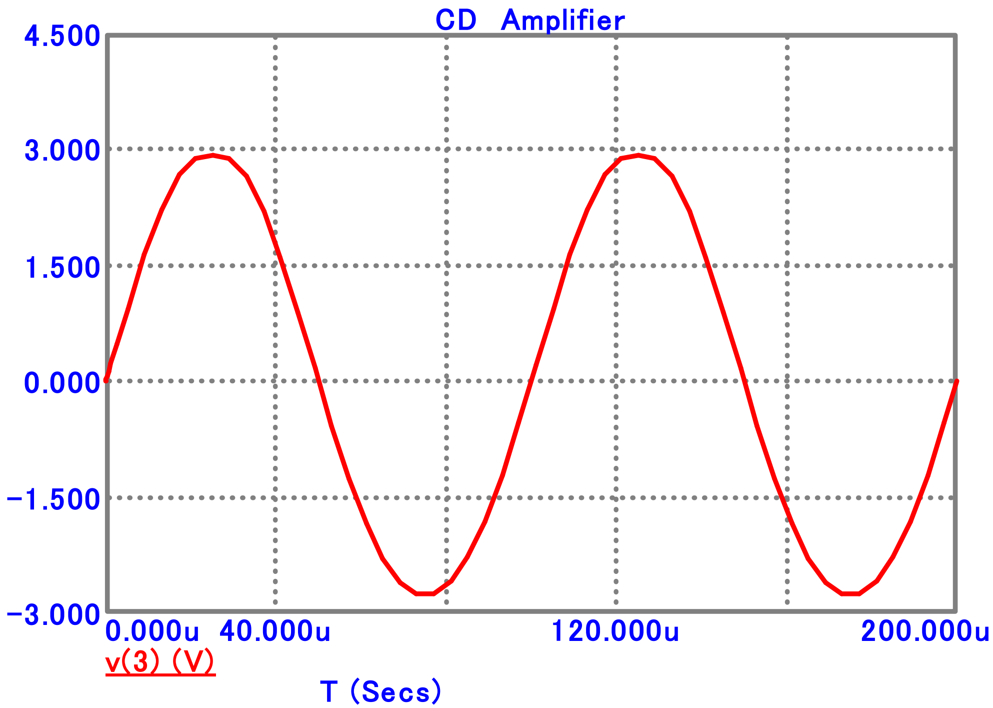




G-Number

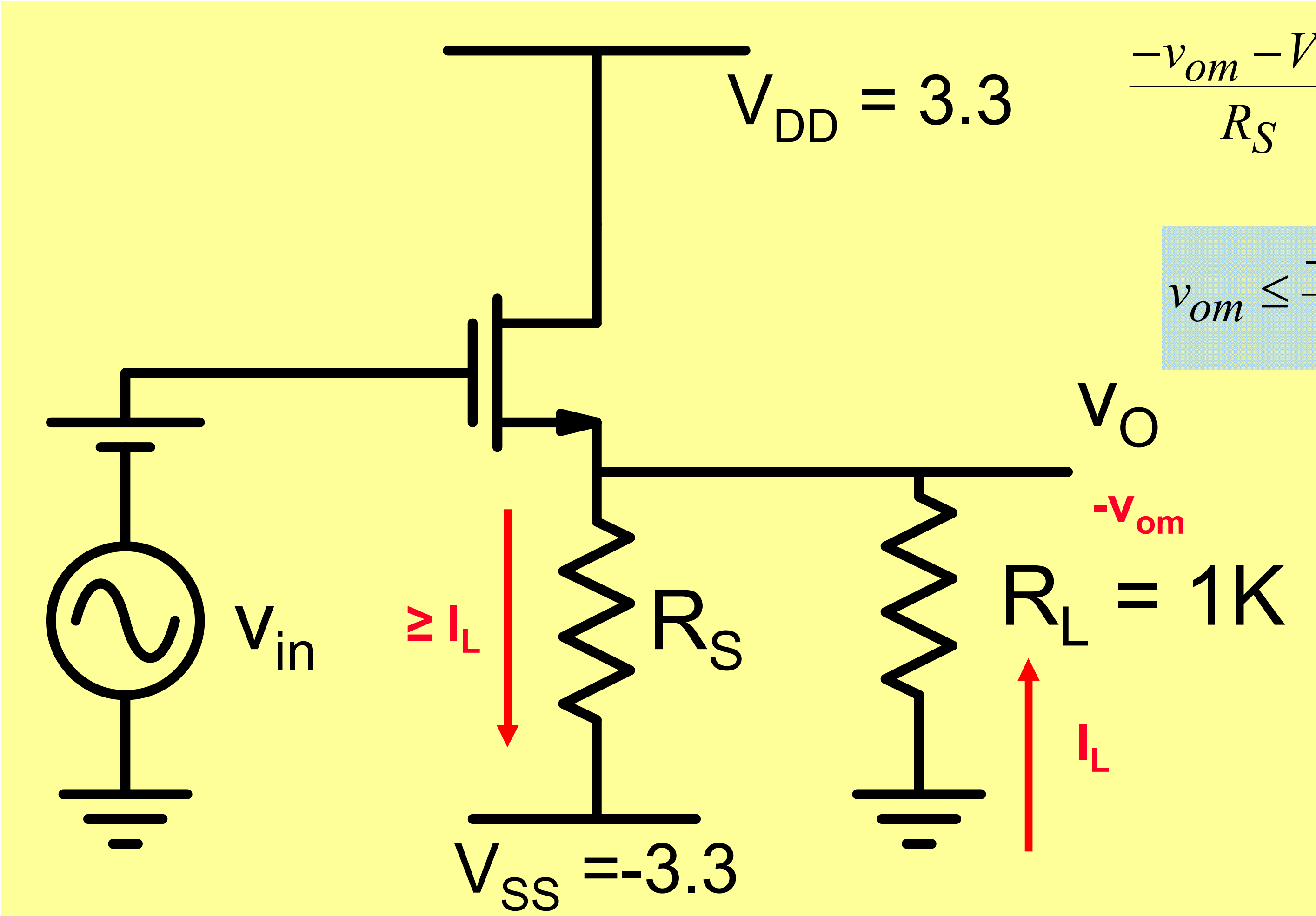






Good swing but input sinusoidal amplitude is 3.7V

Swing limited by Output current drive

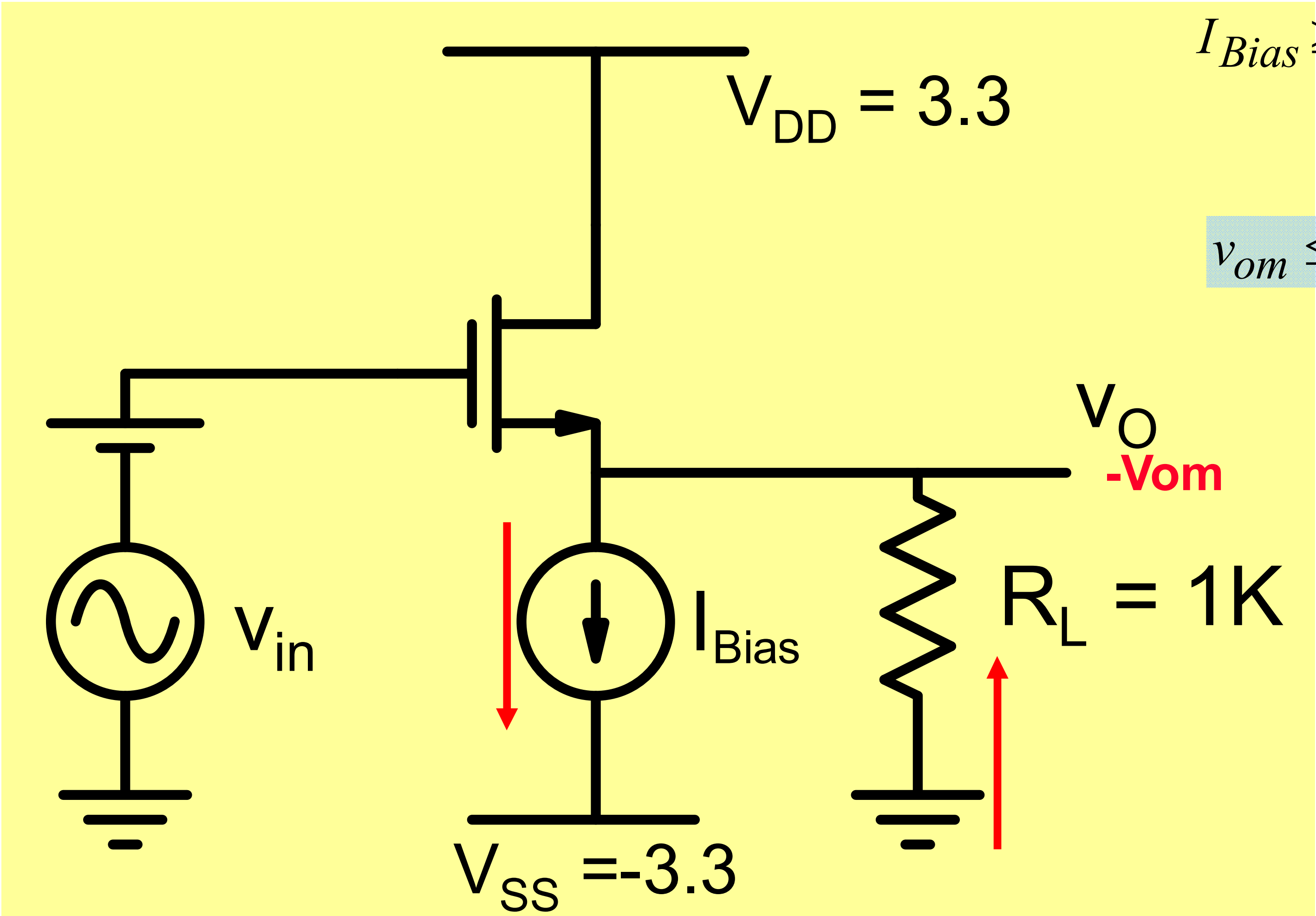


$$\frac{-v_{om} - V_{SS}}{R_S} \geq I_L = \frac{v_{om}}{R_L}$$

$$v_{om} \leq \frac{-V_{SS}}{R_S} \times R_S \| R_L$$

$$I_{DSQ} = \frac{-V_{SS}}{R_S} \geq \frac{v_{om}}{R_S \| R_L}$$

CD amplifier with current source biasing



$$I_{Bias} \geq I_L = \frac{v_{om}}{R_L}$$

$$v_{om} \leq I_{Bias} \times R_L$$

EE210: Microelectronics-I

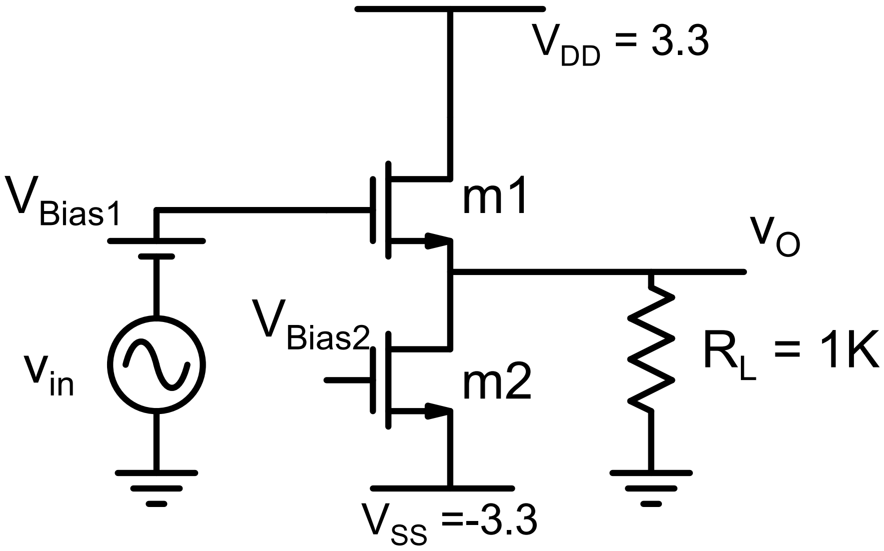
Lecture-44 : MOS Amplifiers_4

http://youtu.be/0Z670Vz_Too

B. Mazhari
Dept. of EE, IIT Kanpur

Example

$$I_{DSQ} = 3.3mA$$



$$\frac{W_1}{L_1} = \frac{200}{1}\ ;\ V_{GS1} = 2.389V ; V_O = 0V$$

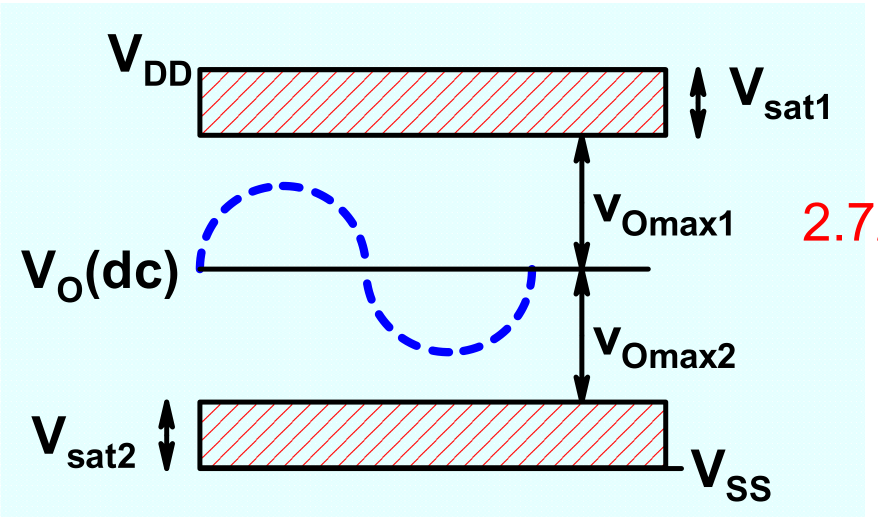
$$\frac{W_2}{L_2} = \frac{200}{1}\ ;\ V_{GS2} = 1.575V$$

$$\Rightarrow V_{bias2} = -1.725V$$

$$V_{sat} = 0.575V$$

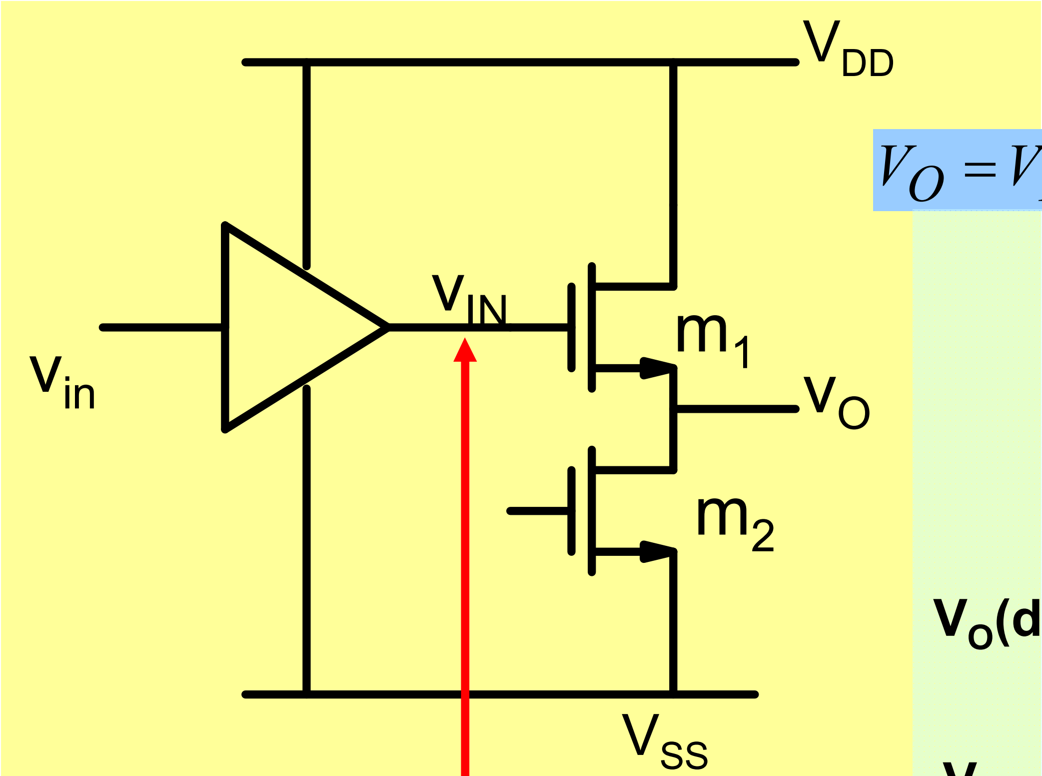
$$A_v = \frac{g_m R_L}{1 + (g_m + g_{mb}) R_L} = 0.79$$

$$R_o = \frac{1}{g_m + g_{mb}} \sim 74\Omega$$

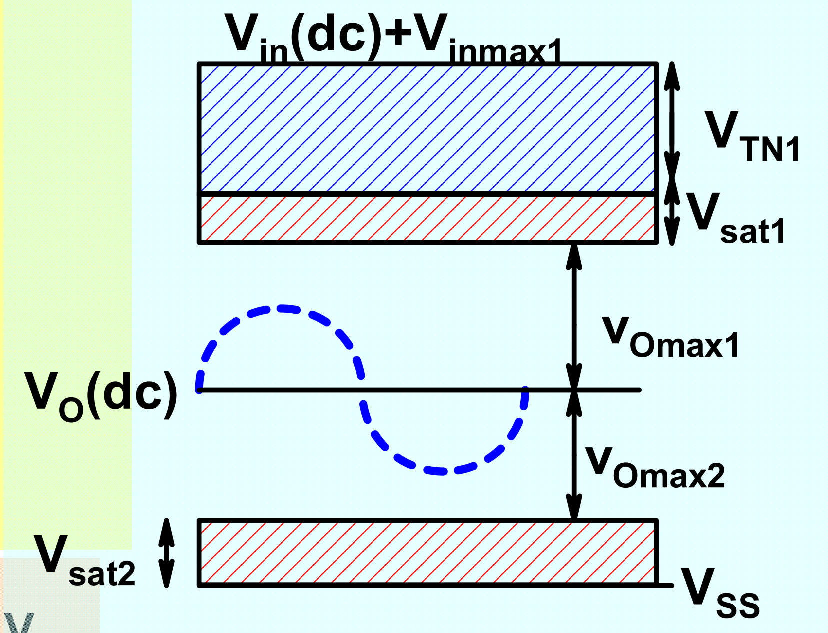


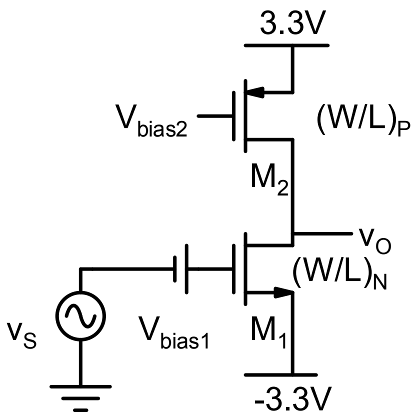
G-Number

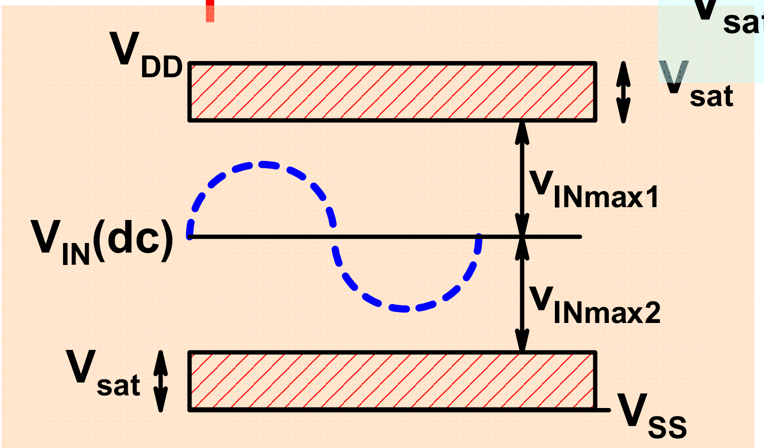
Voltage Swing: Limitations



$$V_O = V_{IN} - V_{GS1} = V_{IN} - V_{TN} - V_{sat1}$$

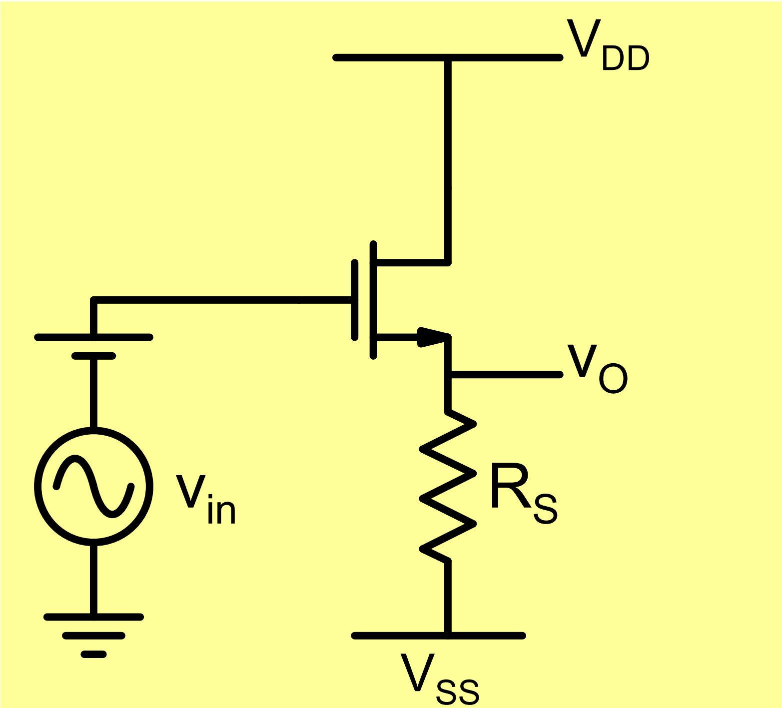


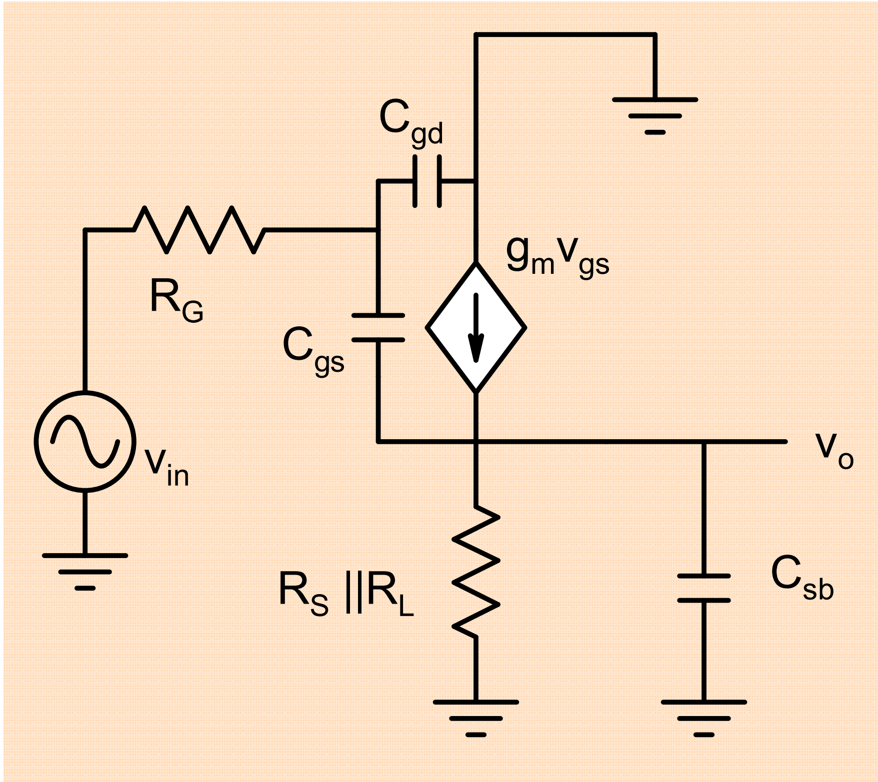




G-Number

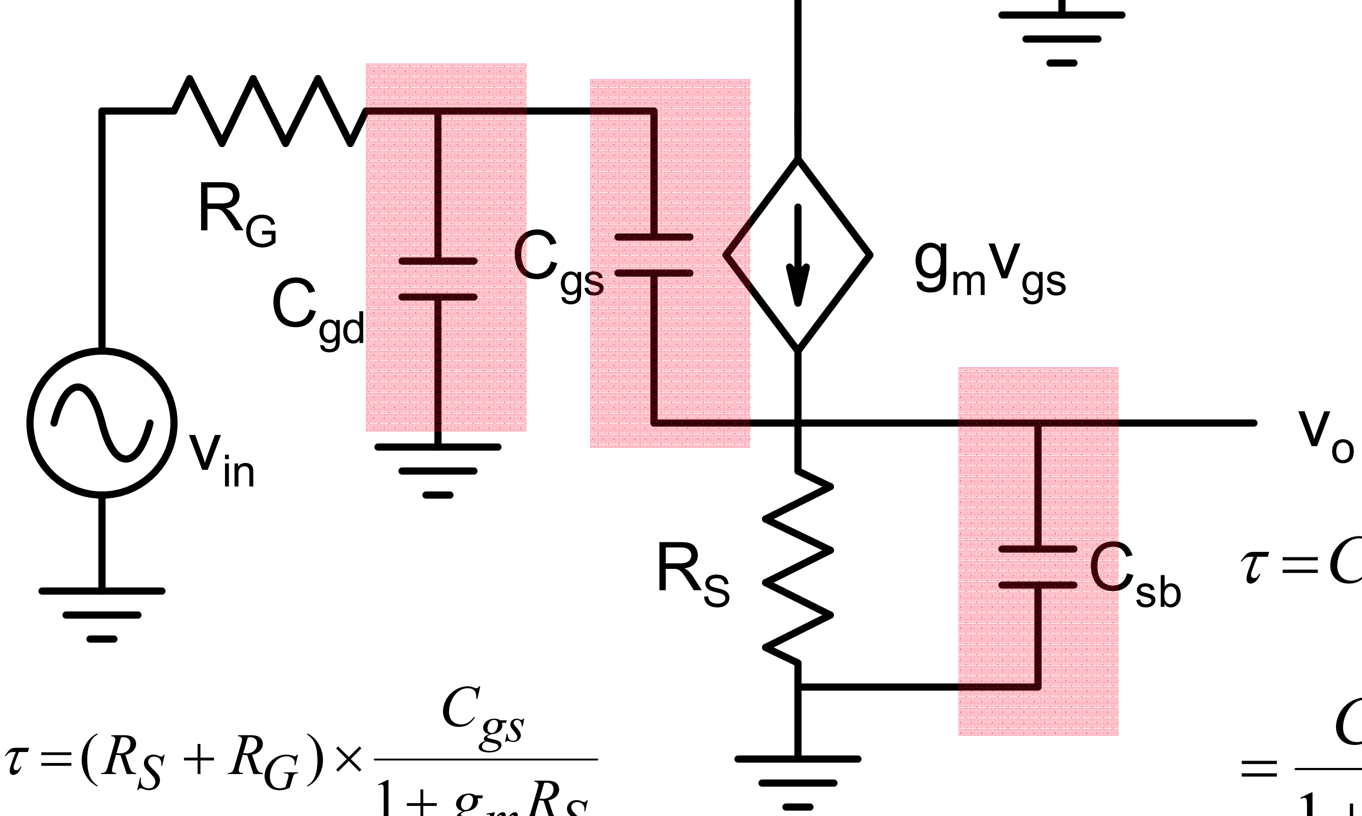
Frequency Response





$$f_{3dB} = \frac{1}{2\pi \sum \tau_j} \qquad \tau_j = R_j C_j$$

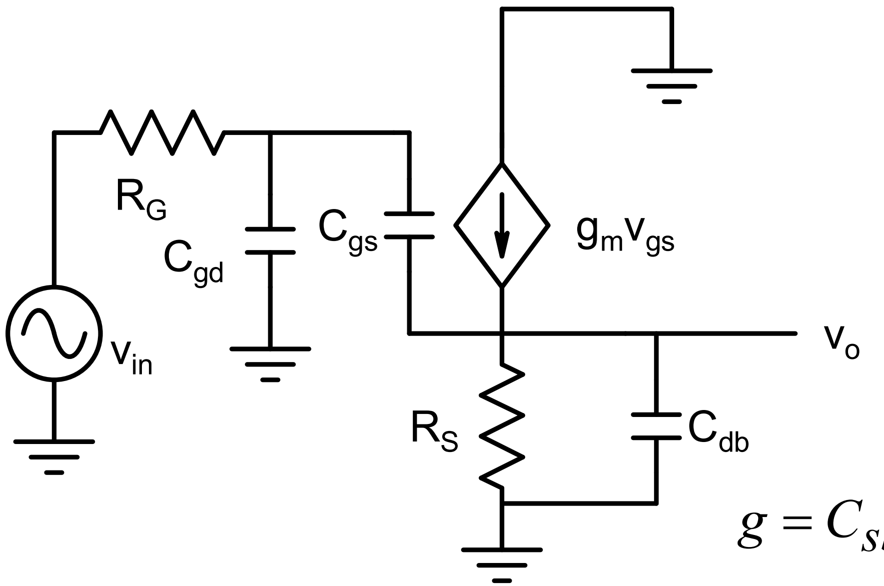
$$\tau = R_G C_{gd}$$



$$\tau = C_{sb} \times R_s \left\| \frac{1}{g_m} \right. = \frac{C_{sb} R_S}{1 + g_m R_S}$$

$$\tau = (R_S + R_G) \times \frac{C_{gs}}{1 + g_m R_S}$$

$$\omega_{3dB} \cong \left\{ R_G C_{gd} + (R_S + R_G) \times \frac{C_{gs}}{1 + g_m R_S} + \frac{R_S C_{sb}}{1 + g_m R_S} \right\}^{-1}$$



$$H(s) = K\frac{1+\frac{s}{z}}{1+gs+hs^2}$$

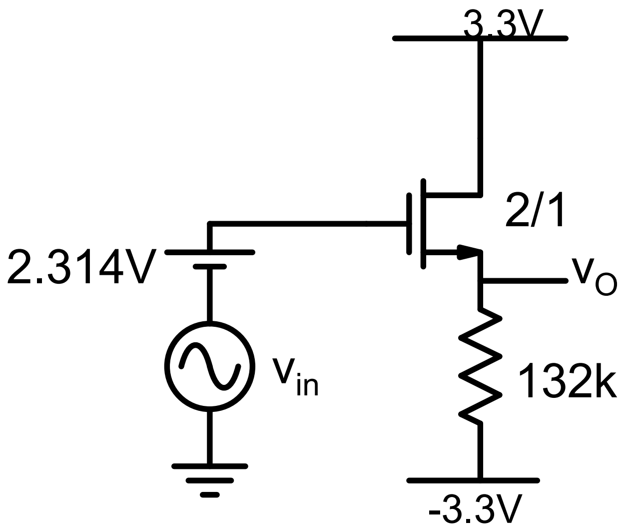
$$z = \frac{g_m}{C_{gs}}$$

$$g = C_{sb}\times\frac{R_s}{1+g_mR_s} + C_{gd}R_g + \frac{C_{gs}}{1+g_mR_S}\times(R_s+R_G)$$

$$h = \frac{R_gR_s}{1+g_mR_s}\times(C_{gd}C_{sb}+C_{sb}C_{gs}+C_{gs}C_{gd})$$

There are, in general, two poles and a zero, all of which will influence 3dB frequency if they are close together.

Case-1: Negligible R_G

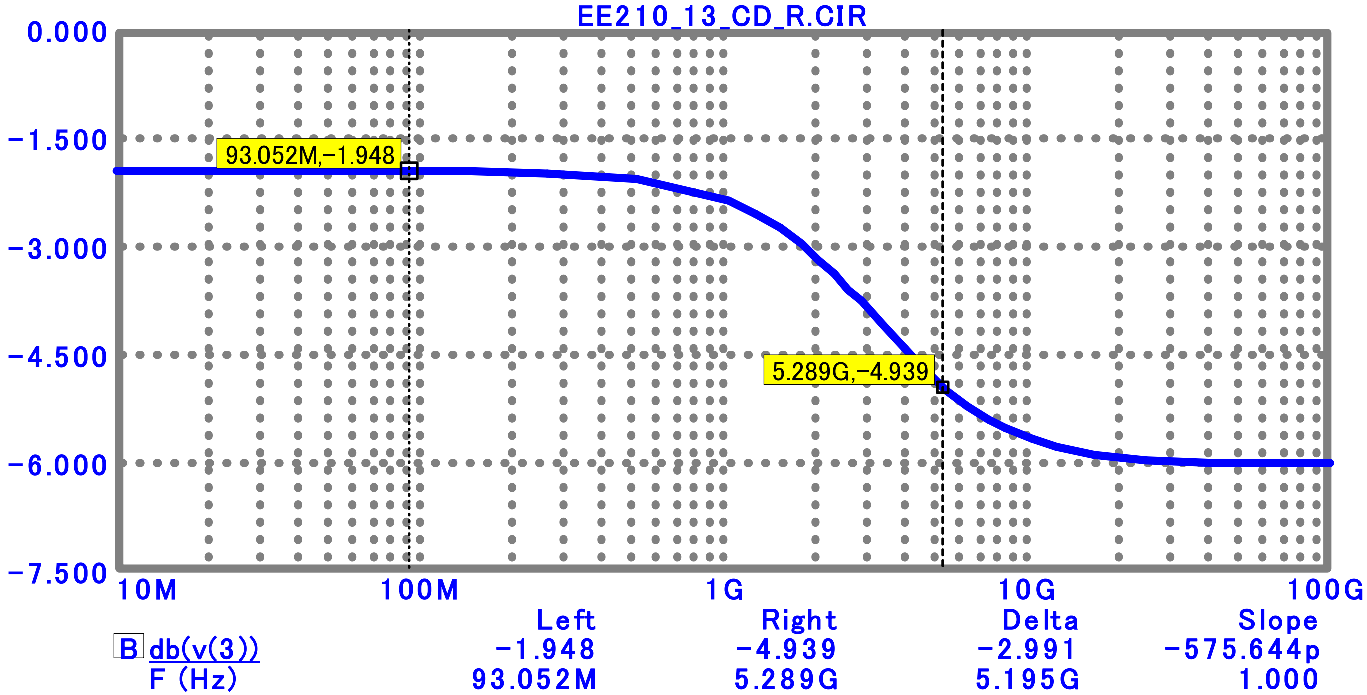


$$2\pi f_{3dB} \cong \{R_G C_{gd} + (R_S + R_G) \times \frac{C_{gs}}{1+g_m R_S} + \frac{R_S C_{db}}{1+g_m R_S}\}^{-1}$$

$$= 2.1\times 10^9\,Hz$$

There is only one pole in this case since h = 0

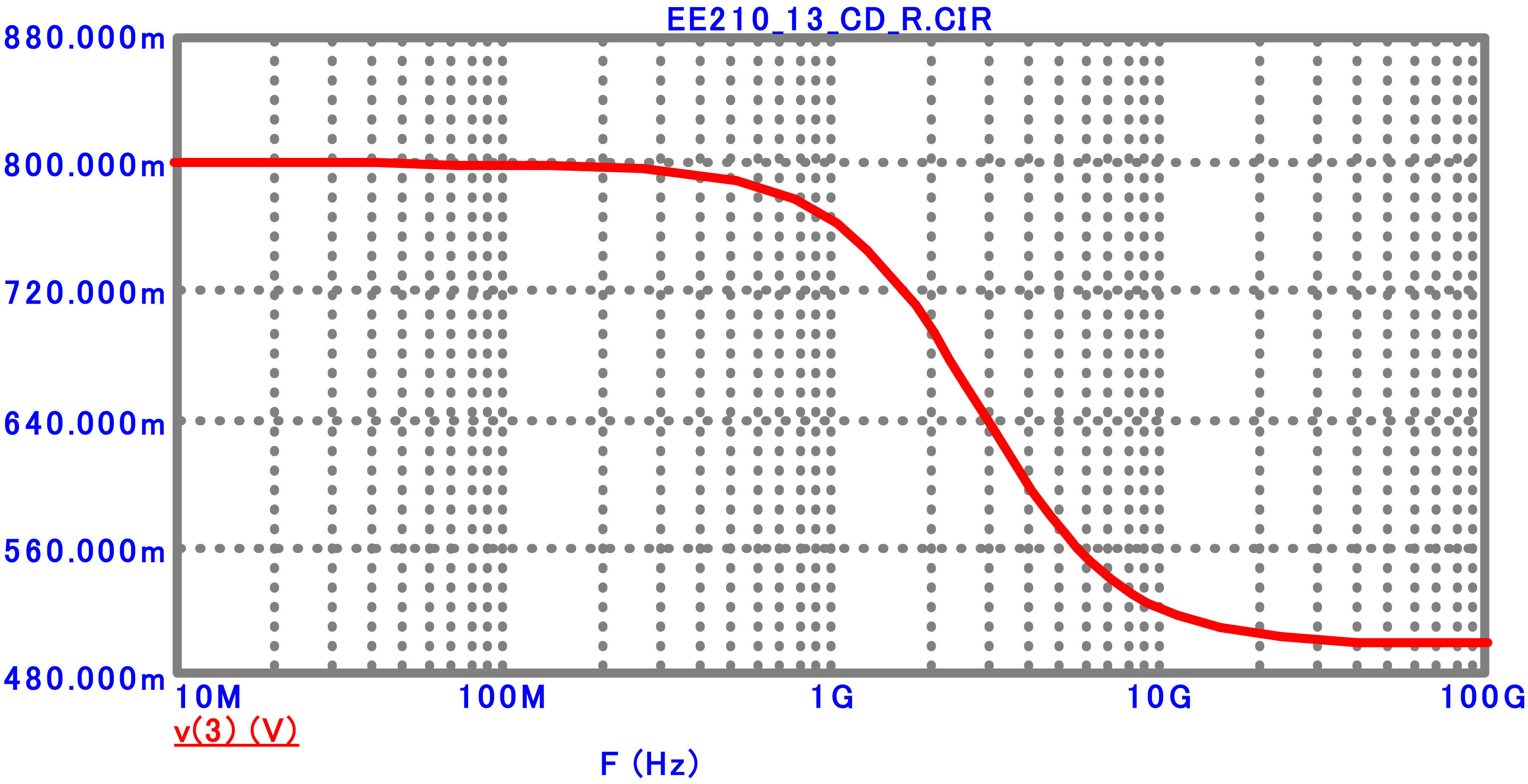
$$p_1 = 2.1\times 10^9\,Hz \qquad z = 3.97\times 10^9\,Hz$$



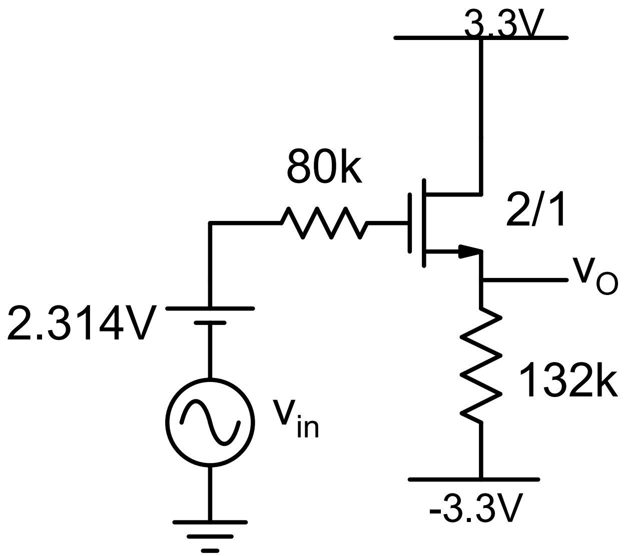
$$f_{3dB} = 5.3\times 10^9\,Hz$$

G-Number

Gain vs. frequency for RG = 0



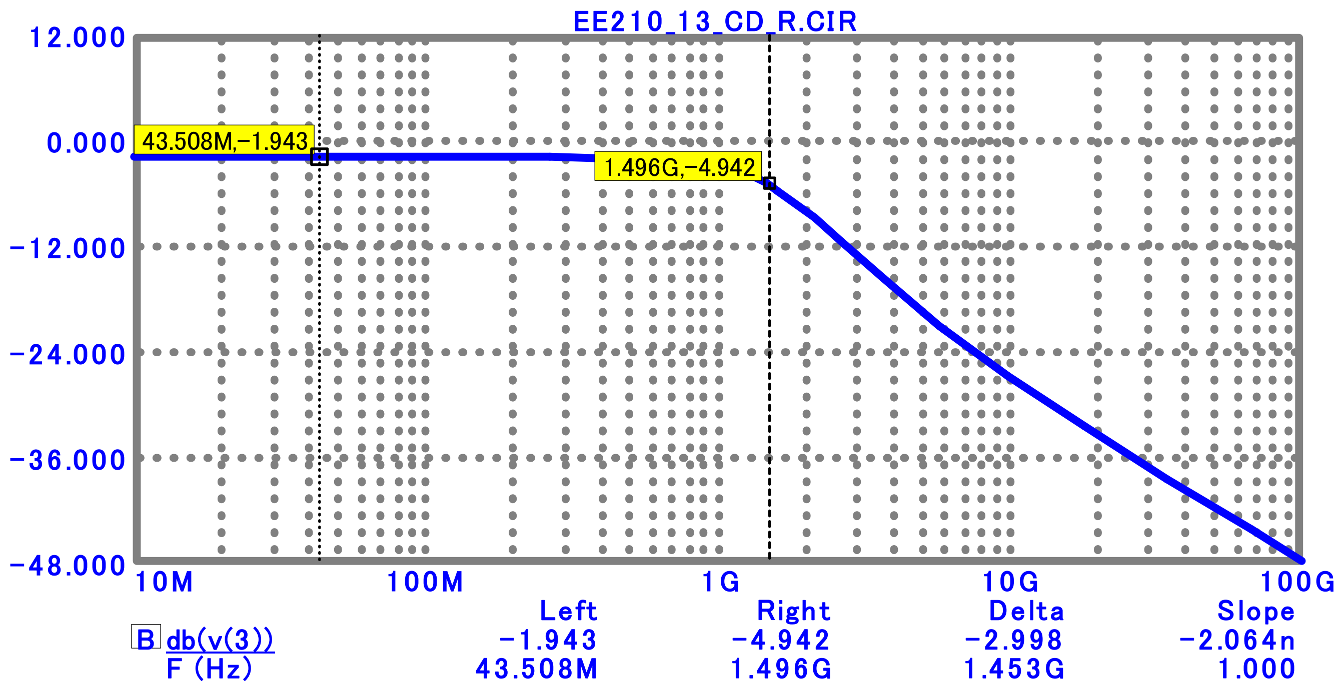
G-Number



$$2\pi f_{3dB} \cong \{R_G C_{gd} + (R_S + R_G)\times\frac{C_{gs}}{1+g_m R_S} + \frac{R_S C_{db}}{1+g_m R_S}\}^{-1}$$

$$= 1.23\times 10^9\,Hz$$

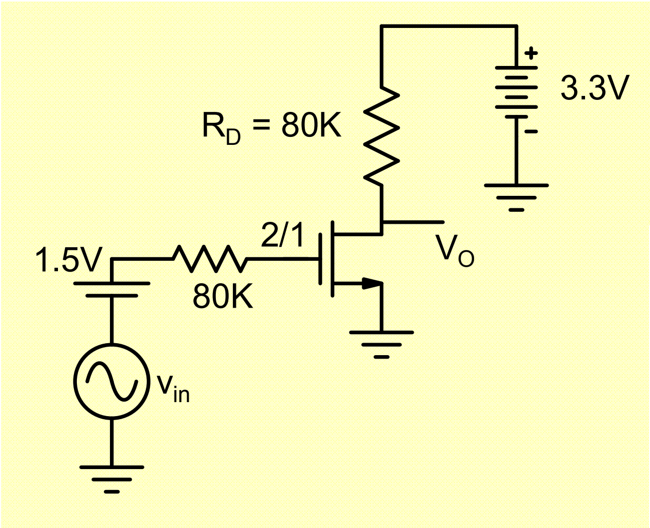
$$p_1 = 1.23\times 10^9\,Hz \qquad z = 3.97\times 10^9\,Hz$$



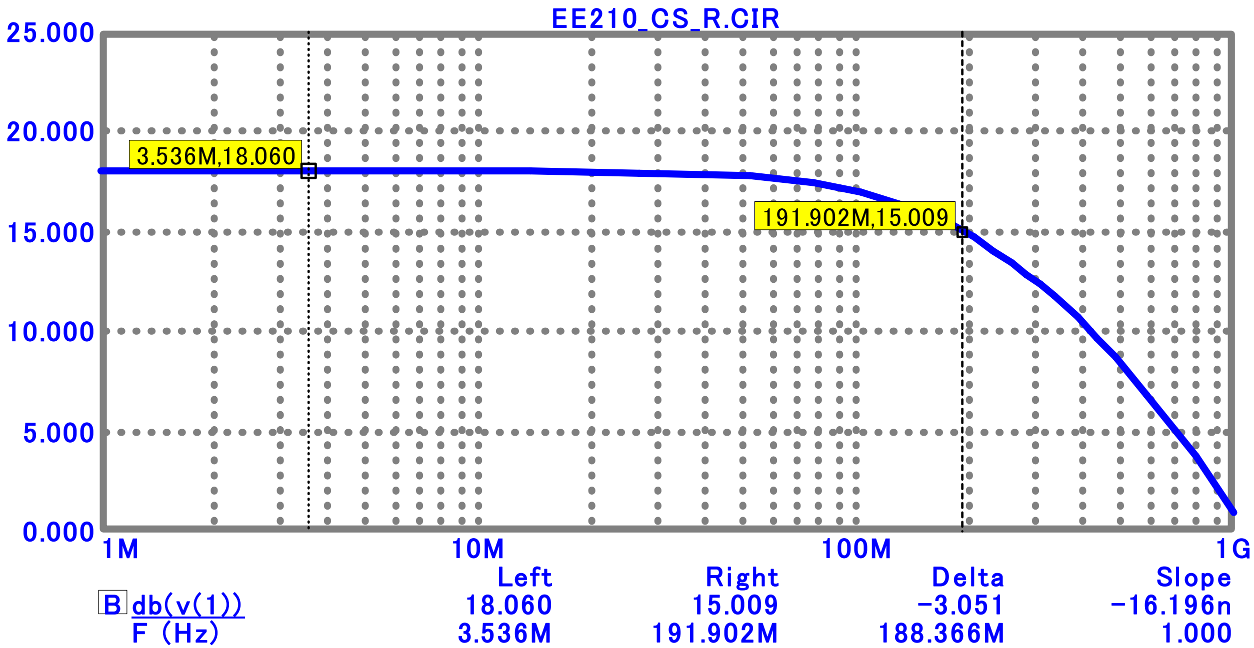
$$f_{3dB} = 1.49\times 10^9\,Hz$$

G-Number

CS amplifier



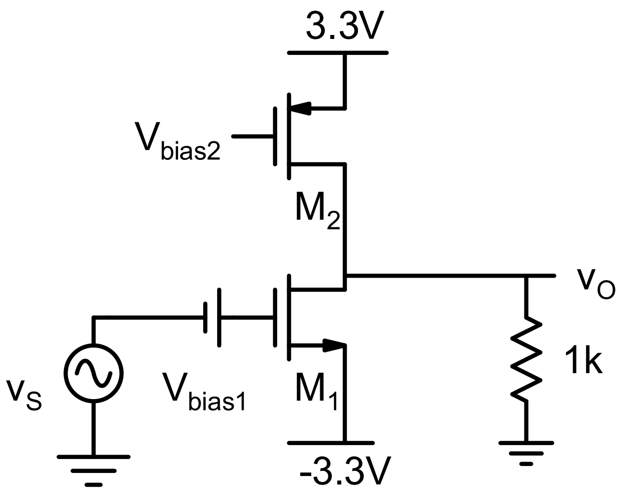
$$f_{3dB} = \frac{1}{2\pi} \times \frac{1}{R_G(C_{gs} + C_{gd}(1 + g_m R_D)) + R_D(C_{gd} + C_{db})}$$



$$f_{3dB} = 0.19GHz$$

CD amplifier has a superior frequency response

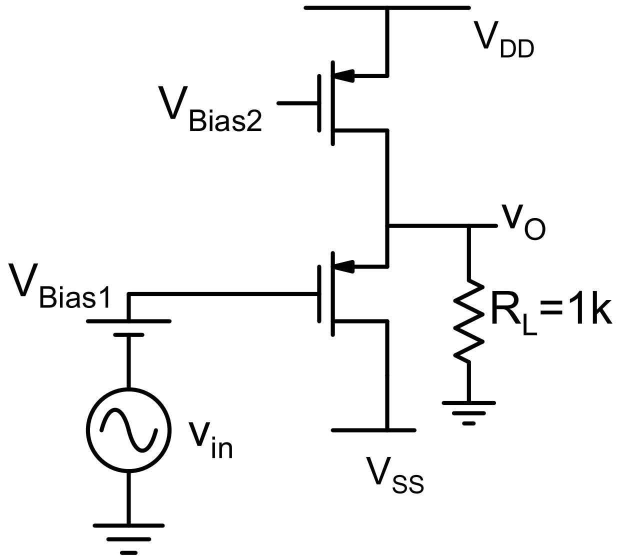
Summary



CS

1. Low Output resistance requires large bias current
2. Rail-to-rail output swing
3. Frequency response suffers from Miller's effect and is inferior.

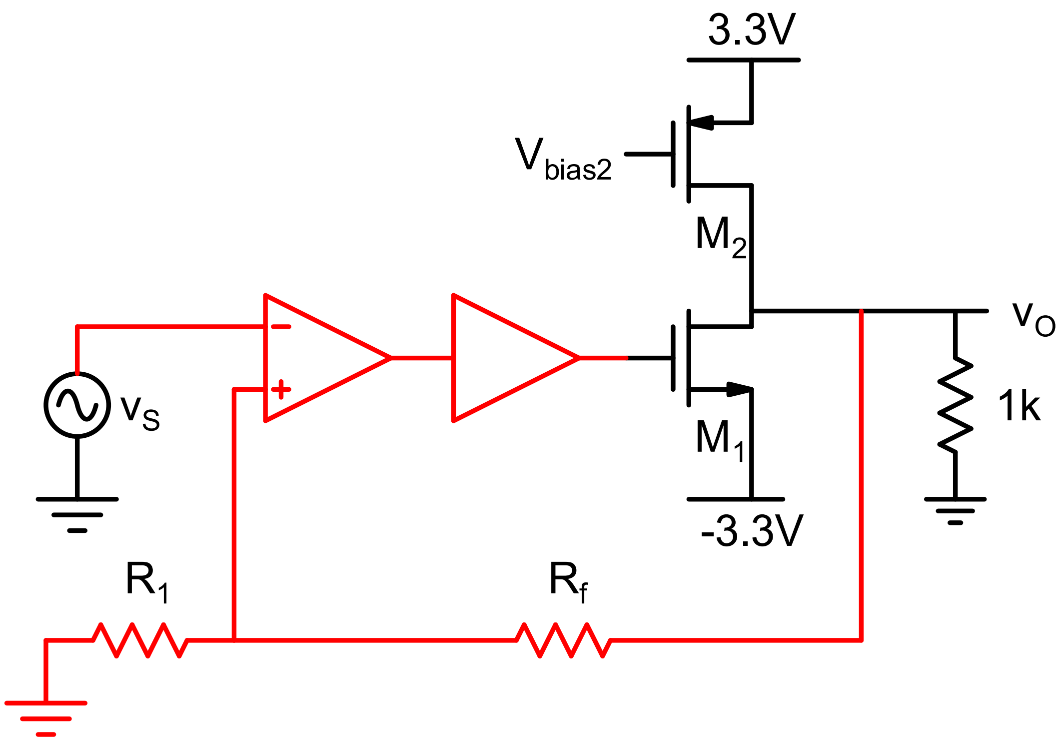
CD



1. Significantly Lower Output resistance can be obtained at same value of bias current
2. Swing lower by about a V_T drop
3. Good frequency response

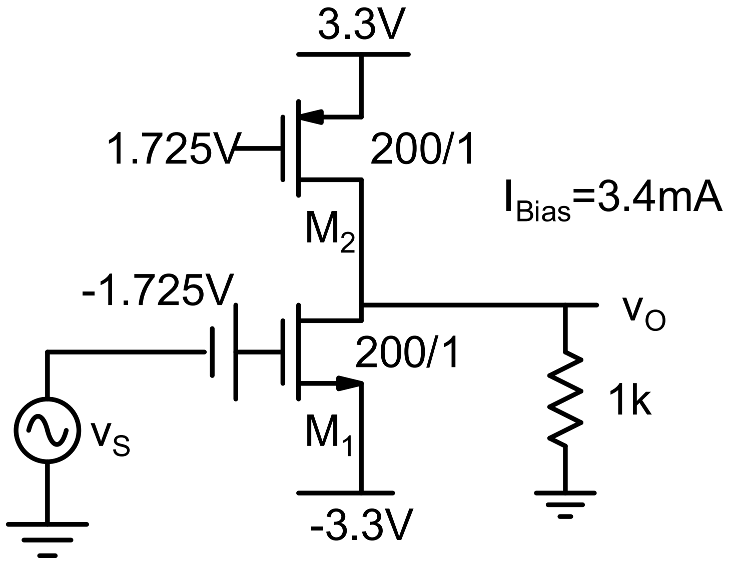
Efficiency limited to < 25% for both the stages

Multistage Amplifiers with Negative feedback

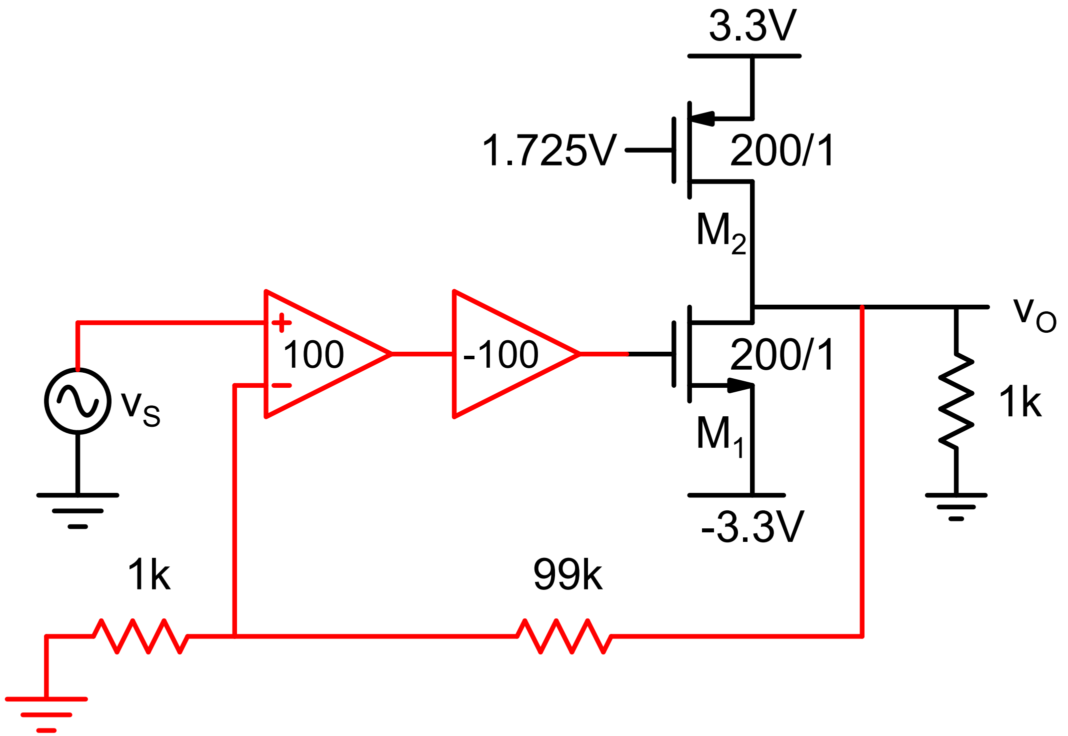


Negative feedback will help lower the output resistance

Example

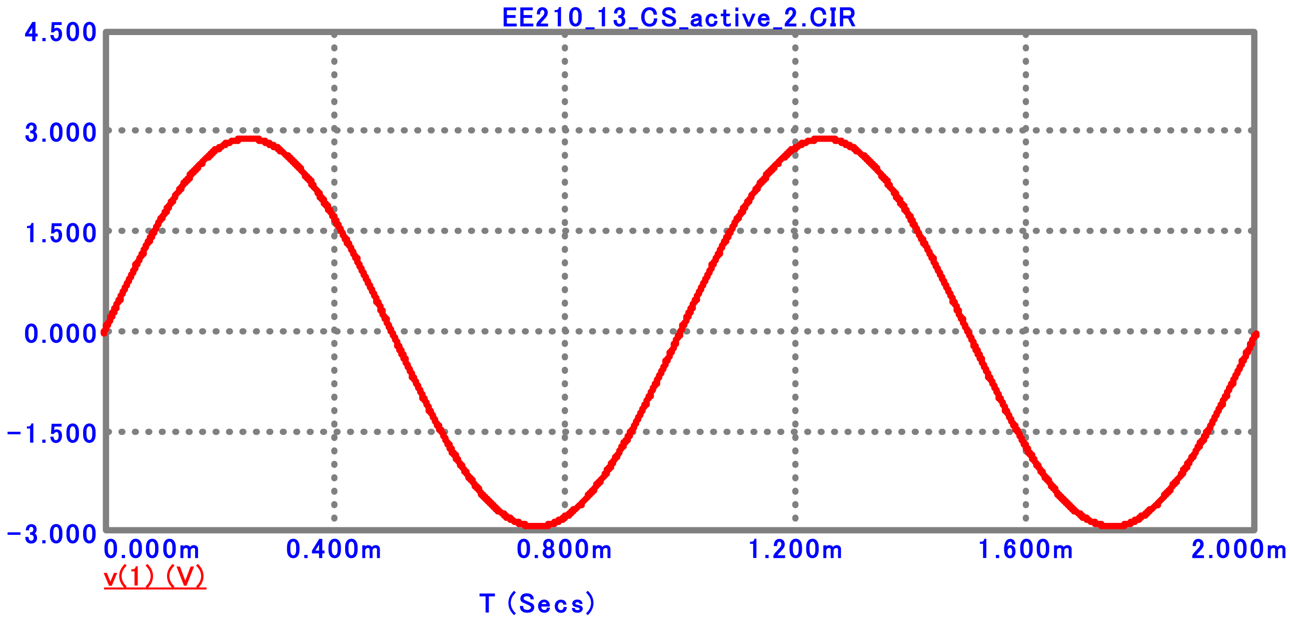


$$A_v = -11\ ;\ R_o = 938\Omega$$



$$A_v = -99.94\ ;\ R_o = 0.8\Omega$$

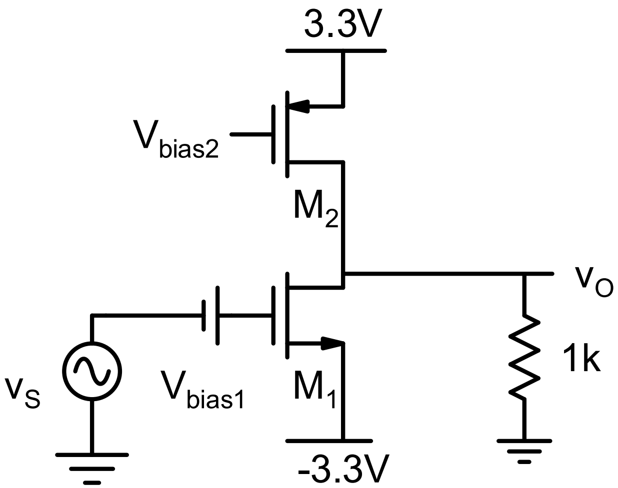
Output voltage for the feedback circuit

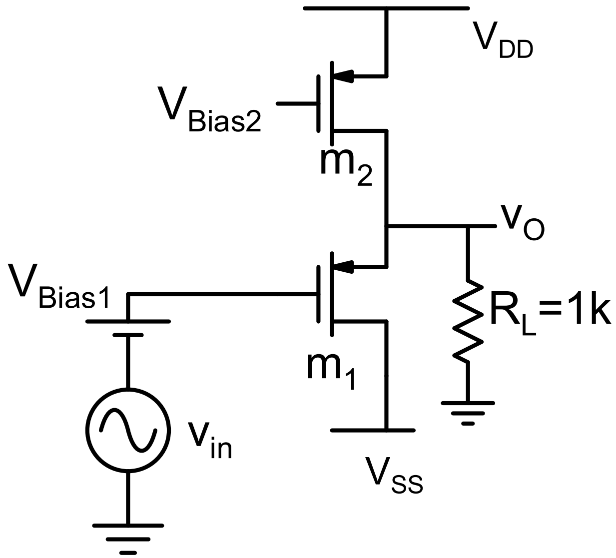


I_{bias} = 3.4mA

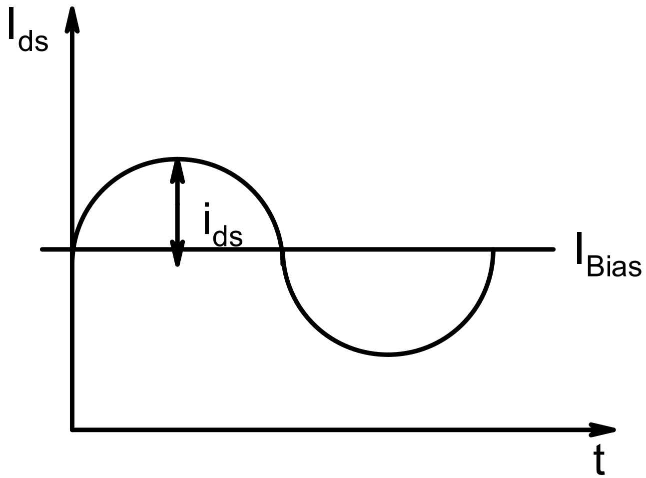
$$\eta = \frac{P_L}{P_{supply}} = \frac{4.5mW}{22mW} = 0.205$$

Class A amplifiers





In both CS and CD amplifiers, the transistor remains ON and conducting throughout out the ac cycle. To achieve this the bias current must be larger than the ac current.



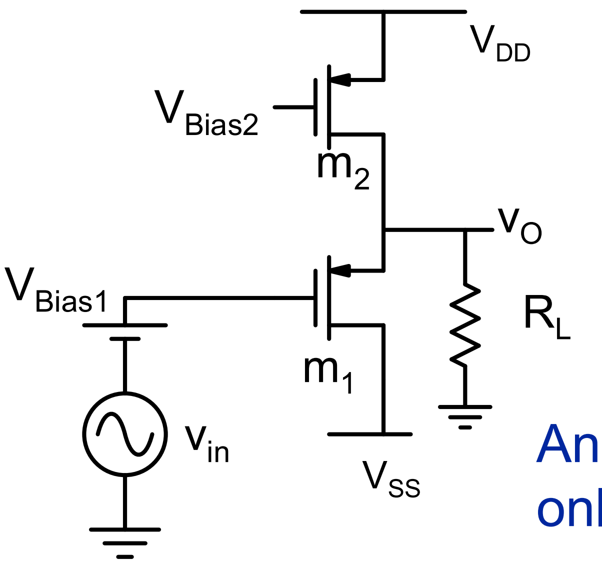
$$I_{ds} = I_{Bias} + i_{ds} \qquad i_{ds} \le I_{Bias} \qquad i_L \le i_{ds} \le I_{Bias}$$

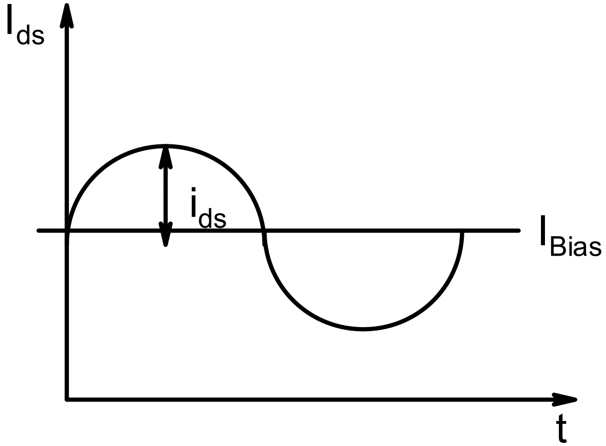
$$v_o \le V_{DD} \qquad P_L = 0.5 v_o \times i_L < 0.5 V_{DD} I_{Bias}$$

$$P_{ss} = -\frac{1}{T}\int_0^T V_{SS} \times I_{ds} dt = -V_{ss} I_{Bias} \qquad P_{dd} = V_{DD} \times I_{Bias}$$

$$\eta = \frac{P_L}{P_{\sup ply}} \le 0.25$$

Amplifiers with negligible stand-by power dissipation

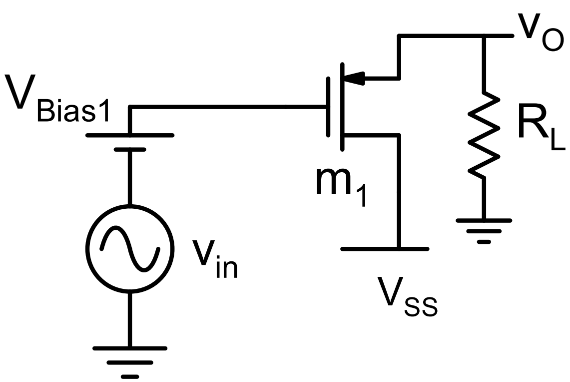


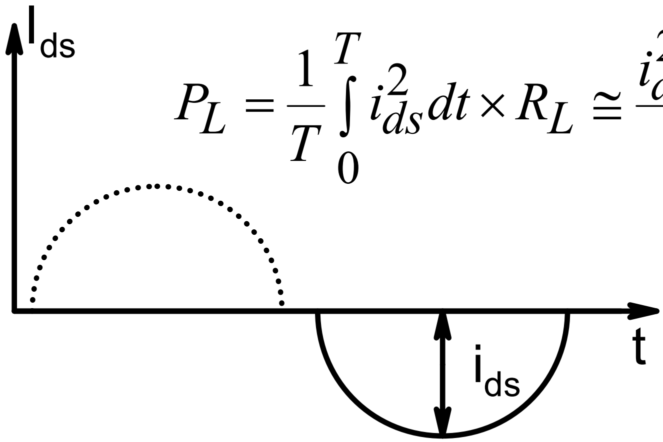


Even when no input is applied, power is drawn from the supply.

An efficient amplifier will take power from the supply only when power is to be delivered to the load.

Bias current must be zero !!



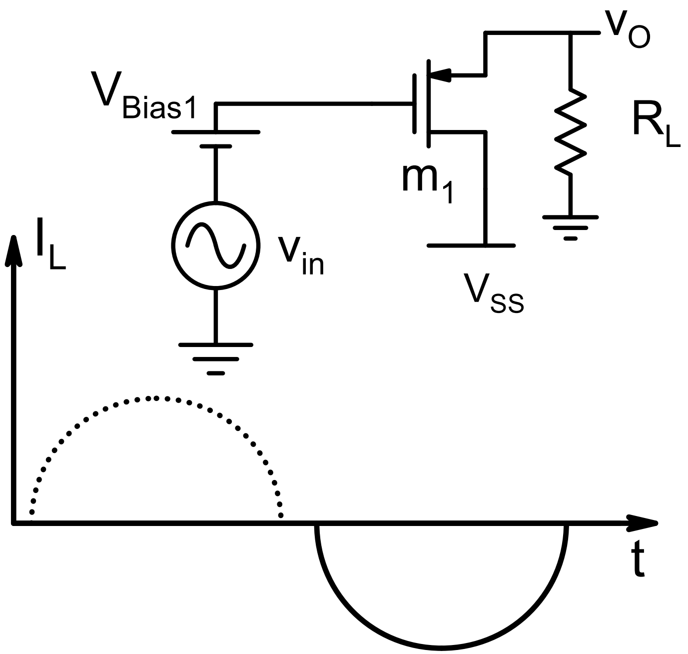


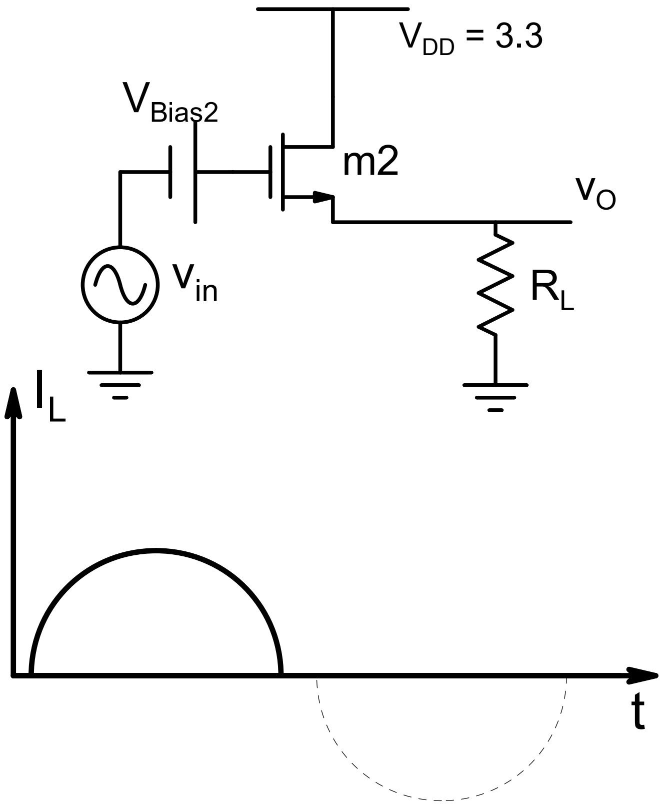
$$P_L = \frac{1}{T}\int_0^T i_{ds}^2 dt \times R_L \cong \frac{i_{ds\max}^2 R_L}{4}$$

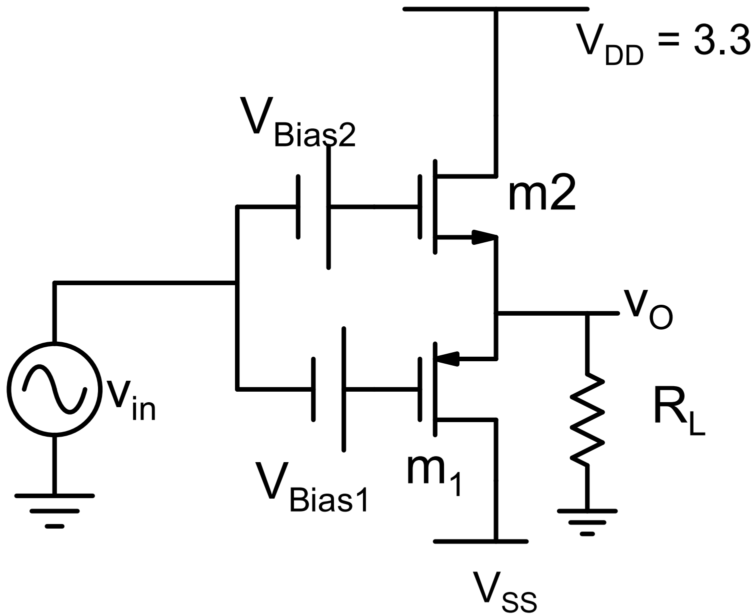
$$P_{ss} = -\frac{1}{T}\int_0^T V_{SS} \times I_{ds} dt \cong \frac{|V_{ss}| \times i_{ds\max}}{\pi}$$

$$\boxed{\eta = \frac{\pi}{4} \times \frac{i_{ds\max} R_L}{|V_{SS}|} = 0.785 \times \frac{v_{o\max}}{|V_{SS}|}}$$

How do we reduce distortion?

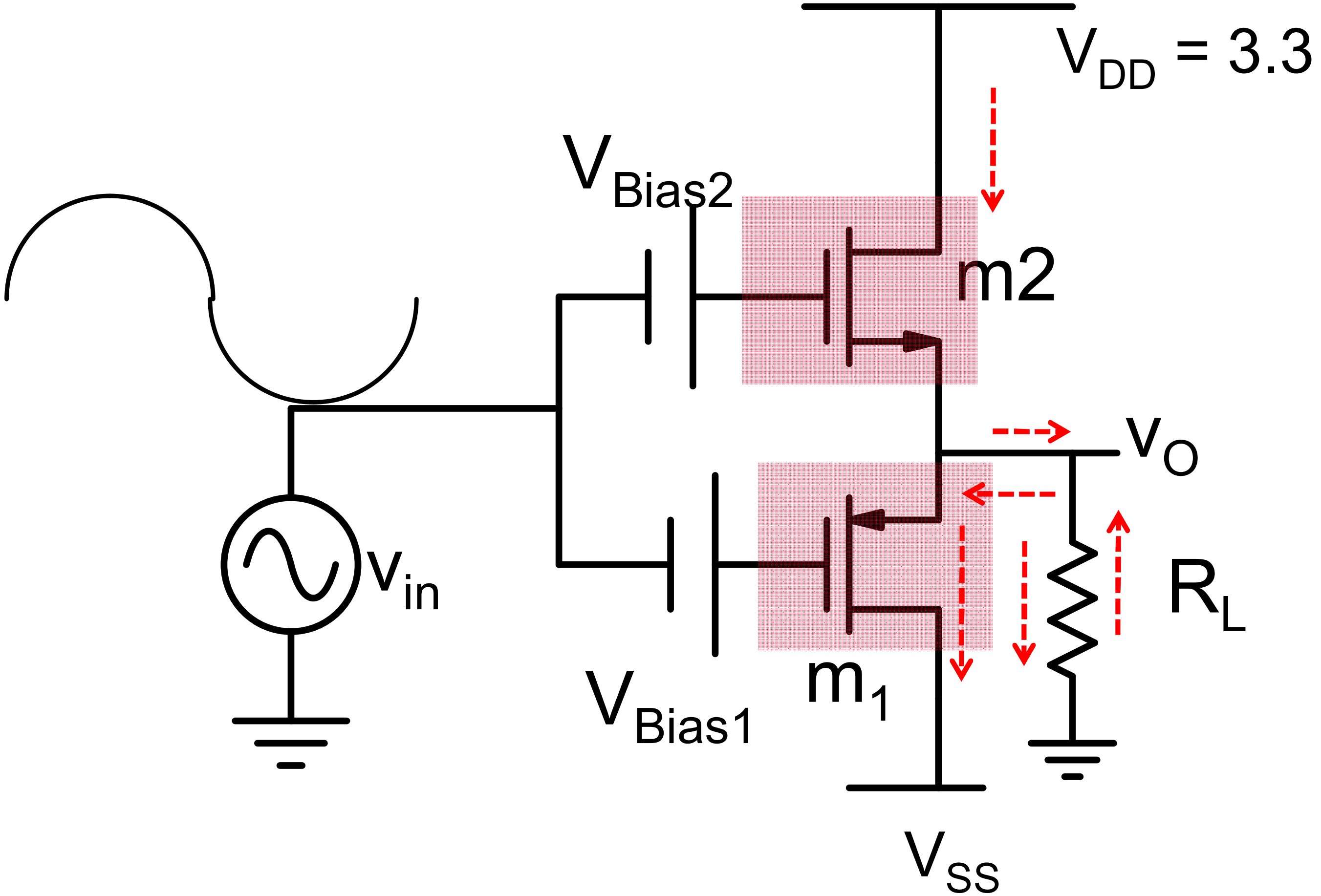






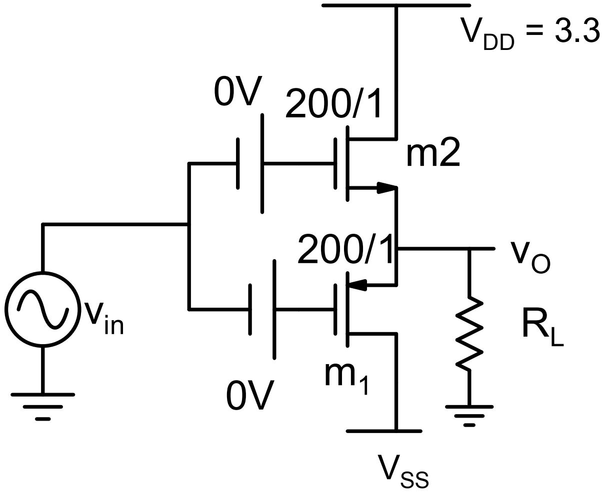
Each Transistor conducts for only half the cycle resulting in Class B operation

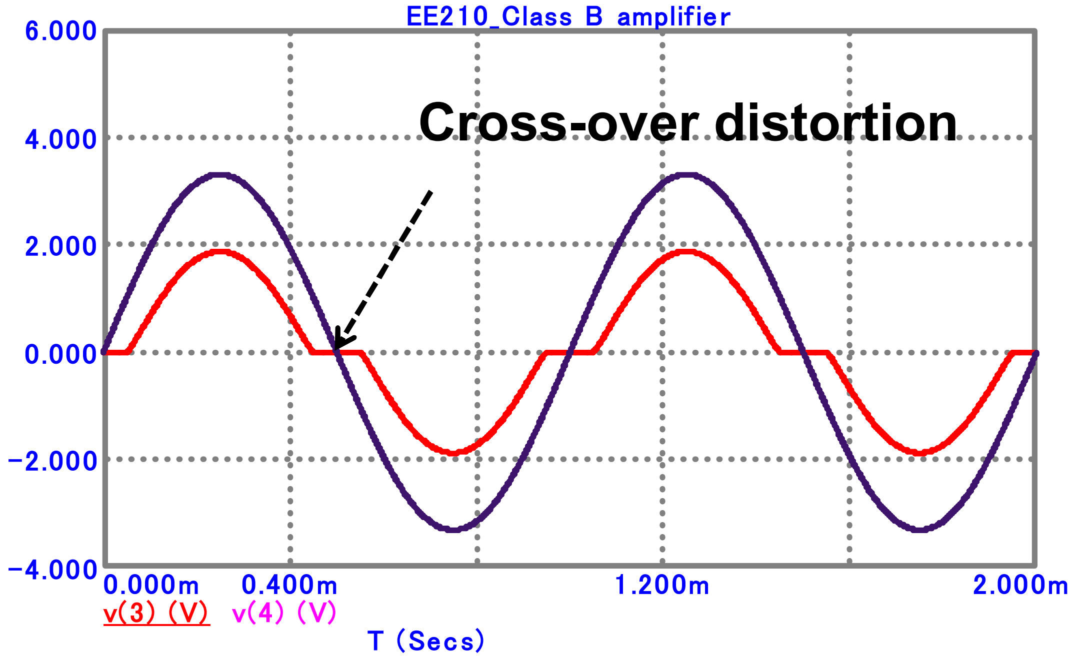
Class B push-pull amplifier



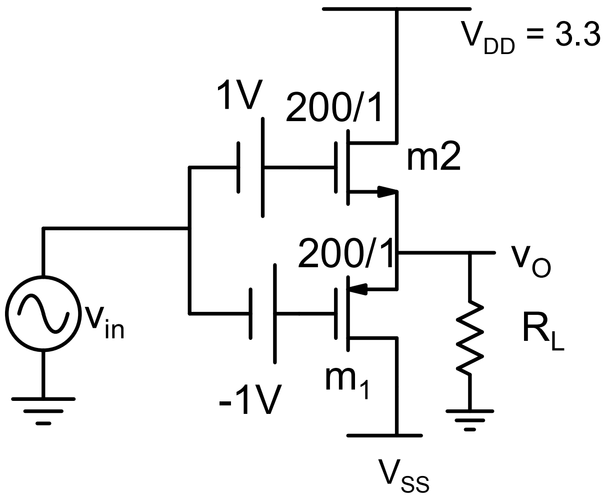
During positive cycle, M2 pushes current into the load, while during the negative cycle, M1 pull current from the load and hence the name Push-Pull amplifier

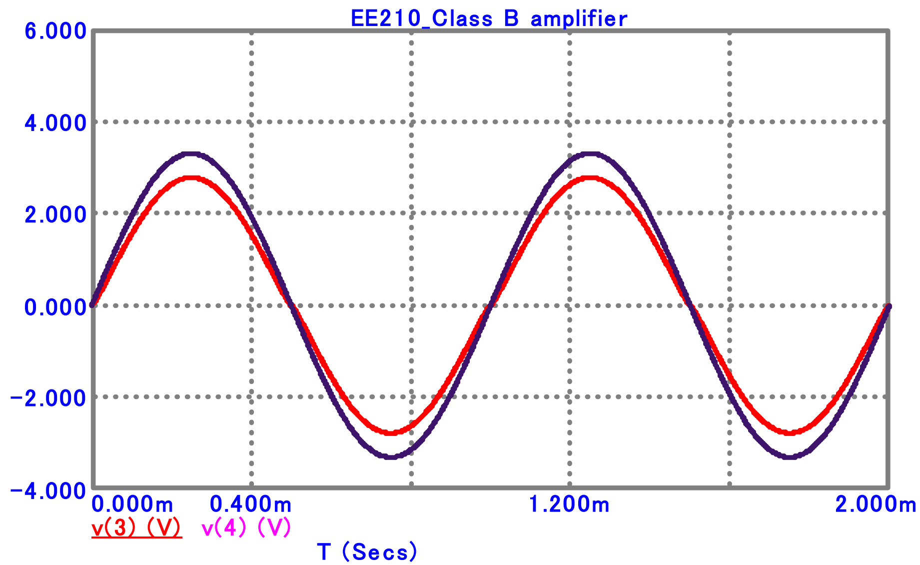
Symmetrical nmos and pmos with identical parameters and no body effect for nmos.





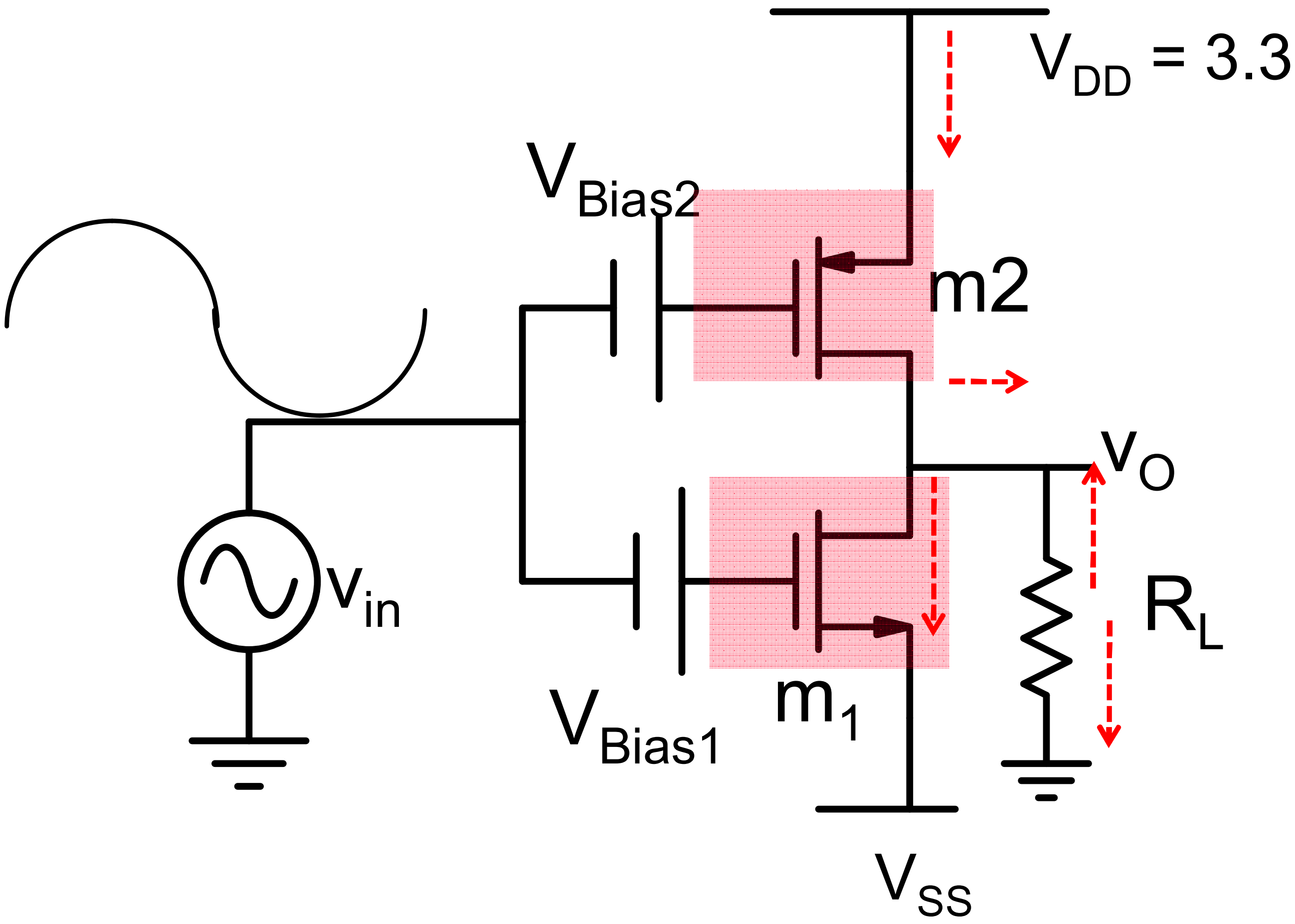
Class AB amplifier



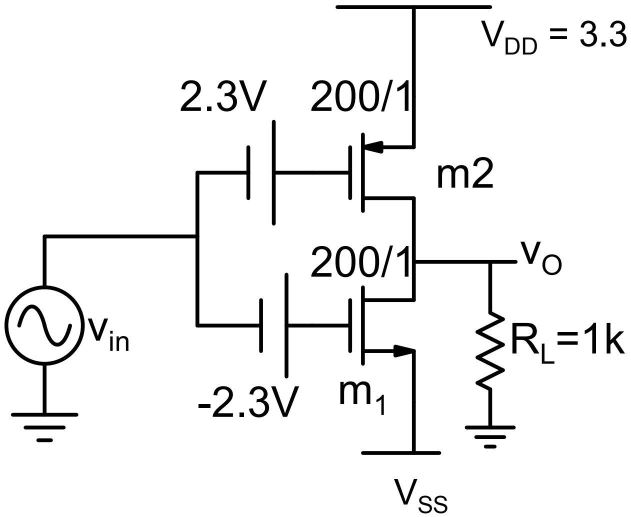


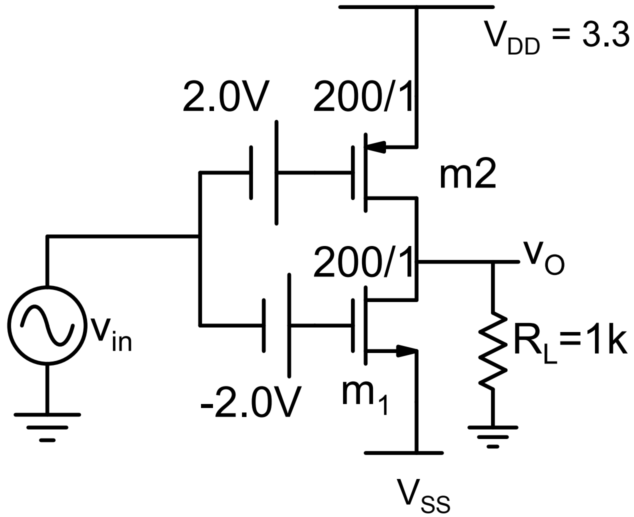
G-Number

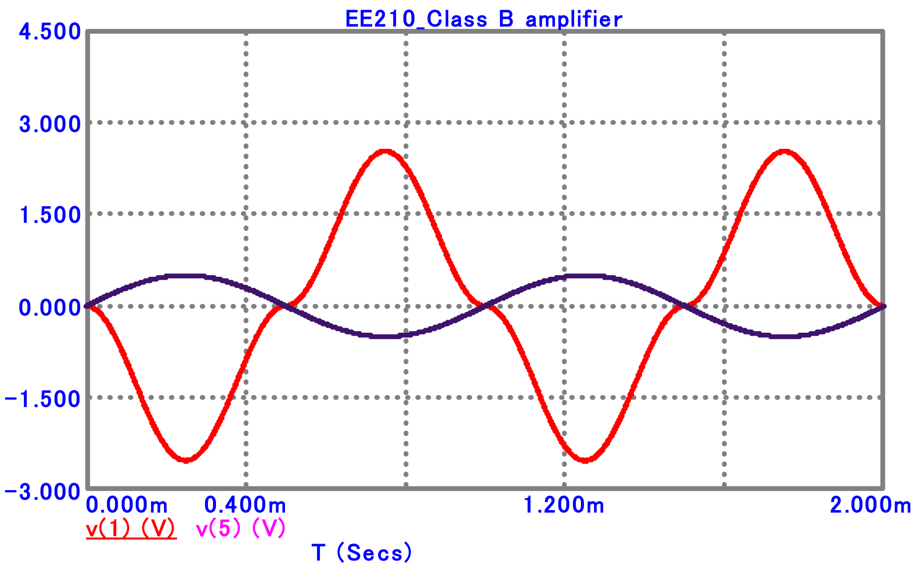
Class AB Push-Pull amplifier using CS stage

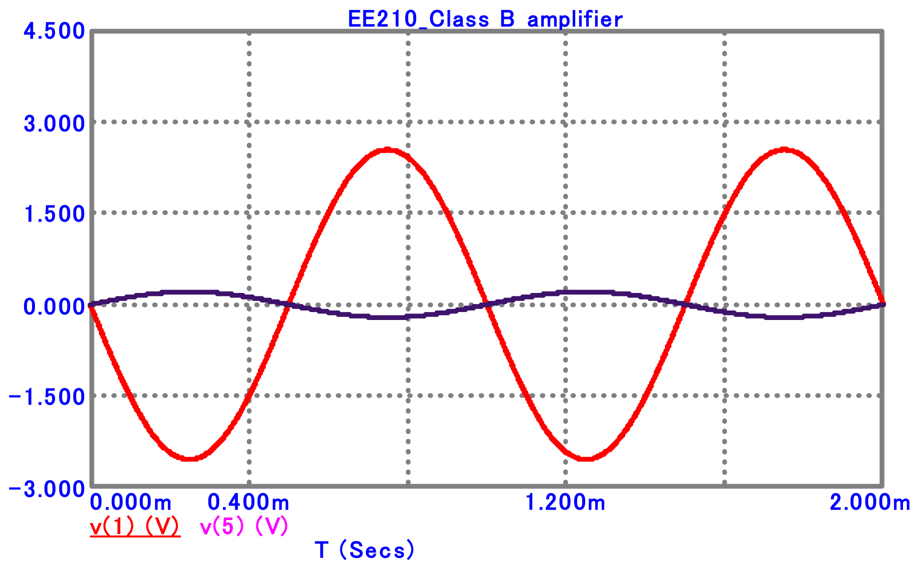


Symmetrical nmos and pmos with identical parameters and no body effect for nmos.

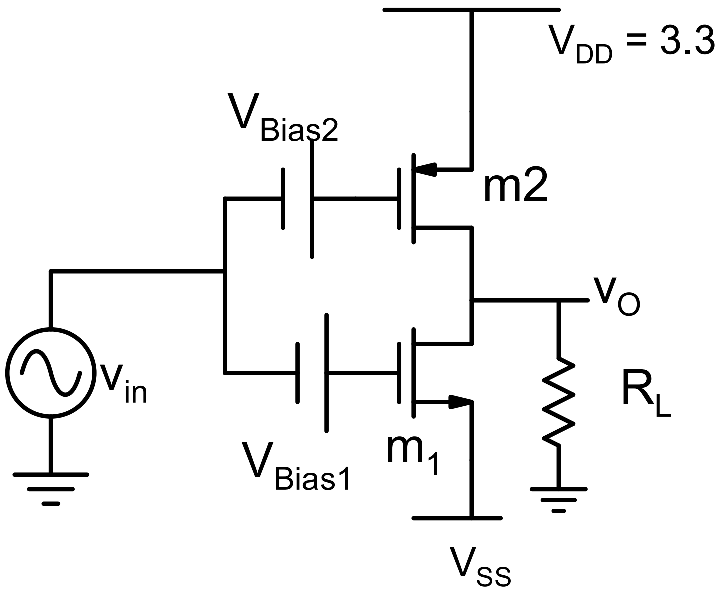


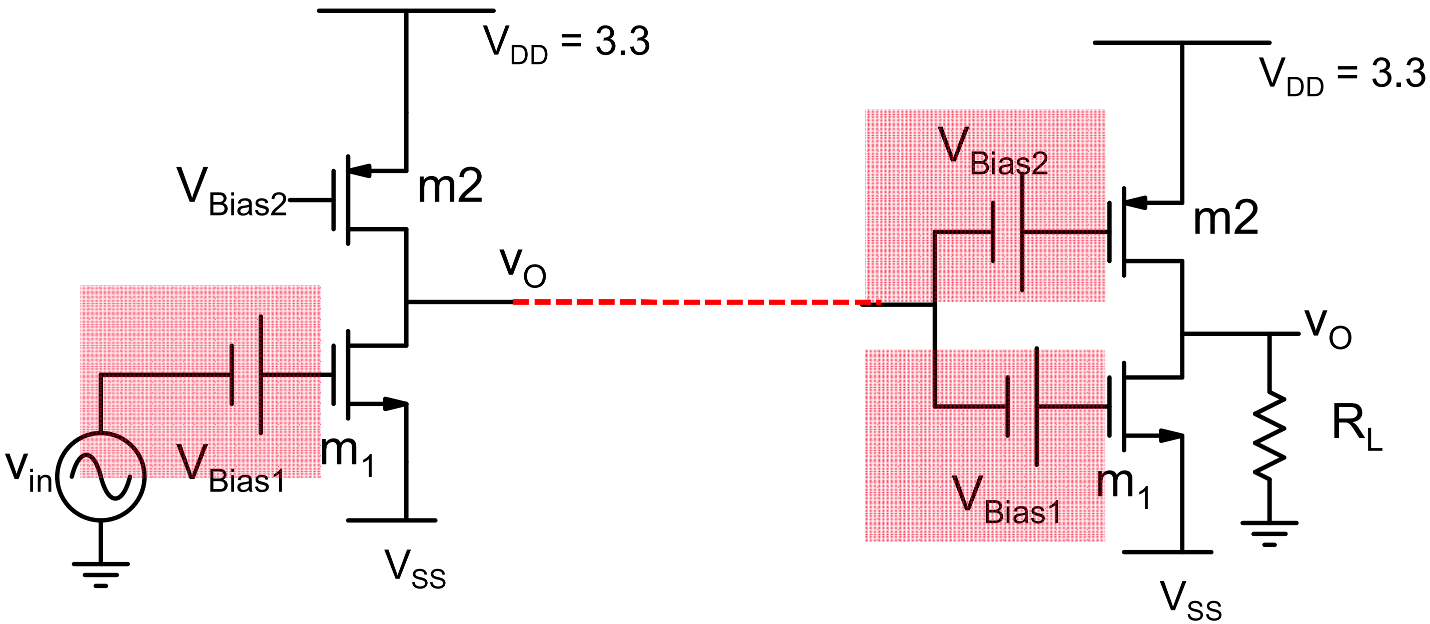






G-Number





Need to generate bias voltages

EEE210: Microelectronics-I

Lecture-45 : MOS Differential Amplifiers

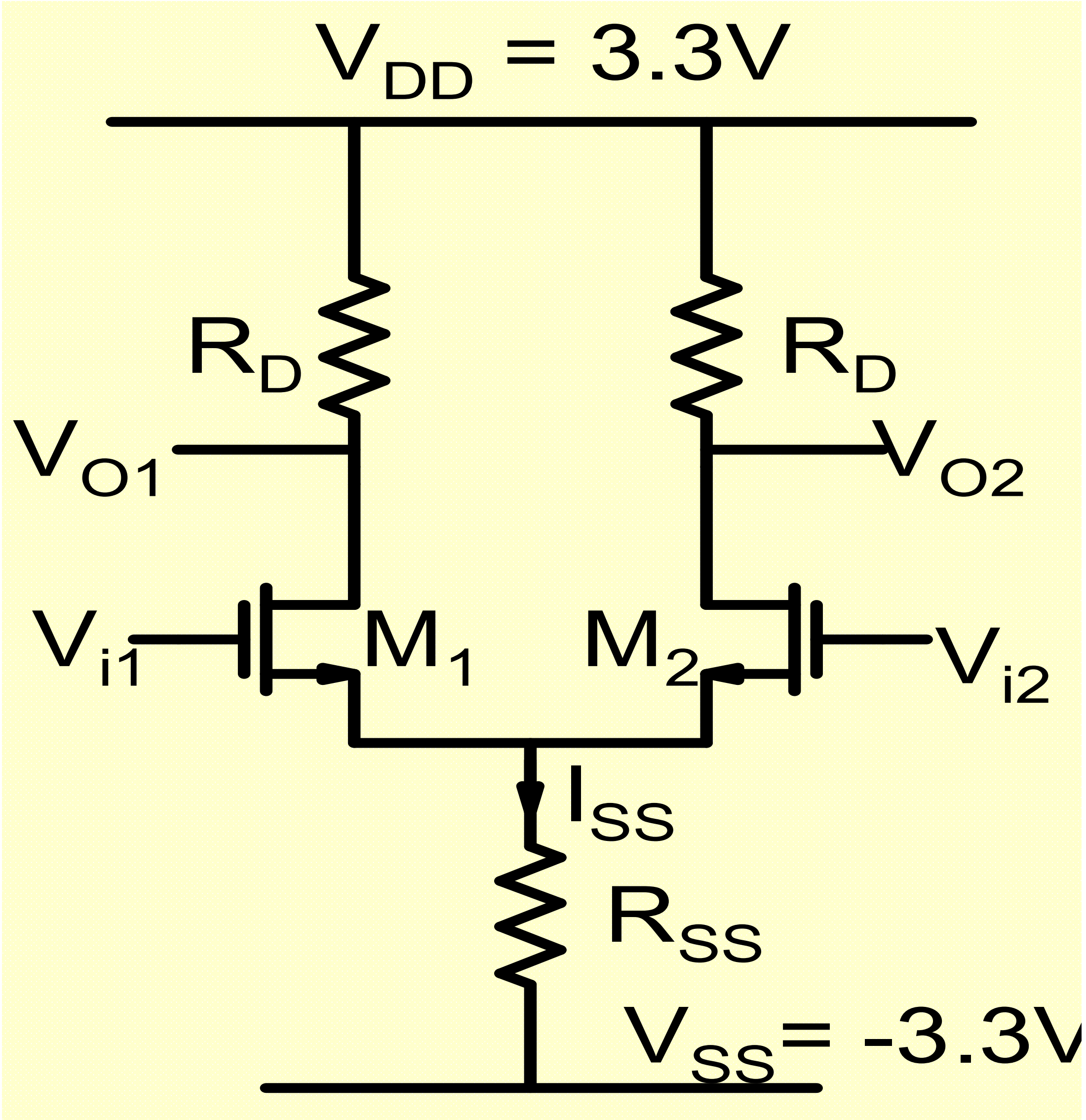
http://youtu.be/SVLGvOAyjhs

B. Mazhari
Dept. of EE, IIT Kanpur

G-Number

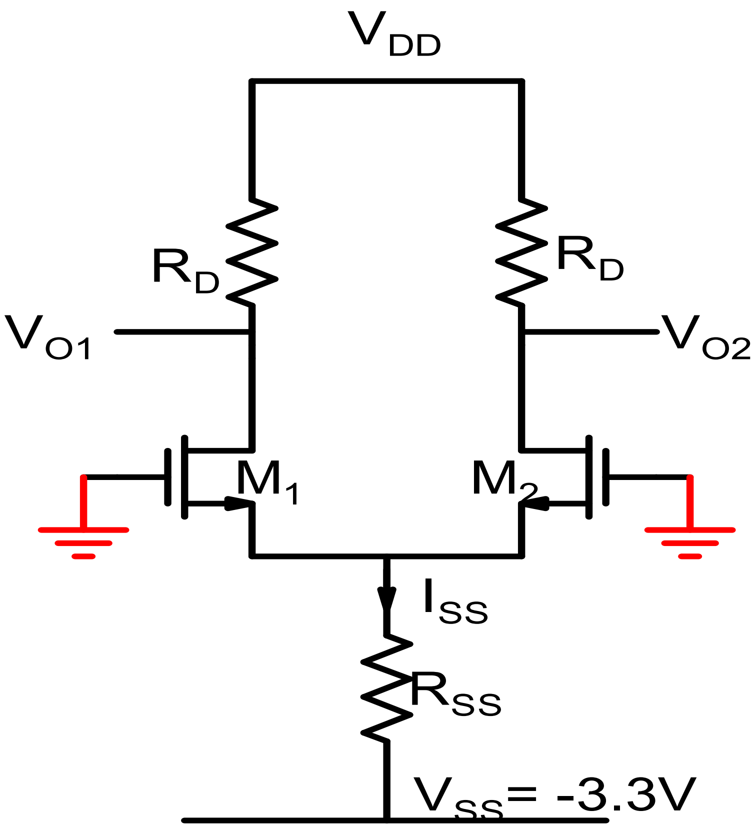
MOS Differential Amplifier

MOS Differential Pair with Resistive Load



Bias Point analysis

M1 and M2 are assumed to be identical



$$I_{DSQ1} = I_{DSQ2} = 0.5 I_{SS}$$

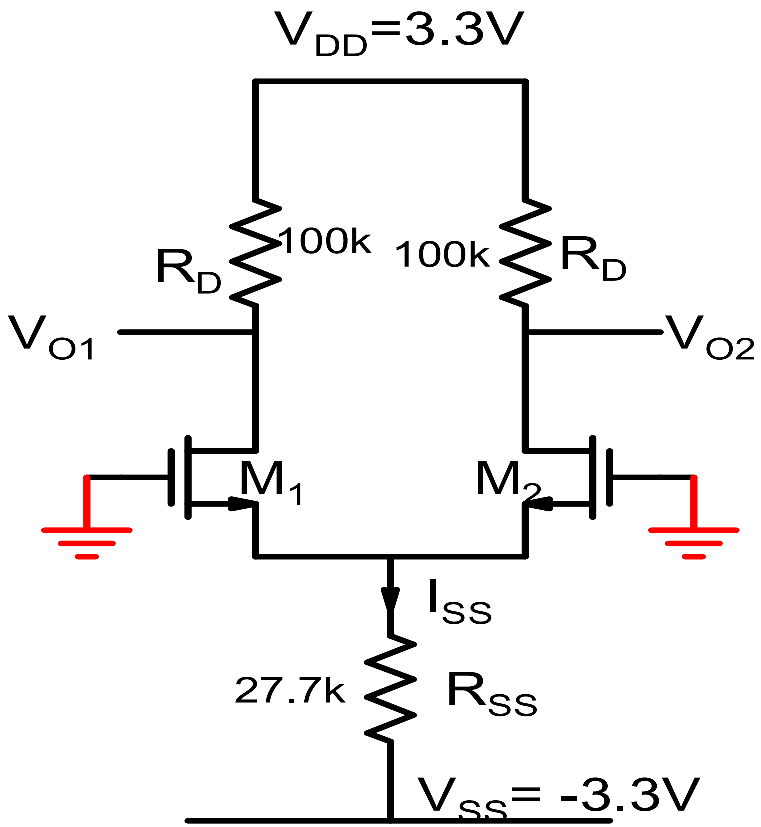
$$I_{SS} = \frac{-V_{GSQ} - V_{SS}}{R_{SS}}$$

$$V_{GSQ} = V_{TN} + \sqrt{\frac{2 I_{DSQ}}{\beta}}$$

$$V_{TN} = V_{TNO} + \gamma\left(\sqrt{2\phi_F + V_{SBQ}} - \sqrt{2\phi_F}\right)$$

$$V_{SBQ} = -V_{GSQ} - V_{SS}$$

Example



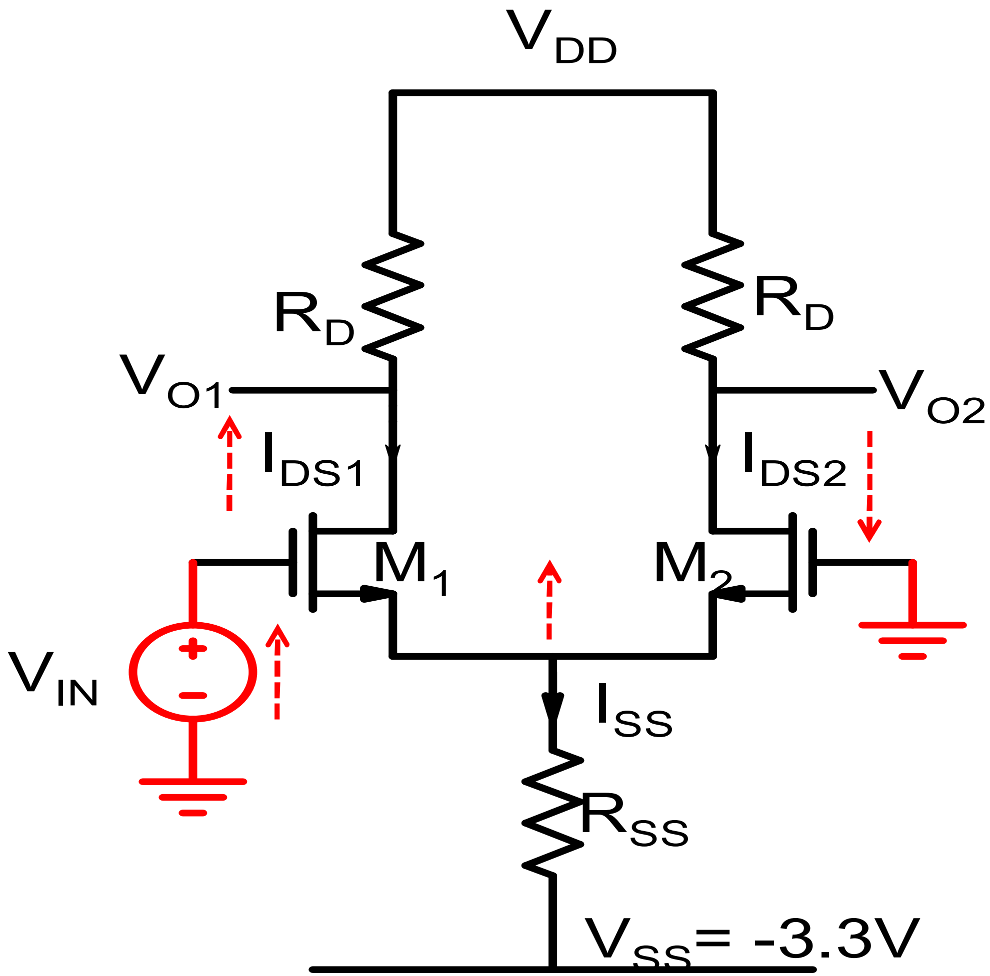
$$I_{SS} = \frac{-V_{GSQ} - V_{SS}}{R_{SS}} = 50\mu A$$

$$I_{DSQ1} = I_{DSQ2} = 0.5 I_{SS} = 25\mu A$$

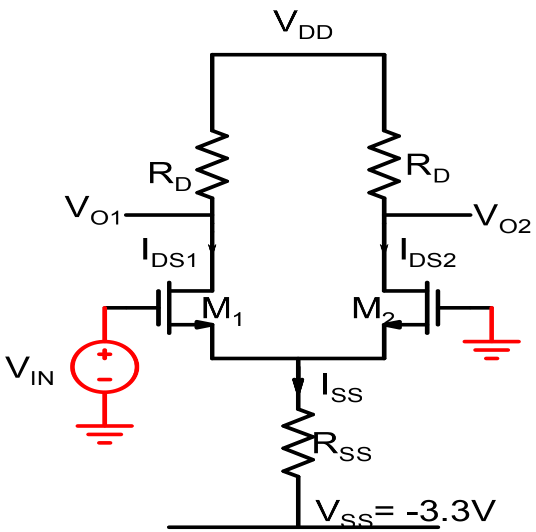
$$V_{TN} = 1.42V \qquad V_{GSQ} = 1.92V$$

$$V_{01}(dc) = 0.8V$$

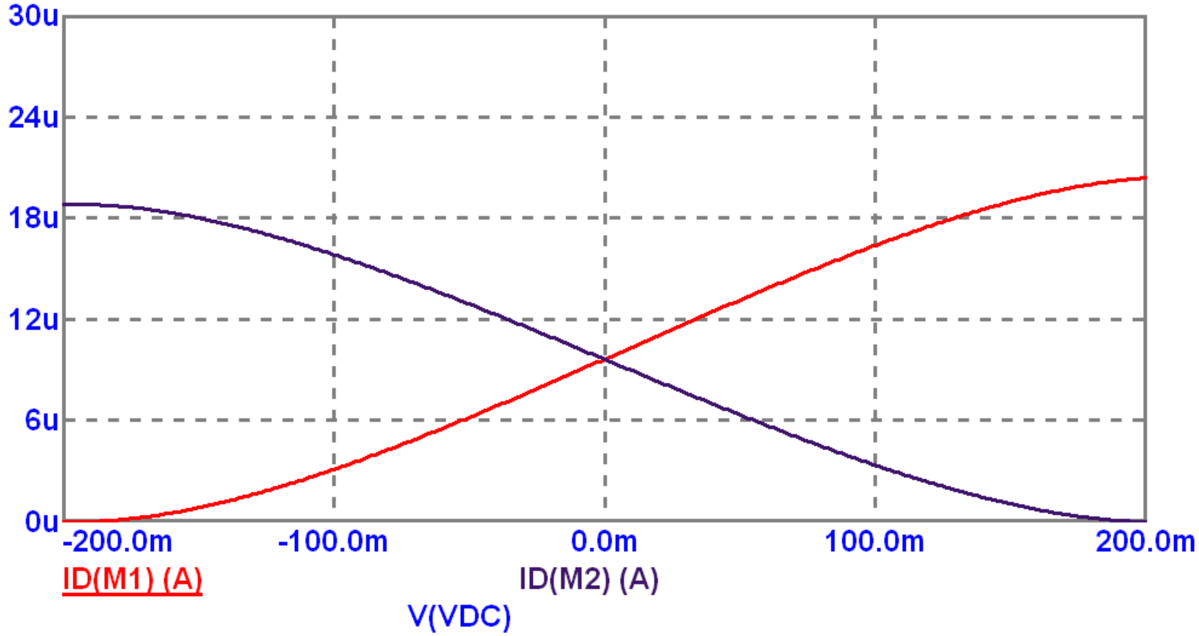
Current Switching in differential Amplifier



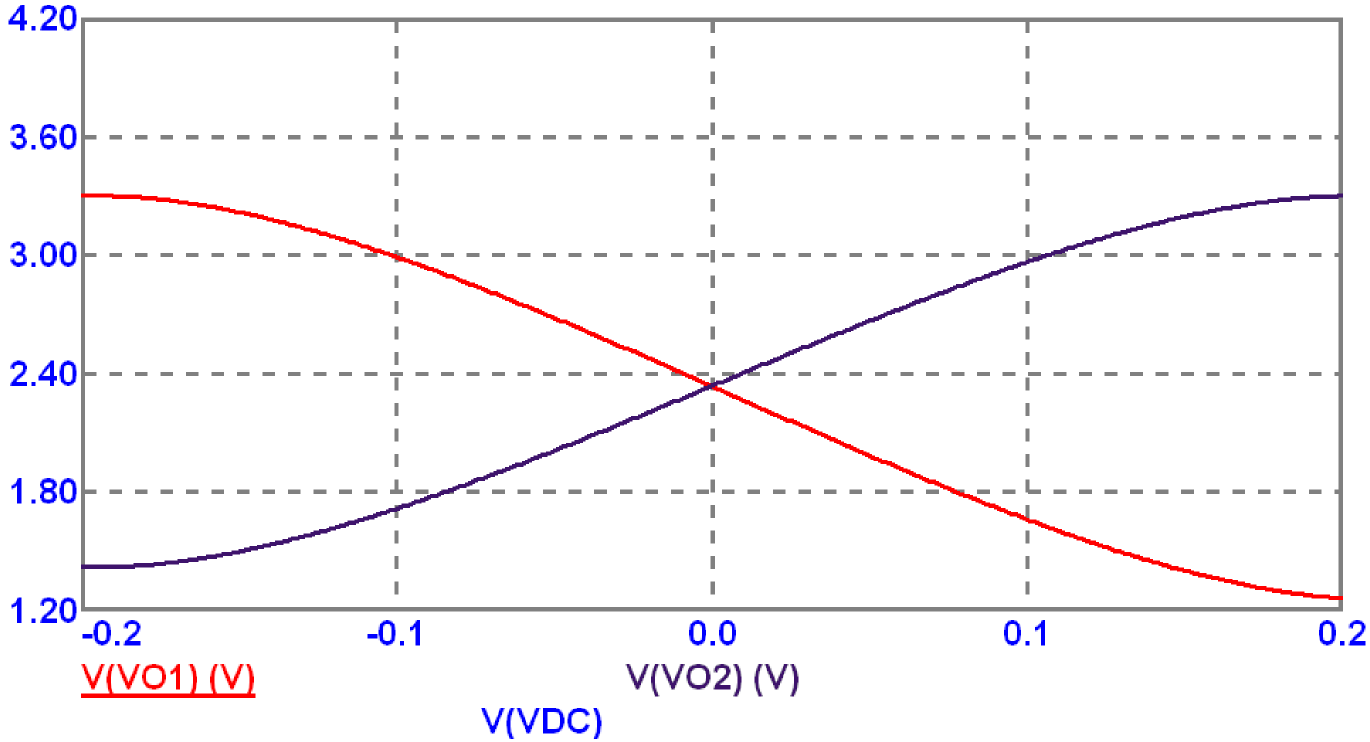
G-Number



VDD
RD
RD
VO1
VO2
IDS1
IDS2
M1
M2
VIN
ISS
RSS
VSS= -3.3V

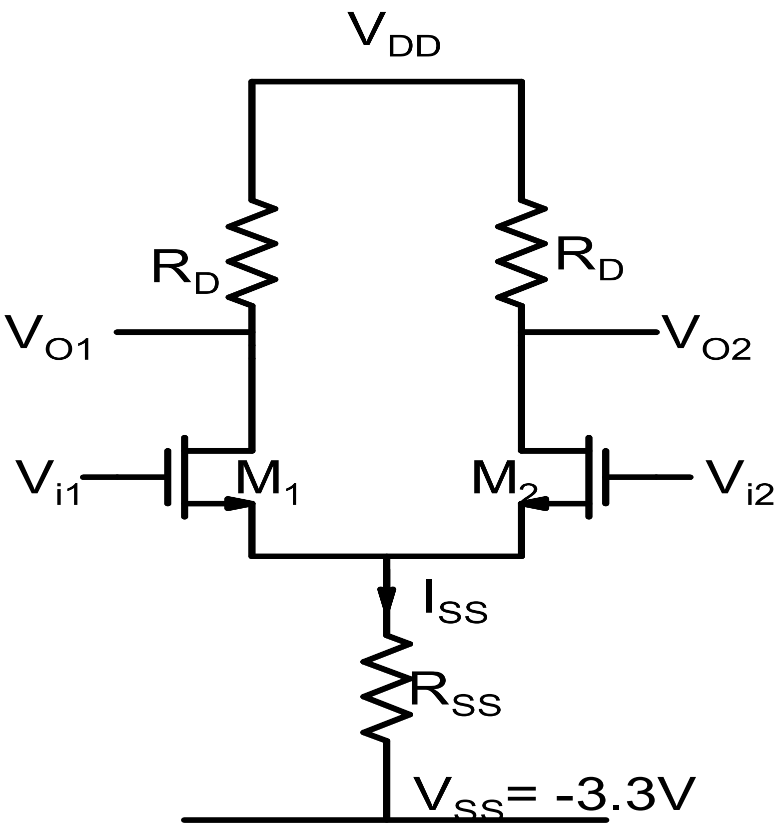


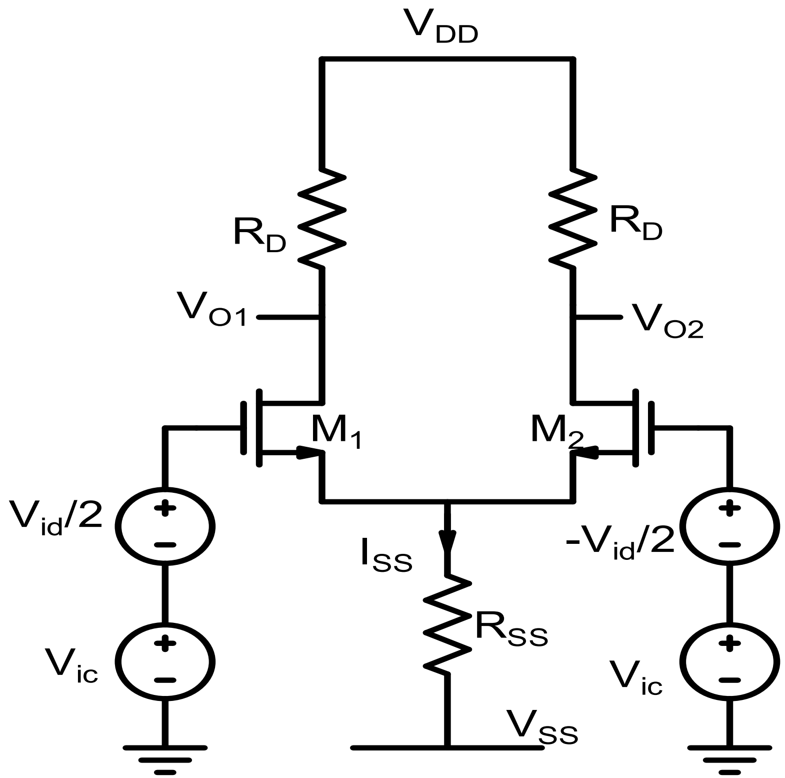
30u
24u
18u
12u
6u
0u
-200.0m
-100.0m
0.0m
100.0m
200.0m
ID(M1) (A)
ID(M2) (A)
V(VDC)



4.20
3.60
3.00
2.40
1.80
1.20
-0.2
-0.1
0.0
0.1
0.2
V(VO1) (V)
V(VO2) (V)
V(VDC)

Differential Amplifier: Small Signal Analysis

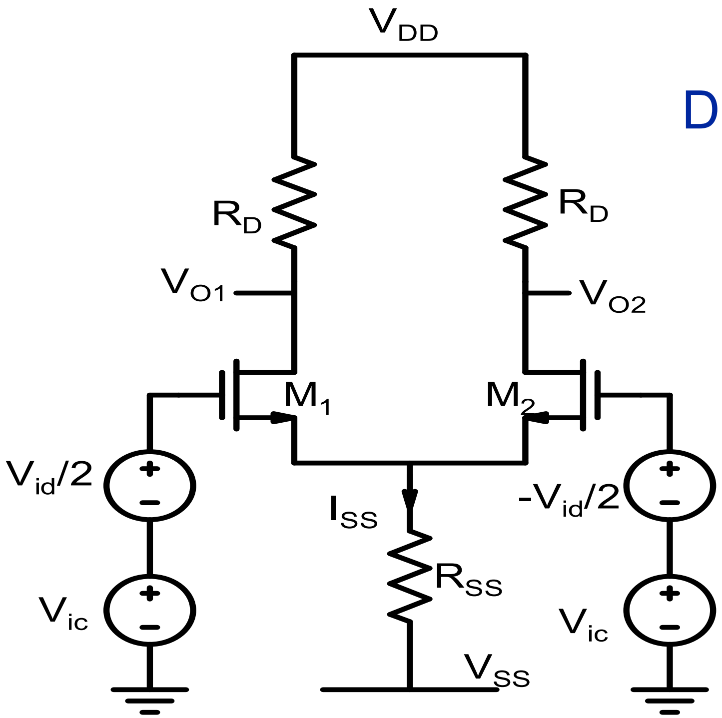




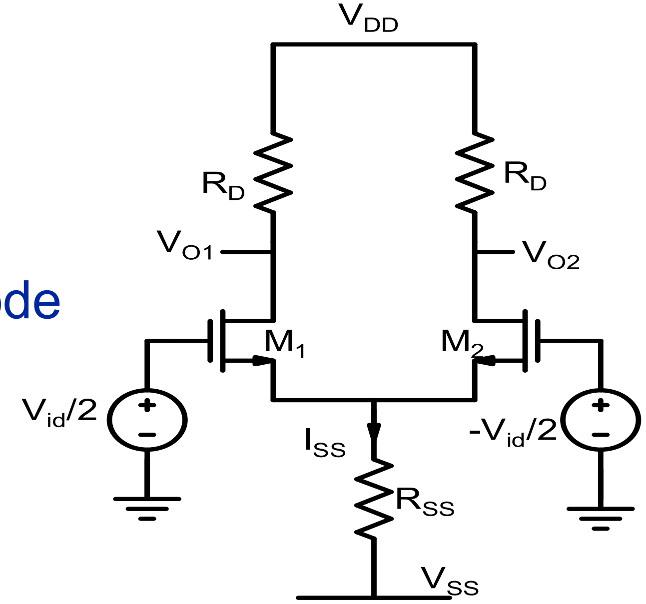
$$v_{id} = v_{i1} - v_{i2} \qquad v_{ic} = \frac{v_{i1} + v_{i2}}{2}$$

$$v_{i1} = \frac{v_{id}}{2} + v_{ic} \qquad v_{i2} = -\frac{v_{id}}{2} + v_{ic}$$

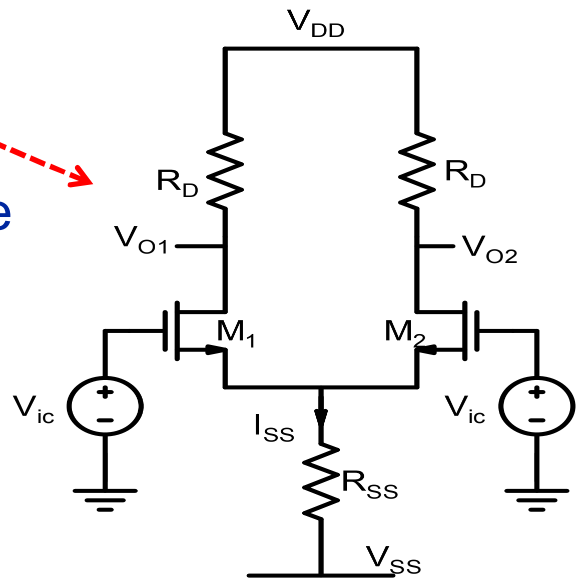
Superposition can be used to simplify analysis



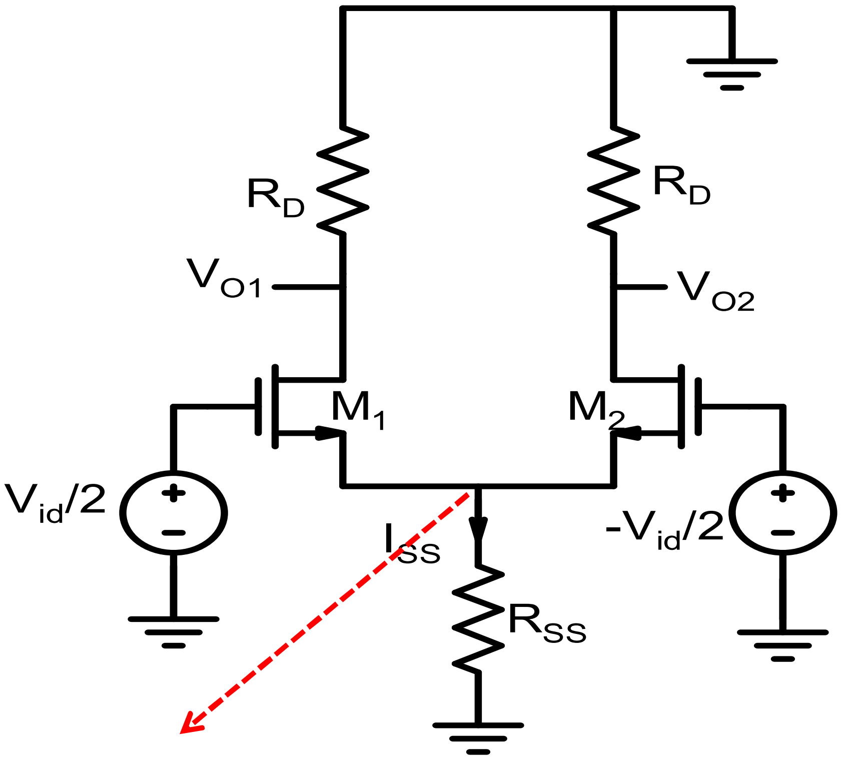
Differential mode



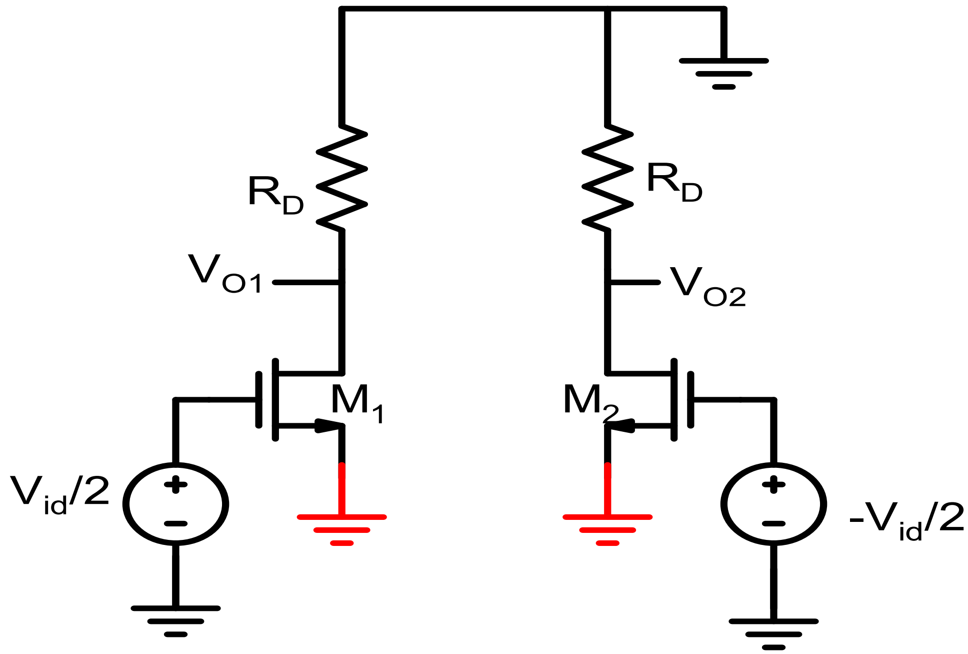
Common mode



Differential Mode Analysis



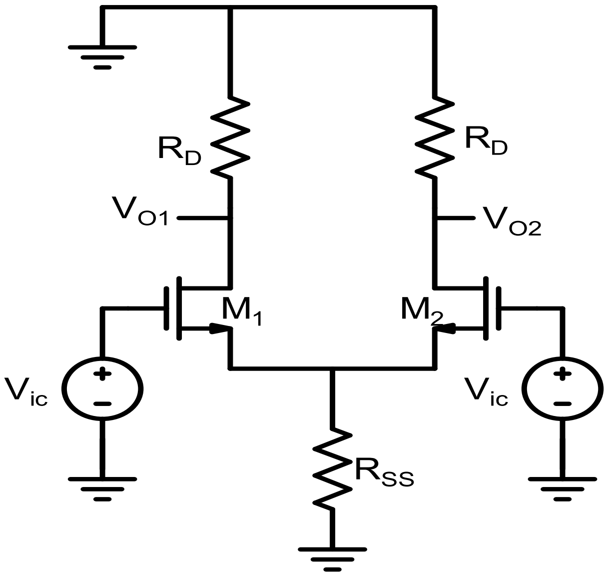
Small-signal ground

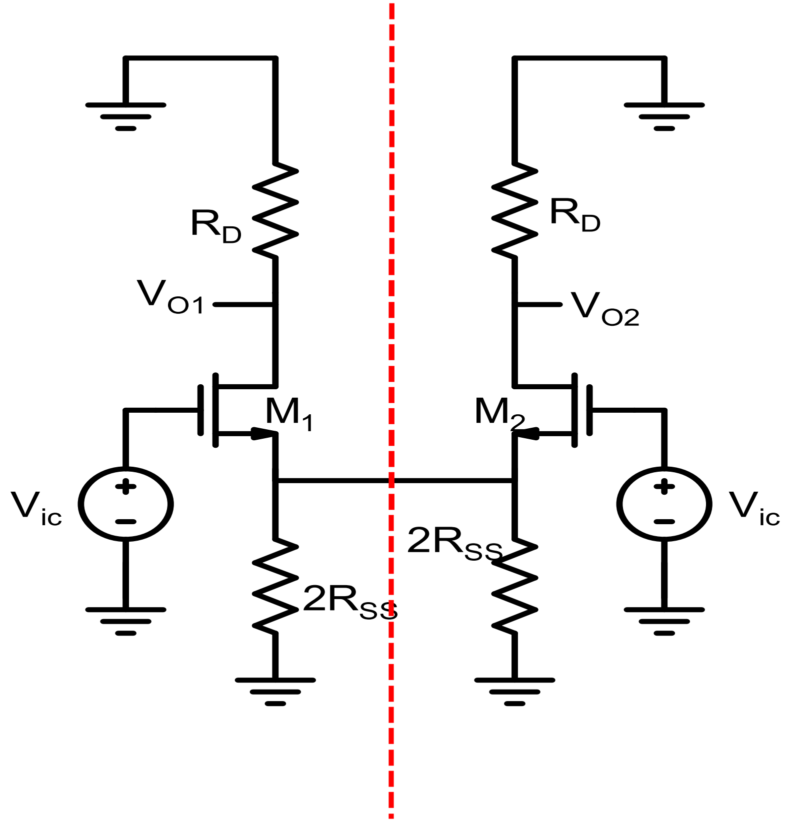


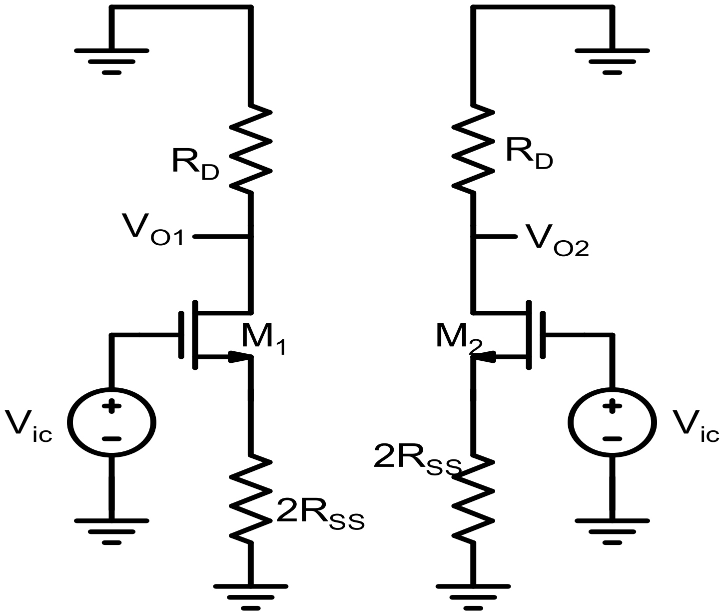
$$\frac{v_{o1}}{\frac{v_{id}}{2}} = -g_m R_D \qquad \frac{v_{o2}}{-\frac{v_{id}}{2}} = -g_m R_D$$

$$\frac{v_{o1}}{v_{id}} = -0.5 g_m R_D \qquad \frac{v_{o2}}{v_{id}} = 0.5 g_m R_D$$

Common Mode Analysis

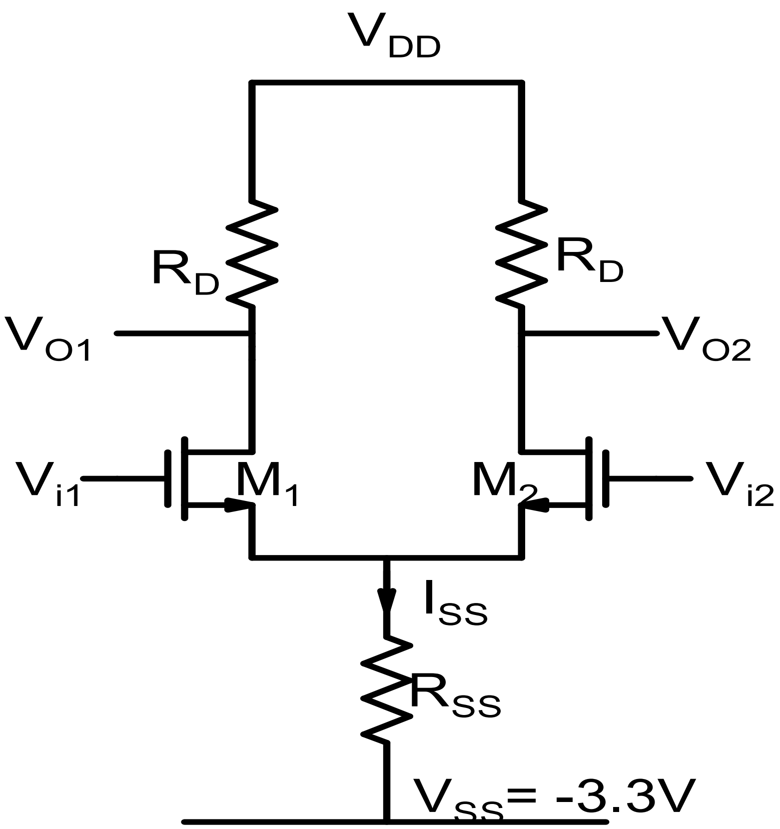






$$\frac{v_{o1}}{v_{ic}} = \frac{v_{o2}}{v_{ic}} = -\frac{g_m R_D}{1 + 2 g_m R_{SS}}$$

Common Mode Rejection Ratio



$$A_{dm} = -0.5 g_m R_D$$

$$A_{cm} = -\frac{g_m R_D}{1 + 2 g_m R_{SS}}$$

$$cmrr = \frac{A_{dm}}{A_{cm}} = 0.5 + g_m R_{SS}$$

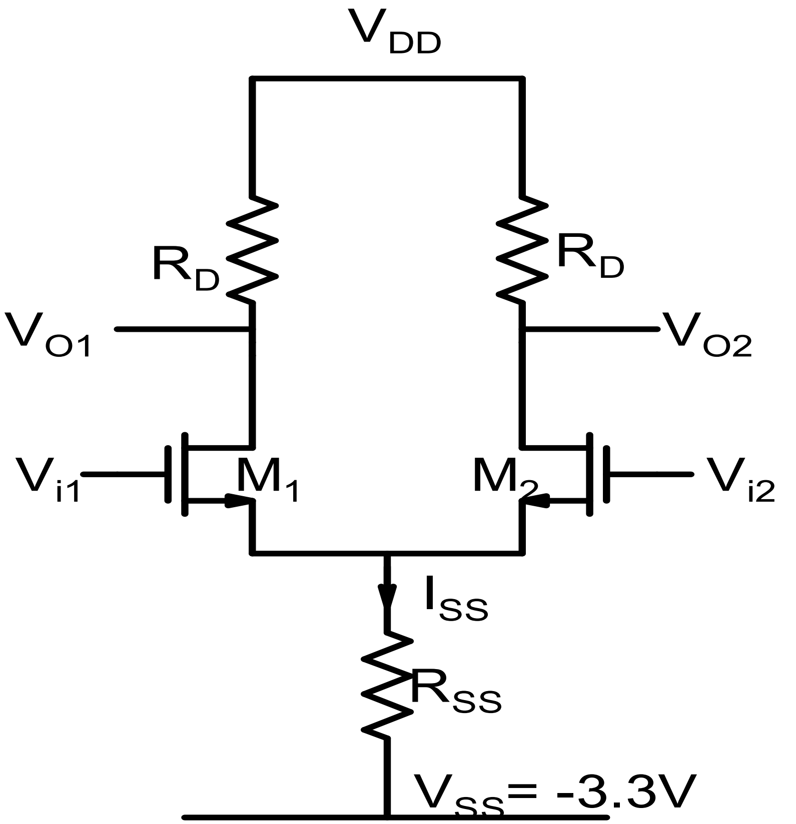
$$CMRR = 0.5 + \frac{2 I_{DSQ} R_{SS}}{V_{GSQ1} - V_T} = 0.5 + \frac{I_{SS} R_{SS}}{V_{GSQ1} - V_T}$$

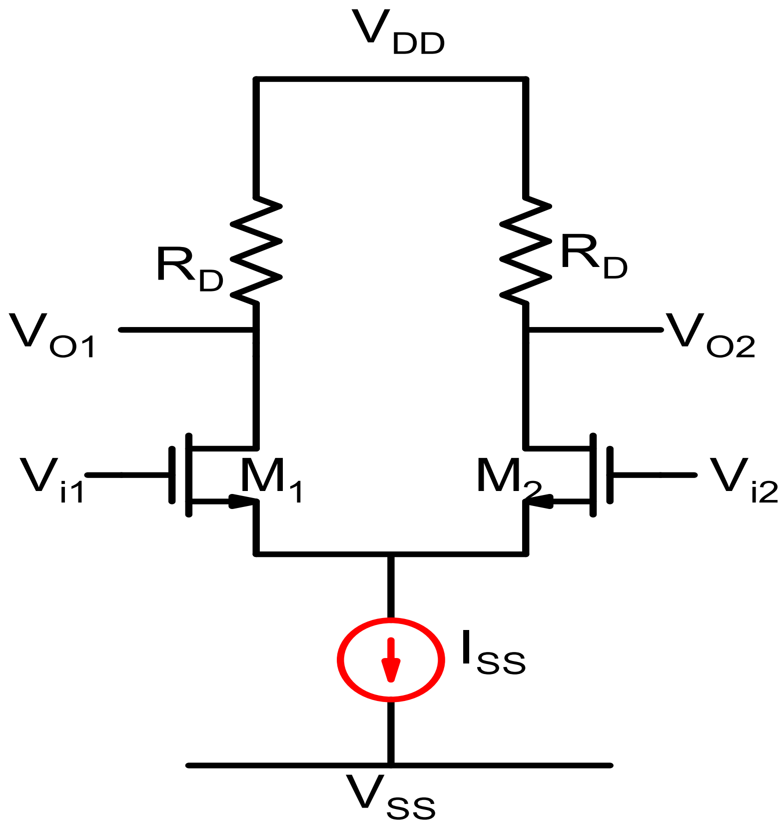
$$CMRR = 0.5 + \frac{-V_{GSQ1} - V_{SS}}{V_{GSQ1} - V_T}$$

Example: $$CMRR = 0.5 + \frac{-1.37 + 3.3}{1.37 - 1.23} = 11.85$$

CMRR is Low !

Common Mode Rejection Ratio

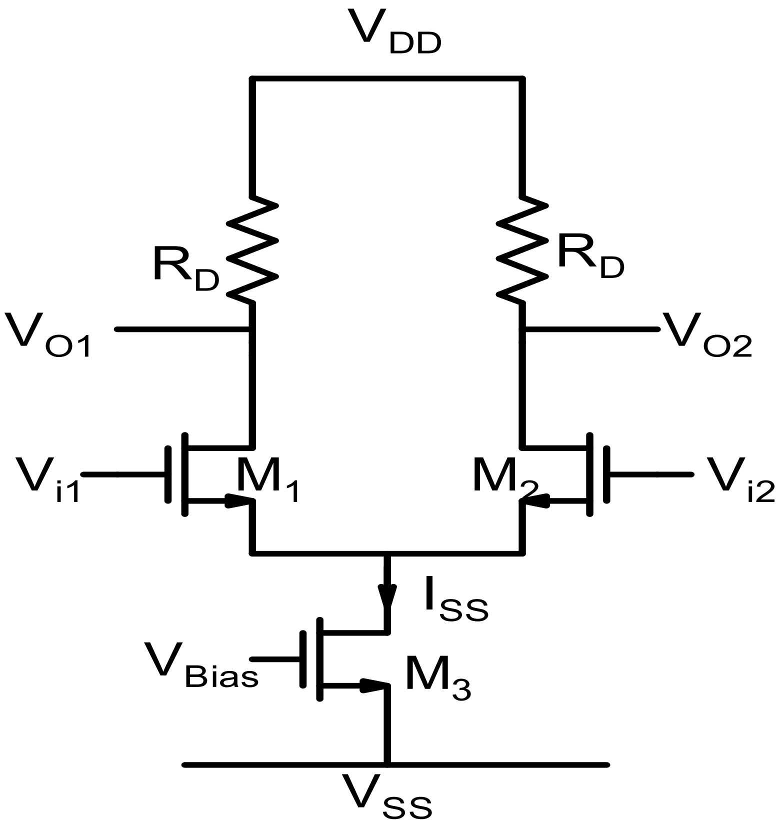




$$CMRR = 0.5 + g_m R_{SS}$$

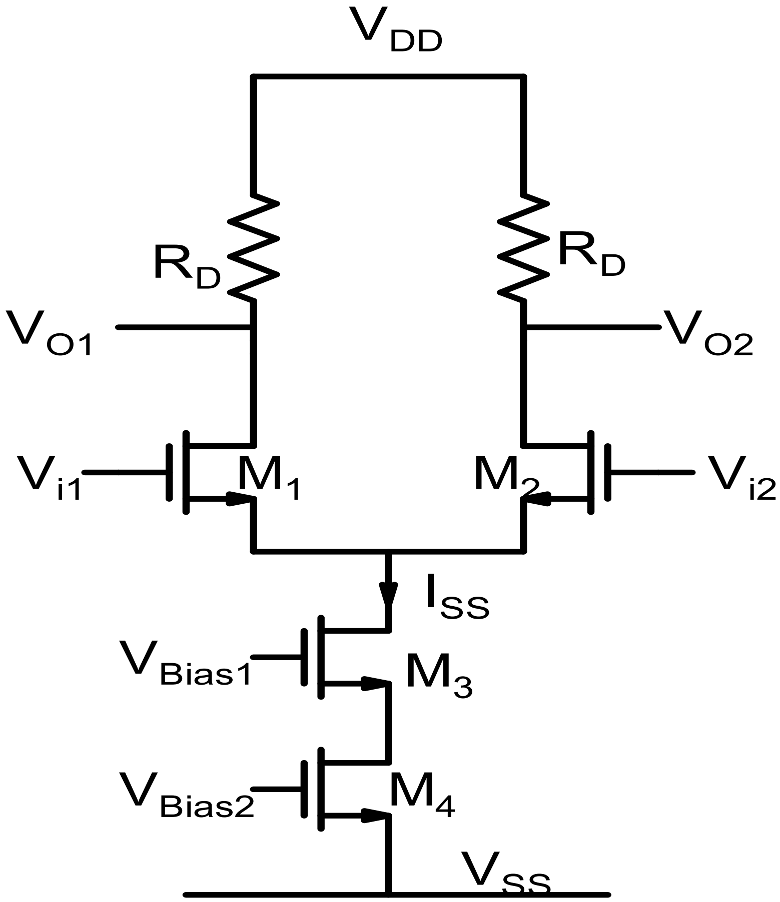
$$CMRR = \infty$$

For a differential amplifier, ideal biasing element is a current source.



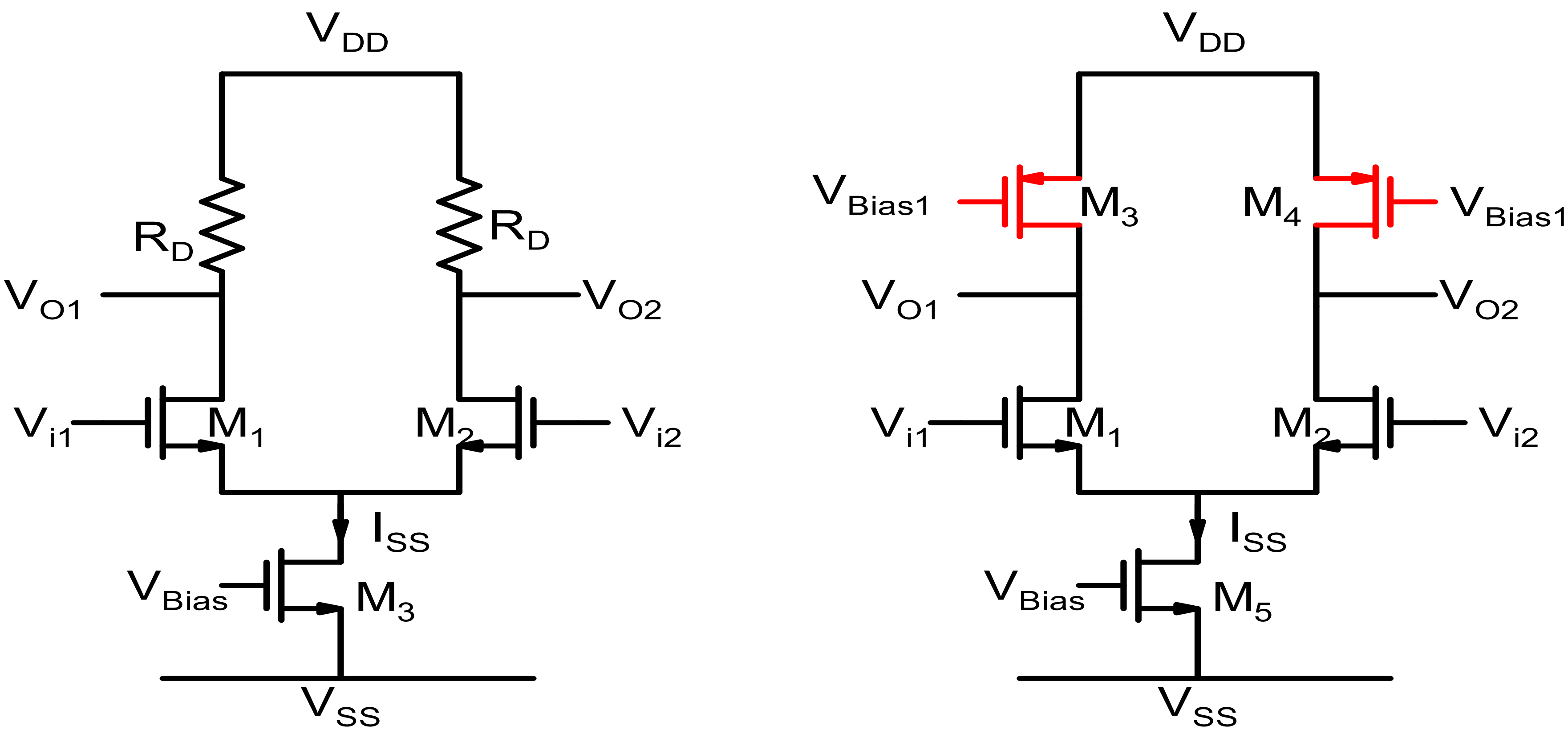
$$CMRR = 0.5 + g_{m1} r_{o3}$$

Very high CMRR can be obtained by building better current sources with higher output resistances



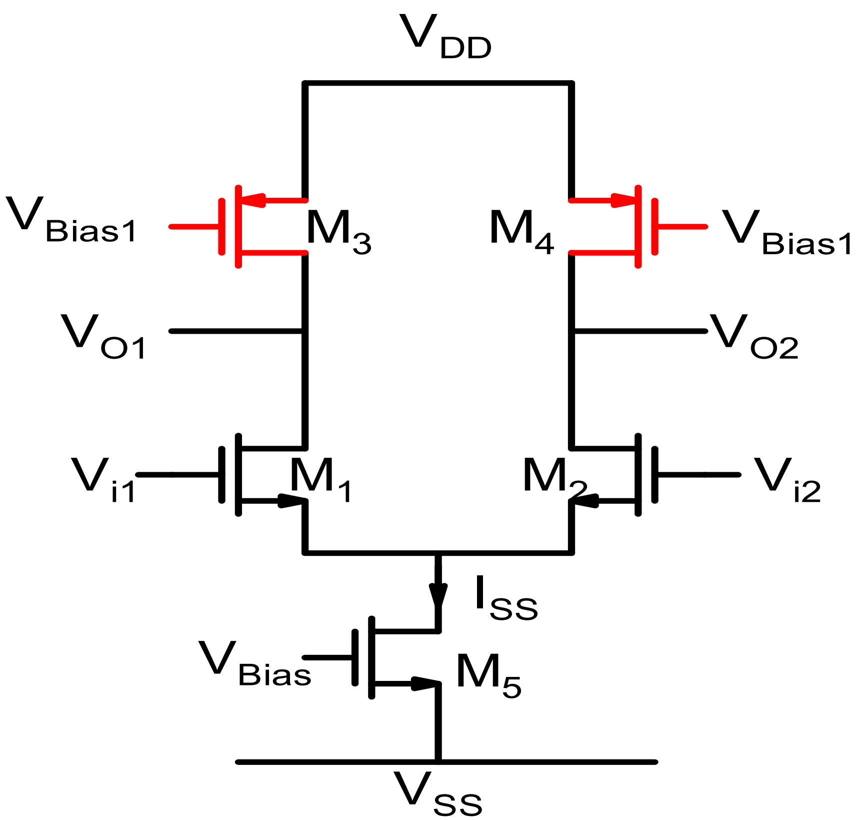
$$CMRR \cong g_{m1} \times (r_{o3} \times g_{m3} r_{o4})$$

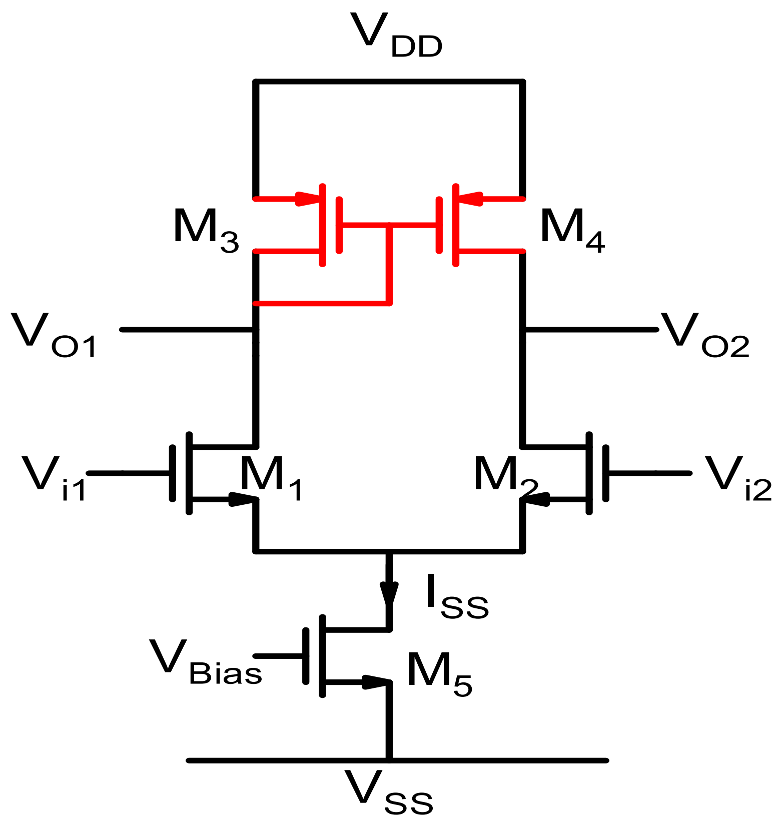
Differential Amplifier with Active Load



The circuit is very sensitive to bias voltage V_{bias1} and requires additional circuit to generate it as well.

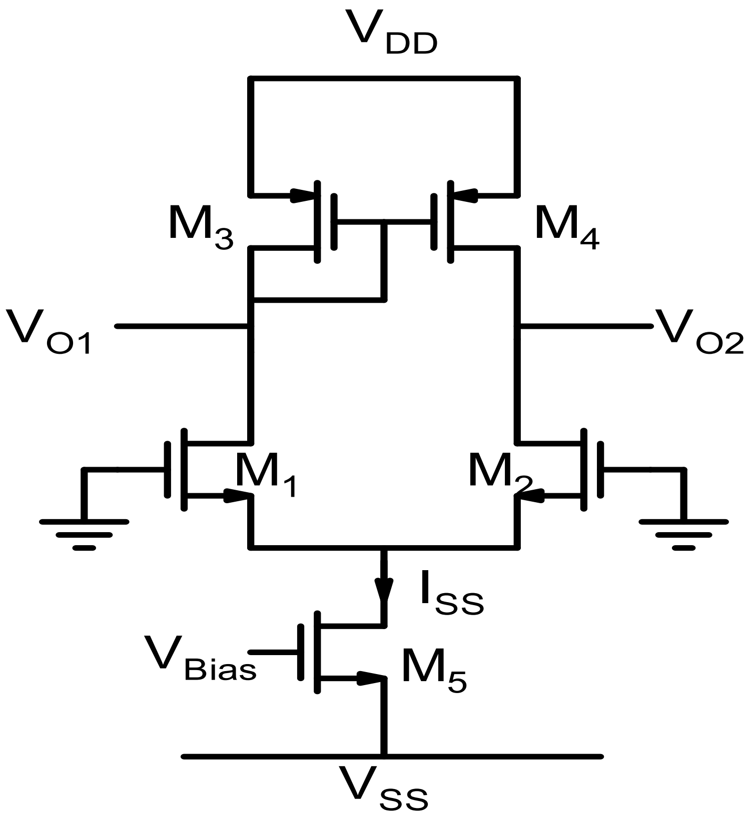
Differential Amplifier with Current Mirror Load





As we shall see later, the new circuit has a well defined dc bias point, does not require additional bias circuitry and other advantages well. However, it is no longer symmetrical and substantial gain is obtained only at V_{o2}

Bias point Analysis



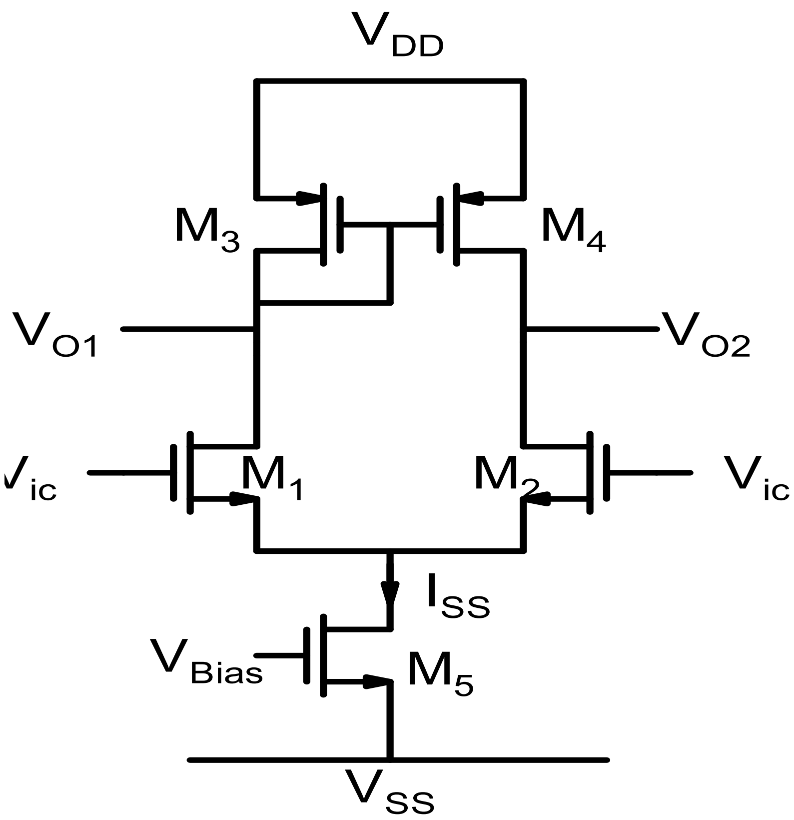
$$I_{SS} = \frac{\beta_5}{2}(V_{Bias} - V_{SS} - V_{TO})^2$$

$$I_{DSQ1} = I_{DSQ2} = 0.5 I_{SS}$$

$$V_{O1Q} = V_{DD} - V_{SG3}$$

$$V_{O1Q} = V_{DD} - |V_{TP}| - \sqrt{\frac{I_{SS}}{\beta_3}}$$

Even though circuit is not symmetrical it can be shown that $V_{o1Q} = V_{o2Q}$



Suppose $V_{o1} > V_{o2}$

$$V_{SD3} < V_{SD4}$$

$$I_{SD3} = \frac{\beta_P}{2}(V_{SG3} + V_{THP})^2(1+\lambda_p V_{SD3}) <$$

$$I_{SD4} = \frac{\beta_P}{2}(V_{SG3} + V_{THP})^2(1+\lambda_p V_{SD4})$$

$$V_{o1} > V_{02} \Rightarrow V_{DS1} > V_{DS2}$$

$$\text{But } I_{DS1} = \frac{\beta_N}{2}(V_{GS1} - V_{THN})^2(1+\lambda_n V_{DS3}) >$$

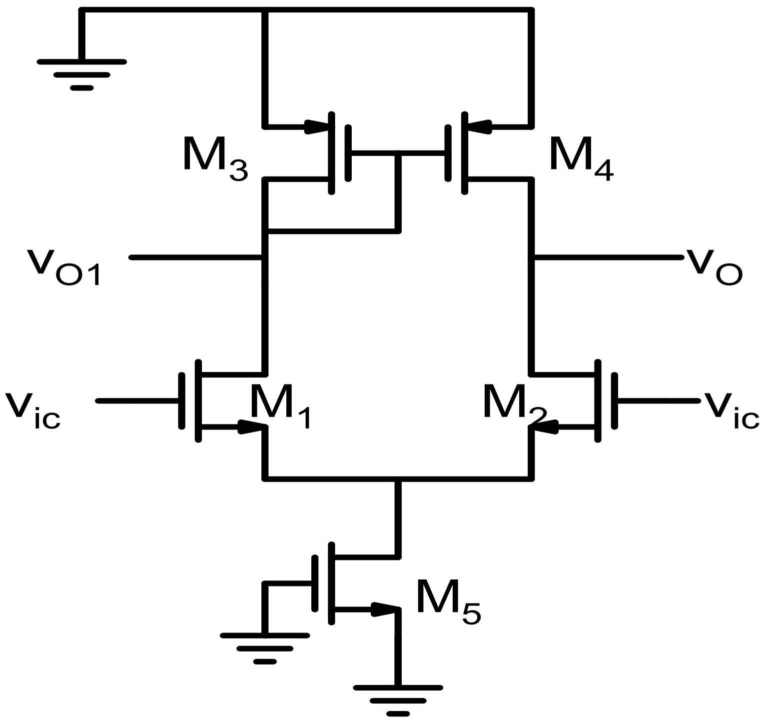
$$I_{DS2} = \frac{\beta_N}{2}(V_{GS1} - V_{THN})^2(1+\lambda_n V_{DS2})$$

$$I_{DS1} = I_{SD3} \;\&\; I_{DS2} = I_{SD4} \Rightarrow I_{SD3} > I_{SD4}$$

Contradiction !! Same occurs if we start with the assumption$V_{o2} > V_{o1}$

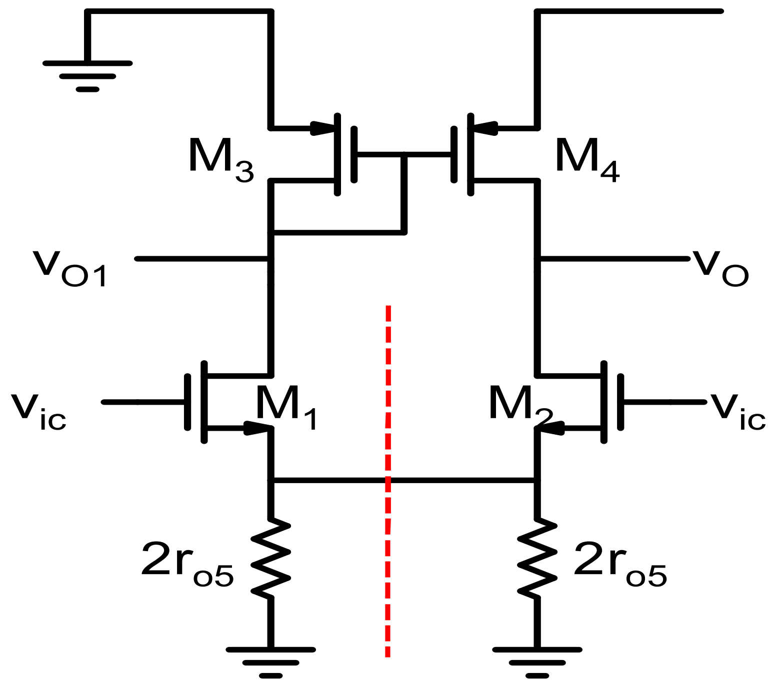
The Only possibility is that both are equal.

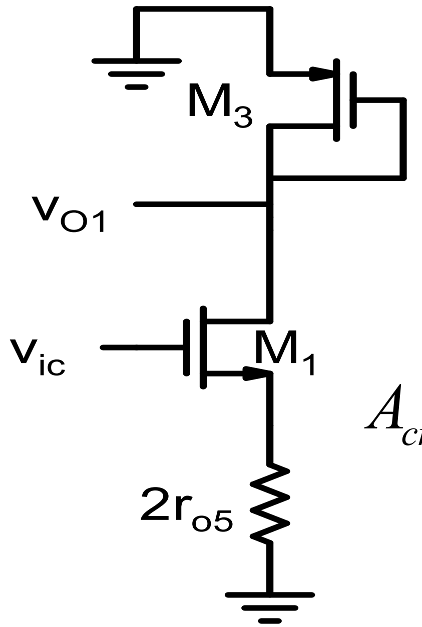
Common Mode Analysis



$$A_{cm} = \frac{v_o}{v_{ic}}$$

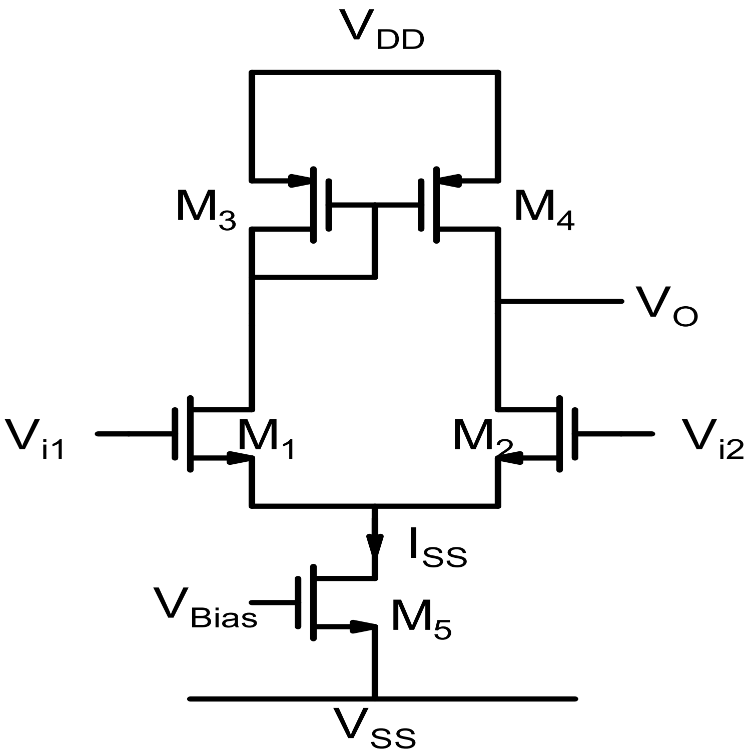
$$v_{o1} = v_0$$



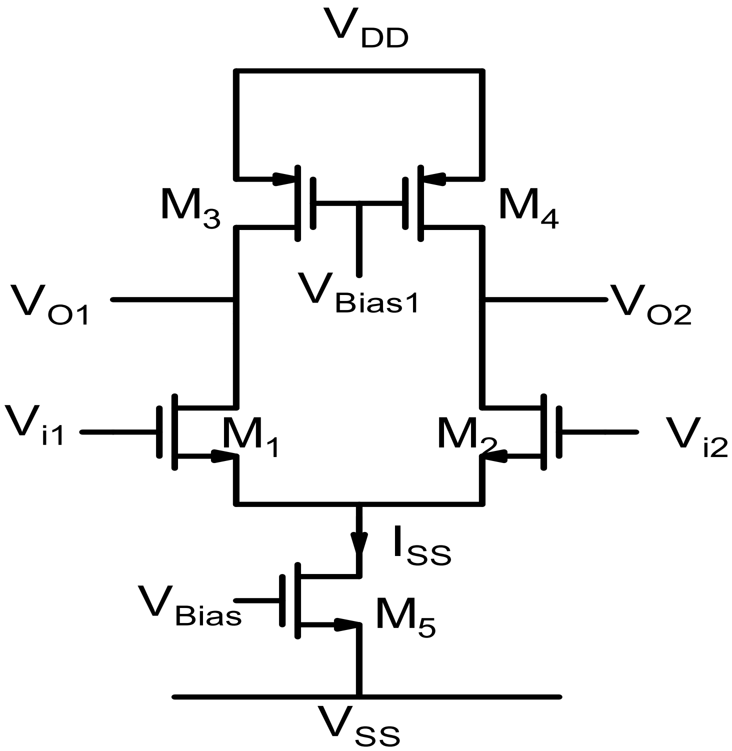


$$A_{cm} = -\frac{g_{m1}}{1 + g_{m1} \times 2r_{o5}} \times \frac{1}{g_{m3}}$$

G-Number



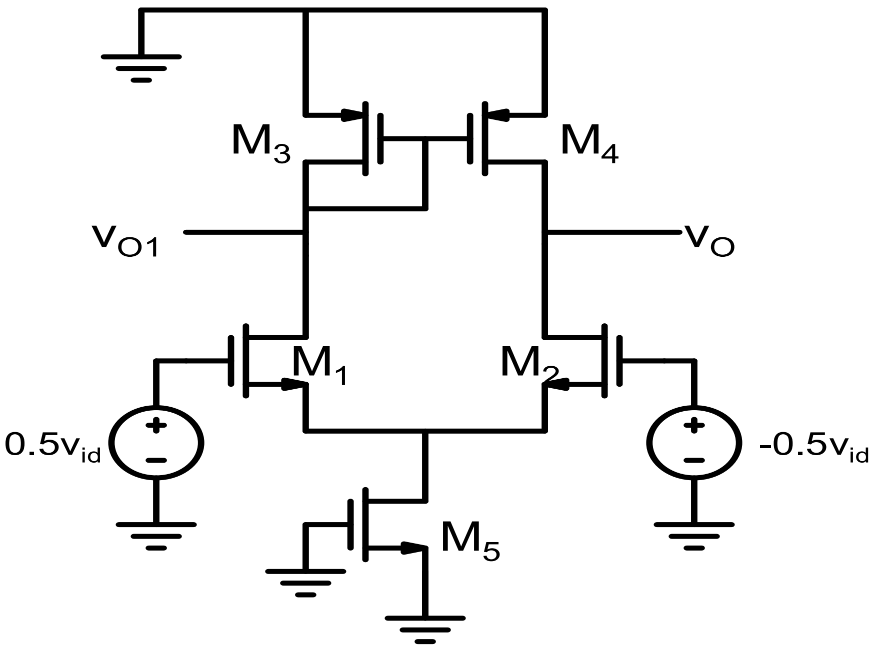
$$A_{cm} = -\frac{g_{m1}}{1+g_{m1}\times 2r_{o5}} \times \frac{1}{g_{m3}}$$

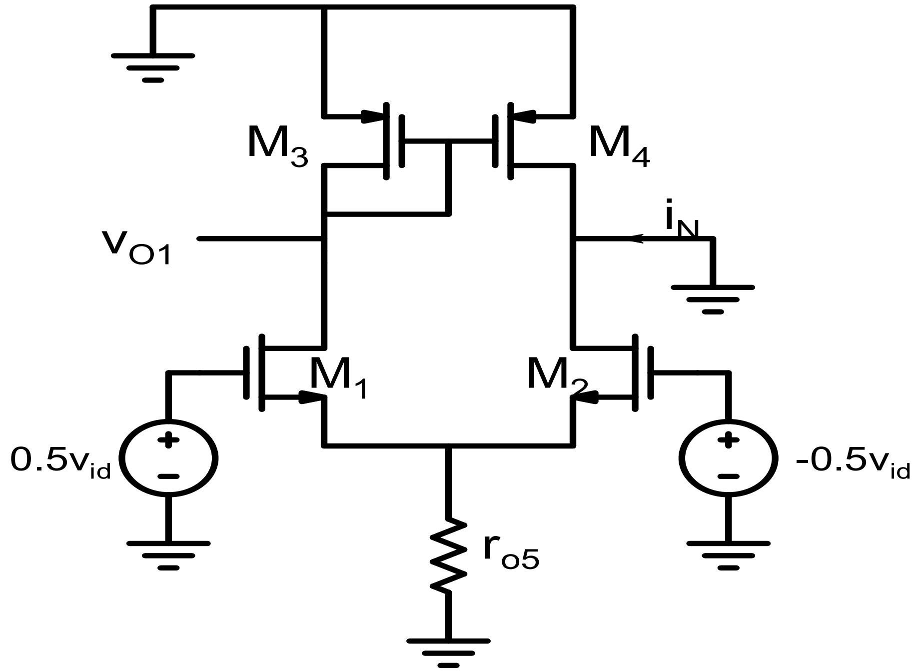


$$A_{cm1} = -\frac{g_{m1}}{1+g_{m1}\times 2r_{o5}} \times r_{o3}$$

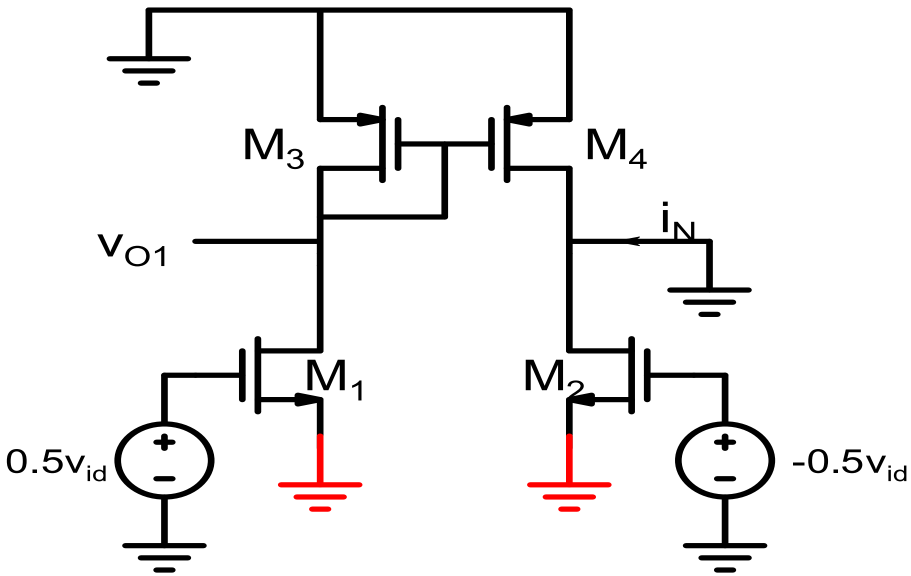
$$\frac{A_{cm1}}{A_{cm}} = g_{m3}\times r_{o3}$$

Differential Mode Analysis

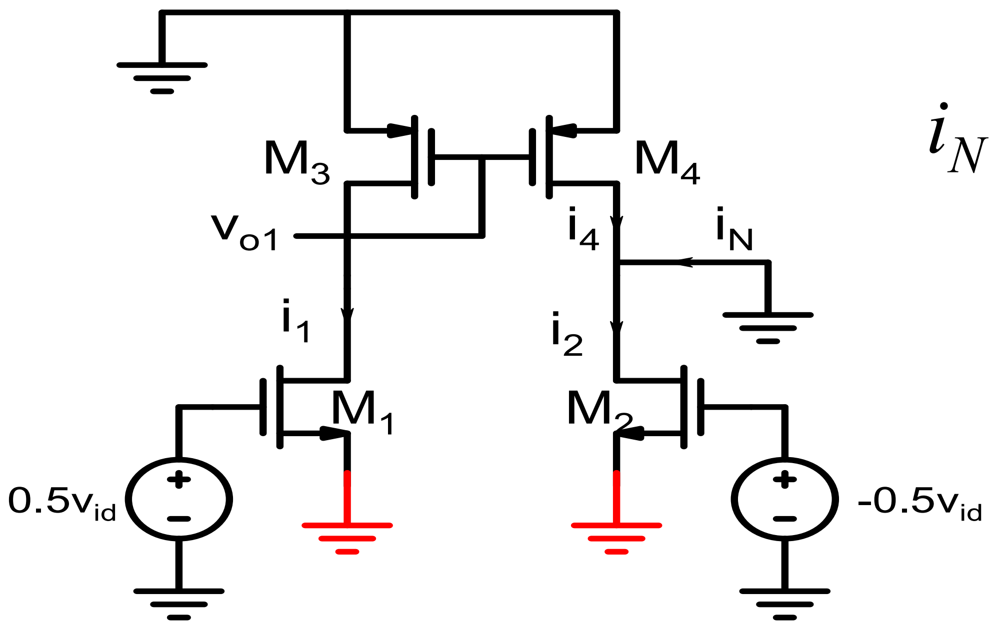




$$v_{o1} \sim 0$$



Norton's Current



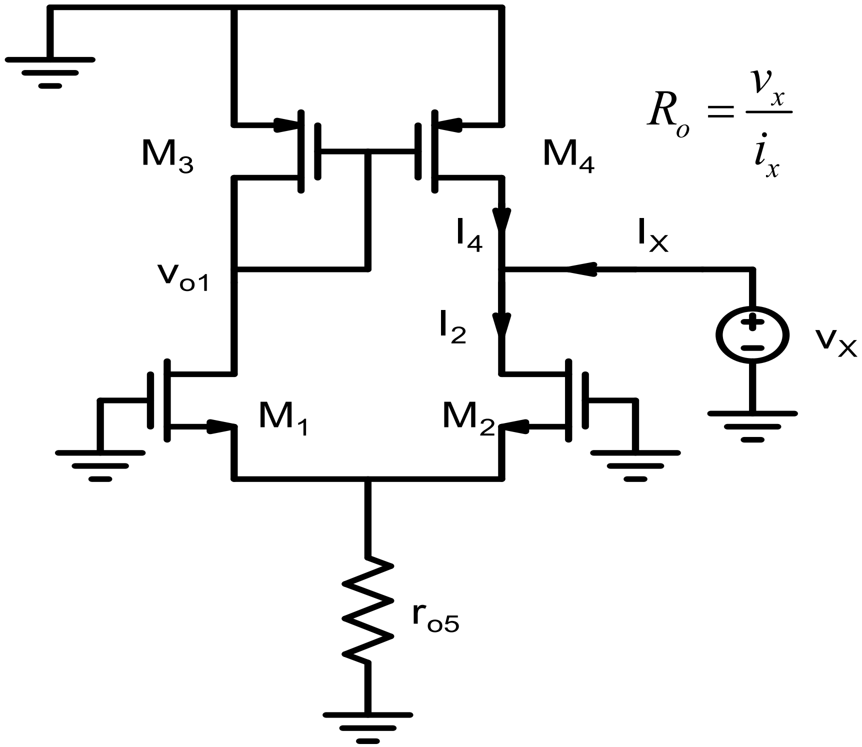
$$i_N = i_2 - i_4 \qquad i_2 = -g_{m2}\frac{v_{id}}{2}$$

$$i_1 = g_{m1}\frac{v_{id}}{2} \qquad v_{o1} = -i_1 \times \frac{1}{g_{m3}}$$

$$i_{ds4} = g_{m4} \times v_{gs4} \qquad i_4 = -i_{ds4} \qquad v_{gs4} = v_{o1} \qquad i_4 = \frac{g_{m4}}{g_{m3}} \times i_1$$

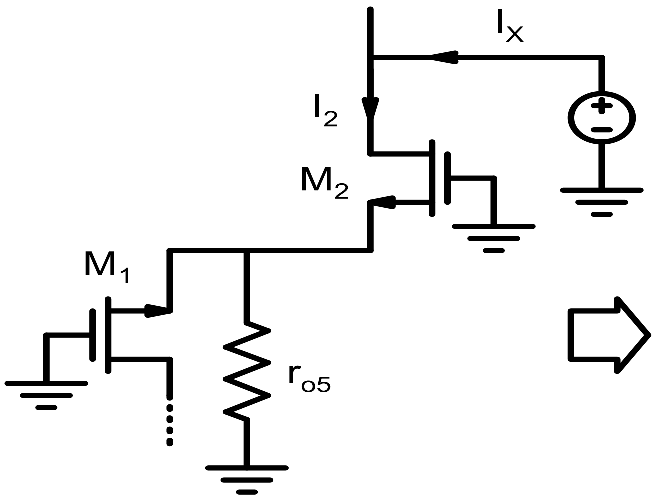
$$g_{m3} = g_{m4} \qquad g_{m1} = g_{m2} \qquad i_N = -g_{m1} \times v_{id}$$

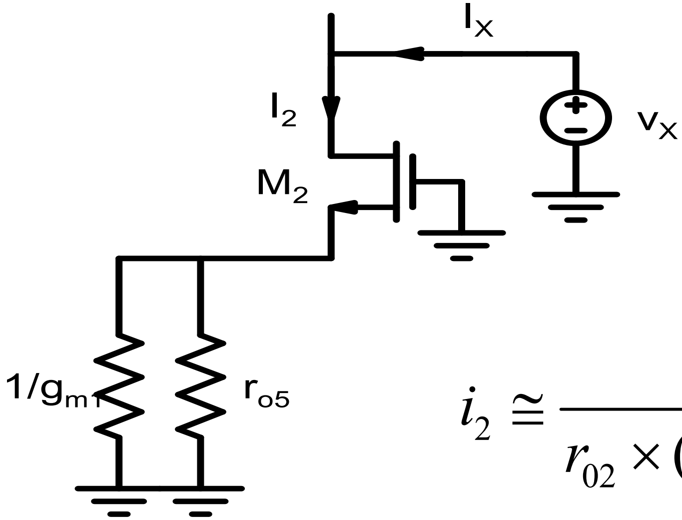
Norton's Resistance



$$R_o = \frac{v_x}{i_x}$$

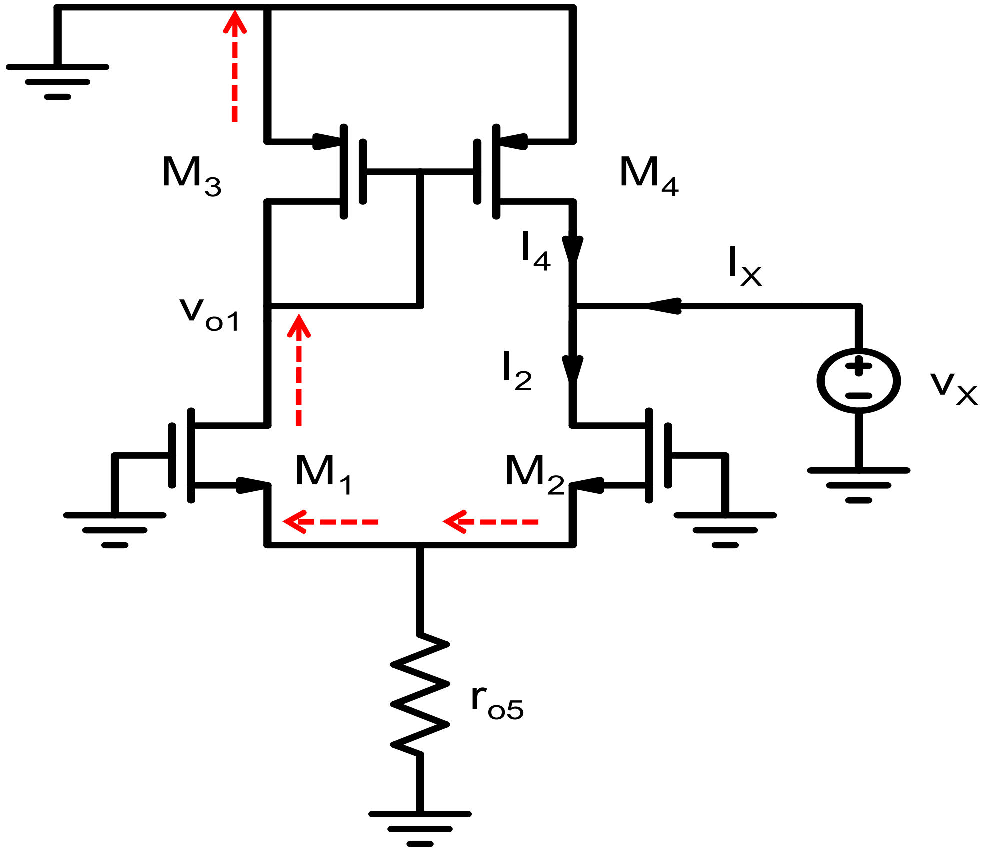
$$i_x = i_2 - i_4$$

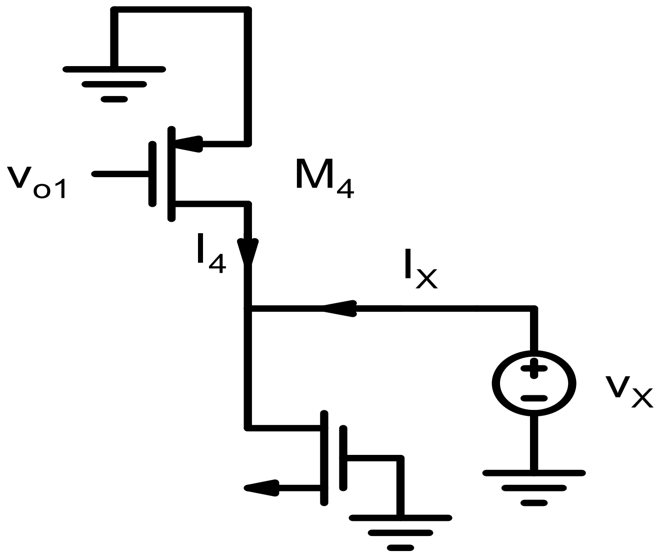




$$i_2 \cong \frac{v_x}{r_{02} \times (1 + g_{m2}/g_{m1})} = \frac{v_x}{2r_{02}}$$

G-Number





$$i_4 = -g_{m4} \times v_{o1} - \frac{v_x}{r_{O4}}$$

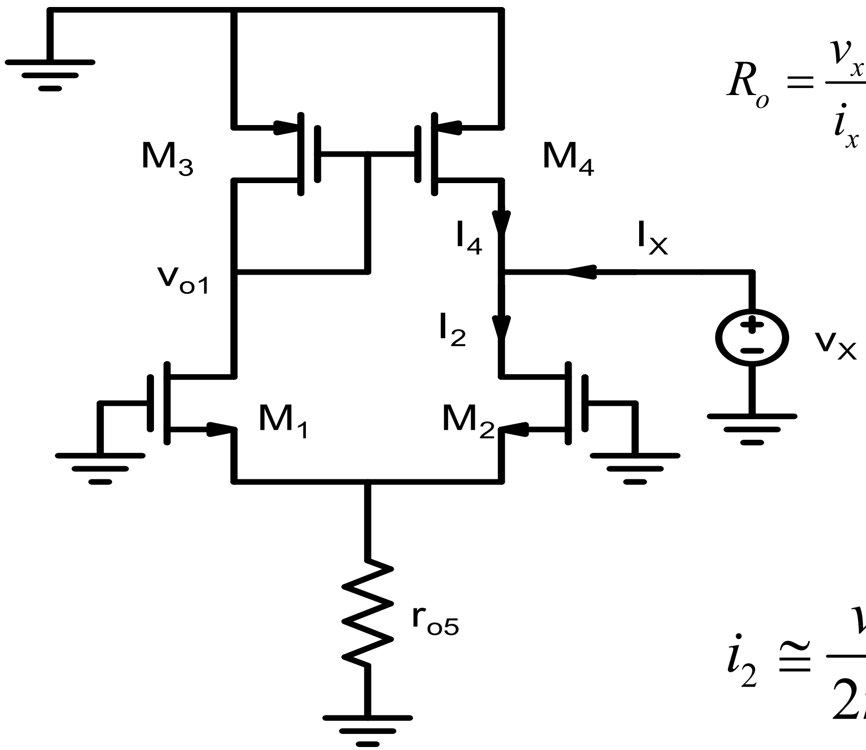
$$i_4 = -g_{m4} \times \frac{i_2}{g_{m3}} - \frac{v_x}{r_{O4}}$$

$$v_{o1} = i_2 \times \frac{1}{g_{m3}}$$

$$i_2 \cong \frac{v_x}{2r_{02}}$$

$$i_4 = -\frac{v_x}{2r_{o2}} - \frac{v_x}{r_{O4}}$$

Norton's Resistance



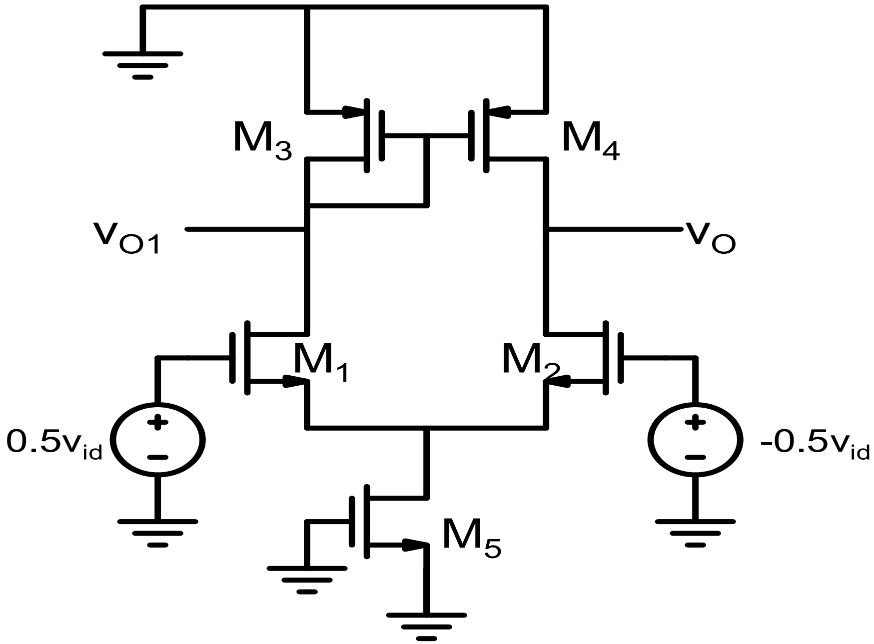
$$R_o = \frac{v_x}{i_x}$$

$$i_x = i_2 - i_4$$

$$i_2 \cong \frac{v_x}{2r_{02}} \qquad i_4 = -\frac{v_x}{2r_{o2}} - \frac{v_x}{r_{O4}}$$

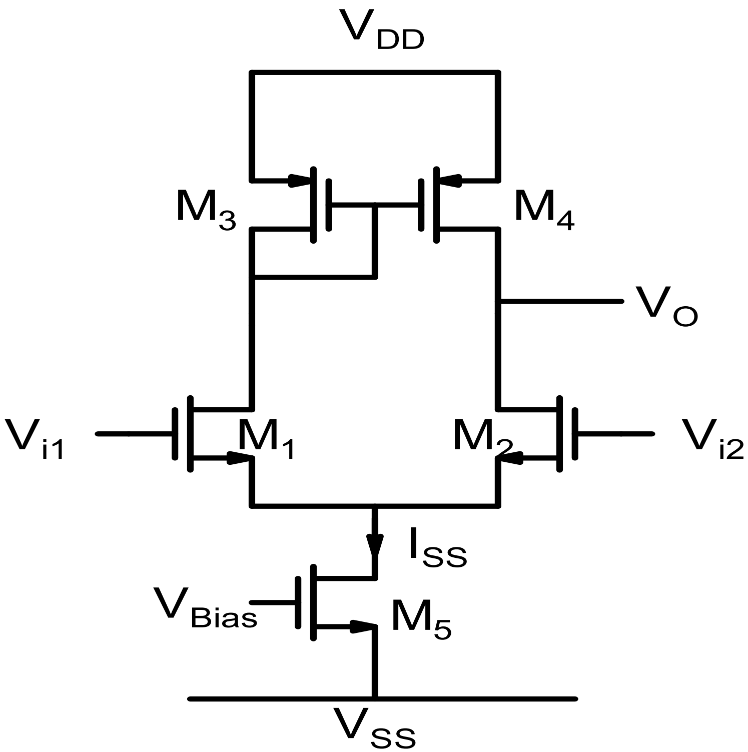
$$i_x = \frac{v_x}{r_{o2}} + \frac{v_x}{r_{O4}}$$

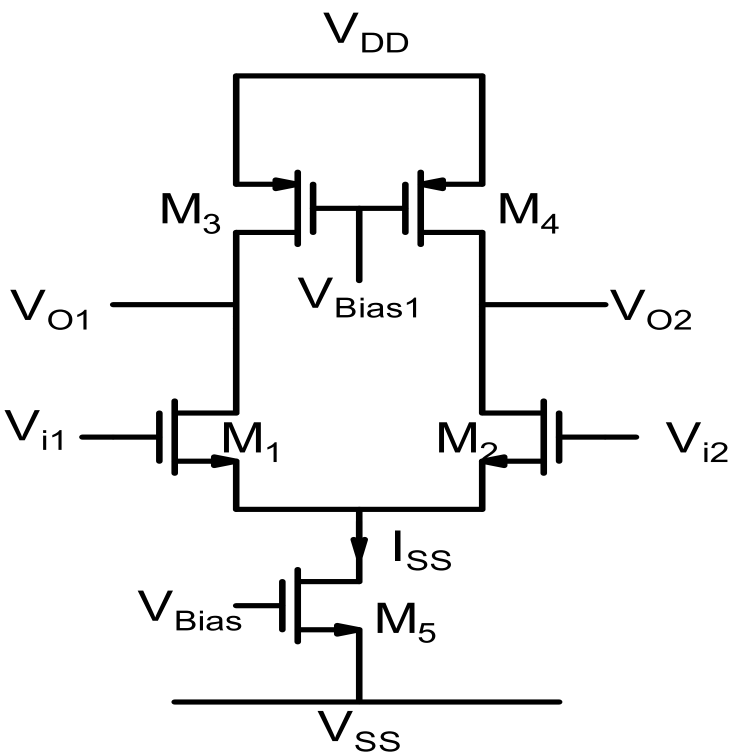
$$R_O = R_N = r_{o2} \| r_{o4}$$



$$R_O = R_N = r_{o2} \| r_{o4} \qquad i_N = -g_{m1} \times v_{id}$$

$$A_{dm} = \frac{v_o}{v_{id}} = +g_m r_{o2} \| r_{o4}$$





$$A_{dm} = \frac{v_o}{v_{id}} = +g_m r_{o2} \| r_{o4}$$

$$A_{cm} = -\frac{g_{m1}/g_{m3}}{1 + g_{m1} \times 2r_{o5}}$$

$$A_{dm} = \frac{v_{o2}}{v_{id}} = +0.5 g_m r_{o2} \| r_{o4}$$

$$A_{cm} = -\frac{g_{m1} r_{o4}}{1 + g_{m1} \times 2r_{o5}}$$

Much higher CMRR is obtained with current mirror load

The END