CMOS ~~CMOS~~ Single stage amplifiers

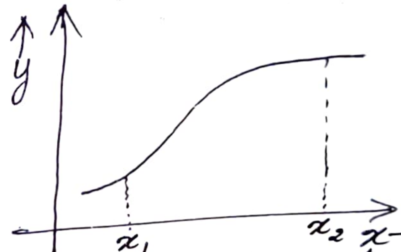
Amplification is an essential function in most analog (and many digital) ckts. We amplify an analog or digital signal because,

a) it may be too small to drive a load.
b) overcome the noise of a subsequent stage or
c) provide logical levels to a digital ckt.
d) Amplification also plays a critical role in feedback systems.

Single stage amplifiers are used in virtually every op-amp design. In this chapter, we study the low freq. behavior of single-stage CMOS amplifiers.

Basic concepts:- The i/p - o/p characteristics of an amplifier is generally a non-linear function (as shown in fig.) that can be approximated by a polynomial over some signal range:

$$y(t) \simeq \alpha_0 + \alpha_1 x(t) + \alpha_2 x^2(t) + \dots + \alpha_n x^n(t); \quad x_1 \le x \le x_2$$



I/p char. of nonlinear system.

The input and output may be current or voltage quantities.

For sufficiently narrow range of x,

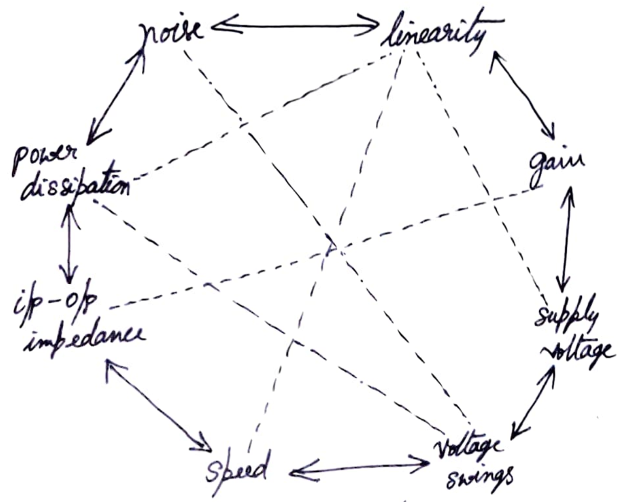
$$y(t) \simeq \alpha_0 + \alpha_1 x(t) \quad \text{—— (2)}$$

where α_0 can be considered the operating point (bias point) and α_1 the small-signal gain. So long as $\alpha_1 x(t) \ll \alpha_0$, the bias point is disturbed negligibly, eqn. (2) provides a reasonable approximation, and higher order terms are insignificant. In other words, $\Delta y = \alpha_1 \Delta x$, indicating a linear relationship between the increments at the i/p and o/p. As $x(t)$ increases in magnitude, higher order terms manifest themselves, leading to non linearity & necessitating large-signal analysis. From another point of view, if the slope of the characteristic (the incremental gain) varies with the signal level, then the system is non linear.

The important performance parameters of an amplifier are:
i) Gain ii) Speed iii) power dissipation iv) supply voltage, v) linearity vi) noise vii) max. voltage swing. viii) I/p and O/p impedance (determines

how the ckt. interacts with preceding and subsequent stages).

In practice, most of these parameters trade with each other, making the design a multidimensional optimization problem. Illustrated in the "analog design octagon" of fig. below, such trade-offs present many challanges in the design of high-performance amplifiers, requiring intuition and experience to arrive at an ~~acceptable~~ acceptable compromise.



Analog Design Octagon

Types of amplifiers :- 1) Common Source Amplifier.
2) Common gate Amplifier.
3) Common ~~[illegible]~~ Drain (Source Follower) Amplifier.

In each case, we begin with a simple model and gradually add second-order phenomena such as channel-length modulation & body effect.

1. Common Source Amplifier :-

a) With resistive load :-

By virtue of its transconductance, a MOSFET converts variation in its gate-source voltage to a small-signal drain current, which can pass through a resistor to generate an o/p voltage. Shown in fig. below a common source (CS) amplifier performs such an operation.

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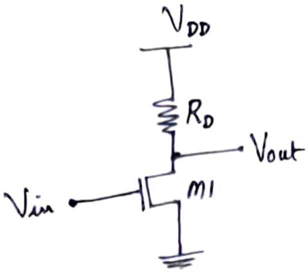
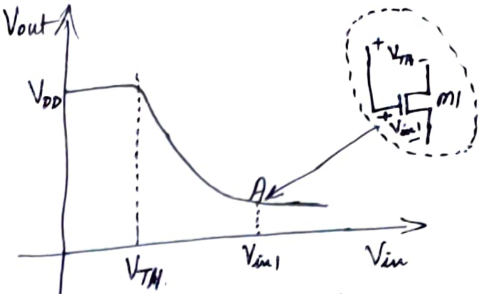
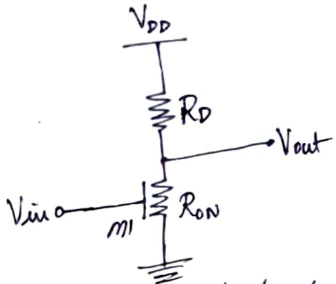


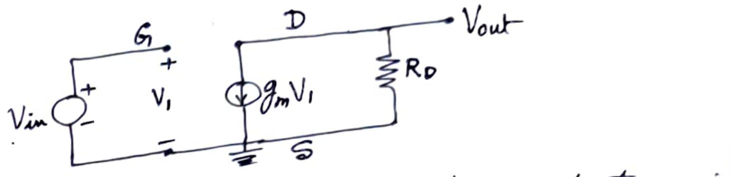
a) CS amp.



b) Input – output chara.



c) Eqv. ckt. in deep triode region



d) Small signal model for the saturation region

We study both the large-signal & the small-signal behavior of the ckt. Note that the input impedance of the ckt. is very high at low frequencies.

If the input voltage increases from zero, M1 is off and $V_{out} = V_{DD}$ [fig (b)]. As V_{in} approaches V_{TH}, M1 begins to turn on, drawing current from R_D and lowering V_{out}. If $V_D > V_G$, M1 is in saturation and we have, $V_{out} = V_{DD} - R_D I_D$

$$V_{out} = V_{DD} - R_D \tfrac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})^2 \quad \text{— (3)}$$

where channel-length modulation is neglected. With further increase in V_{in} exceeds V_{out} by V_{TH} [point A in fig. (b)].

At this point,

$$V_{in1} - V_{TH} = V_{DD} - R_D \tfrac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in1} - V_{TH})^2$$

from which $V_{in1} - V_{TH}$ and hence V_{out} can be calculated.

For $V_{in} > V_{in1}$, M1 is in the triode region:

$$V_{out} = V_{DD} - R_D \tfrac{1}{2}\mu_n C_{ox}\frac{W}{L}\left[2(V_{in} - V_{TH})V_{out} - V_{out}^2\right].$$

If V_{in} is high enough to drive M1 into deep triode region, $V_{out} \ll 2(V_{in} - V_{TH})$, and, from the equivalent ckt. of fig. (c).

$$V_{out} = V_{DD}\frac{R_{on}}{R_{on} + R_D}$$

We have,

$$I_D = \mu_n C_{ox} \frac{W}{L}\left[(V_{GS}-V_{TH})V_{DS} - \frac{1}{2}V_{DS}^2\right]$$

In the deep triode region, $V_{DS} \ll 2(V_{GS}-V_{TH})$ [Here $V_{DS} = V_{out}$, $V_{GS} = V_{in}$]

$$I_D \approx \mu_n C_{ox} \frac{W}{L}(V_{GS}-V_{TH})V_{DS}.$$

or $$R_{ON} = \frac{V_{DS}}{I_D} = \frac{1}{\mu_n C_{ox} W/L\,(V_{GS}-V_{TH})}.$$

Here,

$$V_{out} = V_{DD}\cdot\frac{R_{ON}}{R_{ON}+R_D} = \frac{V_{DD}}{\frac{R_{ON}+R_D}{R_{ON}}} = \frac{V_{DD}}{1+\frac{R_D}{R_{ON}}}.$$

$$= \frac{V_{DD}}{1+R_D\cdot\mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH})}$$

Since the transconductance drops in the triode region, we usually ensure that $V_{out} > V_{in} - V_{TH}$, operating to the left of point A in fig. (c). Using eqn. (3) as the input-output characteristics and viewing its slope as the small signal gain, we have,

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} = -\underbrace{R_D\mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH})}\ \text{[from eqn. (3)]} \quad (4)$$

$$= -g_m R_D.$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}}$$

This result can be directly derived from the observation that M1 converts an input voltage change ΔV_{in} to a drain current change $g_m \Delta V_{in}$, & hence an o/p voltage change $-g_m R_D \Delta V_{in}$. or the gain $A_v = \frac{\Delta V_{out}}{\Delta V_{in}} = -g_m R_D$. The small signal model of fig. (d) yeilds the same result.

Note that g_m varies with the drain current & hence the i/p signal, the gain of the ckt. changes significantly for large input swings, leading to non linearity.

Maximizing the voltage gain :-

From eqn. (4), $A_v = -R_D \mu_n C_{ox} \frac{W}{L}(V_{in} - V_{TH})$

This can be written as, $A_v = -\sqrt{2 \cdot \frac{1}{2} \mu_n^2 C_{ox}^2 \left(\frac{W}{L}\right)^2 (V_{gs} - V_t)^2} \cdot R_D$

$$A_v = -\sqrt{2 \mu_n C_{ox} \frac{W}{L} I_D} \; \frac{V_{RD}}{I_D}$$ where V_{RD} = voltage drop across R_D.

$$\text{or, } A_v = -\sqrt{2 \mu_n C_{ox} \frac{W}{L}} \; \frac{V_{RD}}{\sqrt{I_D}}$$

Thus the magnitude of A_v can be increased by increasing W/L or V_{RD} or decreasing I_D if other parameters are constant.

Trade-offs:

i) Increasing W/L, increases device size which leads to greater device capacitances.

ii) A higher V_{RD} limits the max. voltage swings.

For example: $V_{DS} = V_{DD} - V_{RD} = V_{in} - V_{TH}$, then M1 is at the edge of the triode region, allowing only very small swings at the o/p (and i/p).

iii) If V_{RD} remains constant, and I_D is reduced, then R_D must increase, thereby leading to a greater time constant at the o/p node.

Lower supply voltages further tighten these trade-offs.

For large values of R_D, the effect of channel length modulation in M1 becomes significant.

$\therefore$ Eqn. (3) becomes,

$$V_{out} = V_{DD} - R_D \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{in} - V_{TH})^2 (1 + \lambda V_{out})$$

$$\text{or } \frac{\partial V_{out}}{\partial V_{in}} = -R_D \mu_n C_{ox} \frac{W}{L} (V_{in} - V_{TH})(1 + \lambda V_{out}) - R_D \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{in} - V_{TH})^2 \lambda \frac{\partial V_{out}}{\partial V_{in}}$$

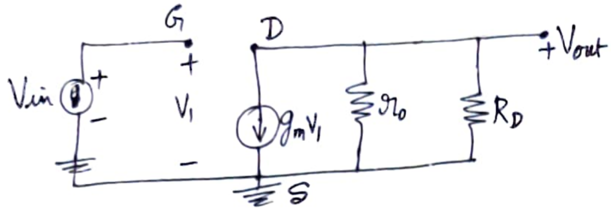
Using the approximation, $I_D \approx \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{in} - V_{TH})^2$ we get,

$$A_v = -R_D g_m - R_D I_D \lambda A_v$$

$$\text{or, } A_v = -\frac{g_m R_D}{1 + R_D \lambda I_D}$$

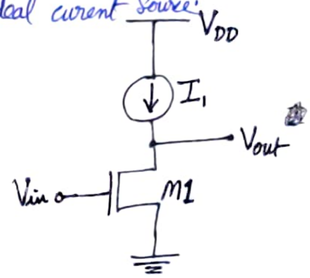
Since $\lambda I_D = 1/r_o$, $\boxed{A_v = -g_m \cdot \frac{r_o R_D}{r_o + R_D}}$

The small signal model of fig. below gives the same result with much less effort. That is, since $g_m V_1 (r_o \| R_D) = -V_{out}$, and $V_1 = V_{in}$, we have $V_{out}/V_{in} = -g_m (r_o \| R_D)$.



$$V_{out} = -g_m V_1 (r_o \| R_D).$$

Example:- 1) Calculate the small-signal voltage gain of the ckt. Assume that M1 is biased in saturation & I_1 is an ideal current source.



$$A_v = -g_m \frac{r_o R_D}{r_o + R_D}$$

Since I_1 introduces an infinite impedance,

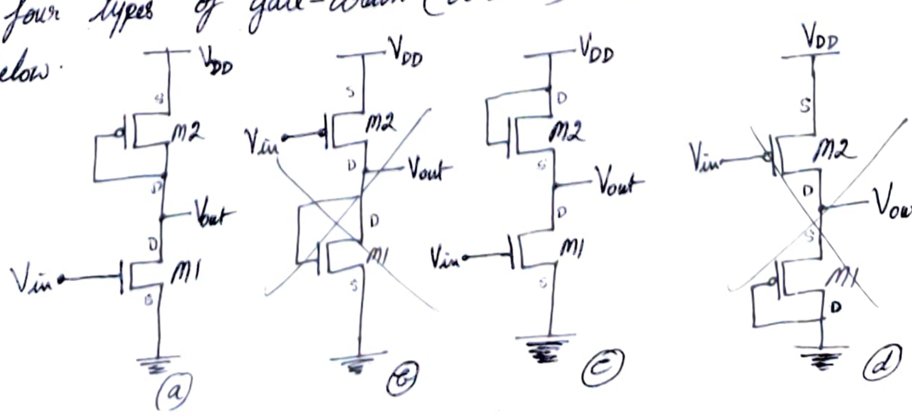
$$A_v = -g_m r_o.$$

This gain is called the "intrinsic gain" of a transistor, this quantity represents the max. voltage gain that can be achieved using a single device. In short channel devices, this, $g_m r_o$ ranges between roughly 10 and 30.

6) CS amplifier with diode (active) connected load:-

In many CMOS technologies, it is difficult to fabricate resistors with tightly-controlled values and they take large amount of chip area. Consequently, it is desirable to replace R_D in fig. ⓐ with a MOS transistor. An active load can also produce higher values of resistance when compared with a passive resistor, resulting in higher gains.

The four types of gate-drain (or diode) connected load amp. are shown below.



All 4 of these configurations are basic common source voltage amplifiers. (4)
In each configuration, M1 and M2 are assumed to be biased in the saturation region.

Now consider fig. (c) above and repeated here again.

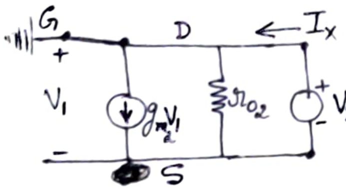


fig. (b)

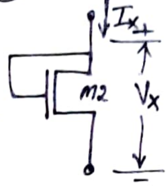
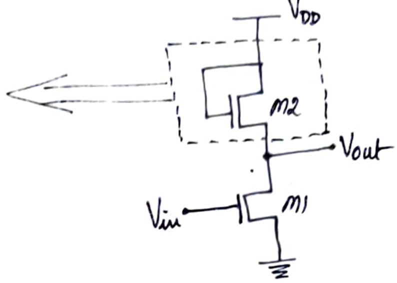


fig. (a)



A MOSFET can operate as a small-signal resistor if its gate and drain are shorted [fig (a)]. This is called a "diode-connected" device. This configuration exhibits a small-signal behavior similar to a 2-terminal resistor. Note that the transistor is always in saturation because the drain and gate have the same potential.

Using the small-signal equivalent shown in fig (b) above to obtain the impedance of the device;

We have, $V_1 = V_X$

$$I_X = \frac{V_X}{r_{o2}} + g_{m_2} V_X = V_X\left(1/r_{o2} + g_{m_2}\right).$$

That is, the impedance of the diode is,

$$Z = 1/g_{m_2} \,||\, r_{o2} \approx \underline{\underline{1/g_{m_2}}} \quad \Big| \because 1/g_{m_2} \ll r_{o2}$$

If body effect exists, then from the fig. below, we have,

$V_1 = -V_X$;

$V_{b2} = -V_X$

And, $(g_{m_2} + g_{mb_2})V_X + \frac{V_X}{r_{o2}} = I_X$

or, $$\frac{V_X}{I_X} = \frac{1}{\frac{1}{(g_{m_2} + g_{mb_2})} + \frac{1}{r_{o2}}}$$

$$= \frac{1}{g_{m_2} + g_{mb_2}} \,||\, r_{o2}$$

$$\approx \underline{\underline{\frac{1}{g_{m_2} + g_{mb_2}}}}$$

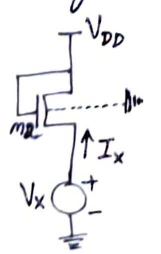


fig (a)

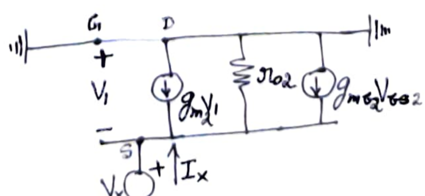
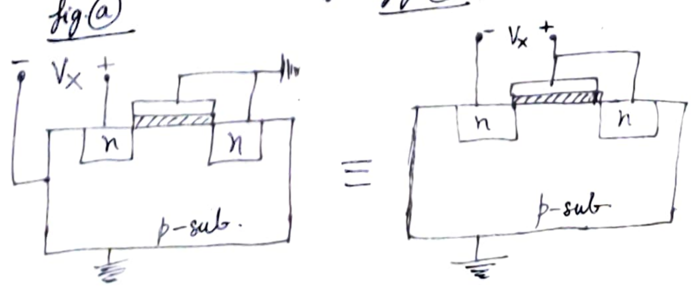


fig (b).



This shows that, the impedance seen at the source of M1 is lower when body effect is included.

Now the gain of the CS amp. with this diode connected load is,

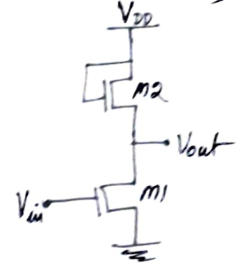


Fig. (a) CS amp.

$$A_v = -g_{m1} R_D$$

$$= -g_{m1} \cdot \frac{1}{g_{m2} + g_{mb2}} = -g_{m1} \cdot \frac{1}{g_{m2}\left(1 + \frac{g_{mb2}}{g_{m2}}\right)}$$

$$= -\frac{g_{m1}}{g_{m2}} \cdot \frac{1}{1+\eta} \quad \text{where } \eta = \frac{g_{mb2}}{g_{m2}}$$

Expressing g_{m1} and g_{m2} in terms of device dimensions and bias currents we have,

$$A_v = -\frac{\sqrt{2\mu_n C_{ox} (W/L)_1 I_{D1}}}{\sqrt{2\mu_n C_{ox} (W/L)_2 I_{D2}}} \cdot \frac{1}{1+\eta}$$

and, since $I_{D1} = I_{D2}$,

$$\boxed{A_v = -\sqrt{\frac{(W/L)_1}{(W/L)_2}} \cdot \frac{1}{1+\eta}}$$

This eqn. reveals an interesting property: if the variation of η with the o/p voltage is neglected, the gain is independent of the bias currents and voltages (so long as M1 stays in saturation). In other words, as the input and o/p signal levels vary, the gain remains relatively constant, indicating that the i/p-o/p characteristic is relatively linear.

The linear behavior of the ckt. can also be confirmed by large signal analysis. Neglecting channel-length modulation for simplicity, we have in fig. above: $I_{D1} = I_{D2}$

$$\frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_1 (V_{in} - V_{TH1})^2 = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_2 (V_{DD} - V_{out} - V_{TH2})^2$$

and hence, $\sqrt{\left(\frac{W}{L}\right)_1} (V_{in} - V_{TH1}) = \sqrt{\left(\frac{W}{L}\right)_2} (V_{DD} - V_{out} - V_{TH2})$

Differentiating both sides w.r.t. V_{in},

$$\sqrt{\left(\frac{W}{L}\right)_1} = \sqrt{\left(\frac{W}{L}\right)_2}\left(-\frac{\partial V_{out}}{\partial V_{in}} - \frac{\partial V_{TH2}}{\partial V_{in}}\right)$$

$V_{TH} = V_{TH0} + \gamma\left[\sqrt{2\phi_F + V_{SB}} - \right.$

which, upon application of the chain rule, $\frac{\partial V_{TH2}}{\partial V_{in}} = \frac{\partial V_{TH2}}{\partial V_{out}} \cdot \frac{\partial V_{out}}{\partial V_{in}}$ ⑤

$$= \eta \cdot \frac{\partial V_{out}}{\partial V_{in}}$$

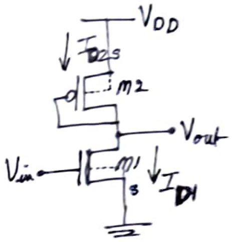
Apply this in the above expn. reduces to,

$$\sqrt{(W/L)_1} = \sqrt{(W/L)_2}\left(-\frac{\partial V_{out}}{\partial V_{in}} - \eta \frac{\partial V_{out}}{\partial V_{in}}\right)$$

$$\sqrt{\frac{(W/L)_1}{(W/L)_2}} = -A_v(1+\eta)$$

or, $$A_v = -\sqrt{\frac{(W/L)_1}{(W/L)_2}} \cdot \frac{1}{1+\eta} \quad \text{—— ⑤}$$

If the diode-connected load is implemented with a PMOS device, the small-signal voltage gain is,



$$A_v = -\sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}} \quad \text{—— ⑥}$$

The ckt. is free from body effect.

where channel-length modulation is neglected.

Eqns. ⑤ & ⑥ indicate that the gain of a CS amp. with diode connected load is relatively a weak function of the device dimensions. For example, to achieve a gain of 10, $\mu_n (W/L)_1 / \mu_p (W/L)_2 = 100$, implying that, with $\mu_n \approx 2\mu_p$, we must have $(W/L)_1 \approx 50 (W/L)_2$.

In a sense, a high gain requires a "strong" input device and a "weak" load device.

Disadvantages:- ① Disproportionately wide or long transistors, & hence a large input or load capacitance.

② A high gain translates to another important limitation: i.e, reduction in allowable voltage swings.

Since $I_{D1} = |I_{D2}|$, $\mu_n \left(\frac{W}{L}\right)_1 (V_{GS1} - V_{TH1})^2 \approx \mu_p (W/L)_2 (V_{GS2} - V_{TH2})^2$

or $$\sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}} = \frac{|V_{GS2} - V_{TH2}|}{V_{GS1} - V_{TH1}}$$

or, $$\boxed{A_v = \frac{|V_{GS2} - V_{TH2}|}{V_{GS1} - V_{TH1}}}$$

In the above example, ∴ the over drive voltage of M2 must be 10 times that of M1. For example, with $V_{GS1} - V_{TH1} = 200\text{ mV}$ and $V_{TH2} = 0.7\text{V}$, we have $|V_{GS2}| = 2.7\text{ V}$, severely limiting the o/p swing. Note that, the diode-connected loads, the swing is constrained by both the required overdrive voltage and the threshold voltage. That even with a small overdrive the o/p level cannot exceed $V_{DD} - |V_{GS2}|$.

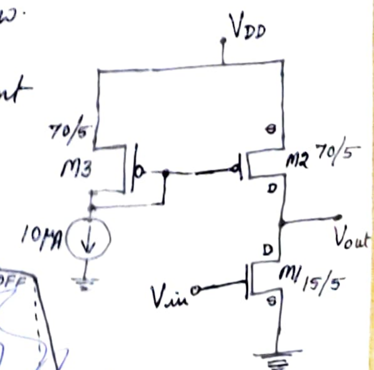
We should also mention that in today's CMOS technology, channel-length modulation is quite significant and, more importantly, the behavior of transistors notably departs from the square law. Thus, the gain of the amp. of fig. above must be expressed as,

$$A_v = -g_{m1}\left(\frac{1}{g_{m2}} \,\|\, r_{o1} \,\|\, r_{o2}\right).$$

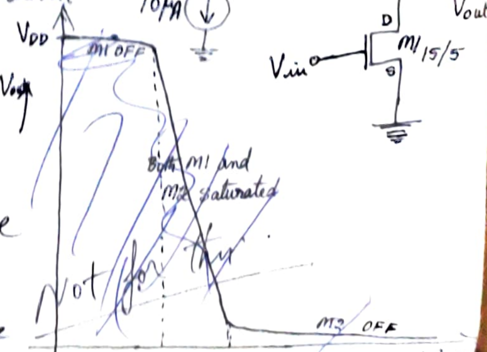
CS amplifier with current source load :- Razavi, page: 55.

In applications requiring a large voltage gain in a single stage the relationship $A_v = -g_m R_D$ suggests that we increase the load impedance of the CS stage. With a resistor or diode connected load however, increasing the load resistance limits the o/p voltage swing.

A more practical approach is to replace the load with a current source. The current source load provides an amplifier with the largest possible load resistance available in the CMOS process. The resulting ckt. is shown in the fig. below.

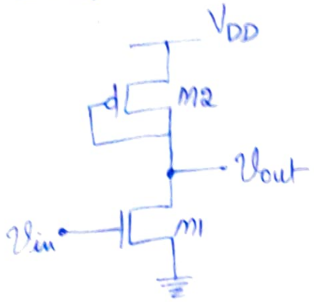


The MOSFET M1 is the common source component of the amplifier, while the M2 is the current source load. ~~The DC transfer char are shown for this amplifier.~~

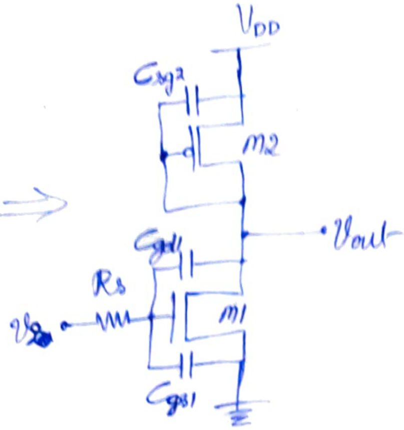


~~The slope of the line when both transistors are saturated corresponds to the small-signal gain of the amplifier. If we bias M1 and M2 so that they are both saturated, we see that the DC o/p voltage~~

Freqr. response :-



$\Rightarrow$



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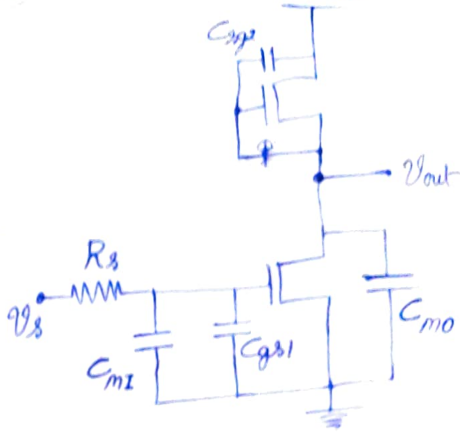


After applying Miller's theorem $\left\{ \begin{aligned} C_{mI} &= C_{gd1}(1+|A_v|) \\ C_{mo} &= C_{gd1}(1+1/|A_v|) \end{aligned} \right.$

$$A_v = \frac{-g_{m1}}{g_{m2}}$$

The 2 RC time constants exists in this circuit: one on the input of the circuit and one on the output of the circuit.

They are, $\tau_{in} = R_s(C_{mI} + C_{gs1})$ where $C_{mI} = C_{gd1}\left(1+\frac{g_{m1}}{g_{m2}}\right)$

$$\tau_{out} = \frac{1}{g_{m2}}\left(C_{sg2} + C_{mo}\right) \text{ where } C_{mo} = C_{gd1}\left(1+\frac{g_{m2}}{g_{m1}}\right)$$

The freqr. response of the amplifier is,

$$A_v(f) = \frac{\frac{-g_{m1}}{g_{m2}}}{\left(1+j\,f/f_{in}\right)\left(1+j\,f/f_{out}\right)}$$

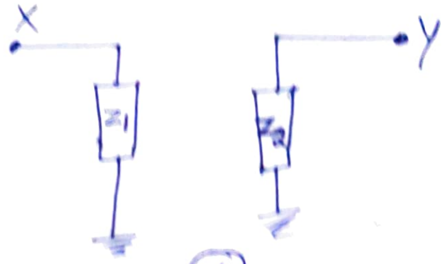
where the pole associated with the amplifier's input is located at,

$$f_{in} = \frac{1}{2\pi\,\tau_{in}}$$

and the pole associated with amplifier's output is located at,

$$f_{out} = \frac{1}{2\pi\,\tau_{out}}$$

ⓐ



ⓑ

Miller's theorem: In the ckt. of fig ⓐ can be converted to that of fig. ⓑ, then $Z_1 = Z/(1-A_V)$ & $Z_2 = Z/(1-1/A_V)$

where $A_V = V_y/V_x$

Proof: The current flowing through Z from X to Y is equal to $\frac{V_x - V_y}{Z}$. For the two circuits to be equivalent, the same current must flow through Z_1.

Thus, $\frac{V_x - V_y}{Z} = \frac{V_x}{Z_1} \Rightarrow Z_1 = \frac{Z\,V_x}{V_x - V_y}$

$$\text{or, } Z_1 = \frac{Z\,V_x}{V_x\left(1 - \frac{V_y}{V_x}\right)}$$

$$= \frac{Z}{1 - A_V} \quad \text{— (1)}$$

$$\text{IIIly, } Z_2 = \frac{Z}{1 - 1/A_V} \quad \text{— (2)}$$

If Z is a capacitor,

Then, $Z = \frac{1}{C_{gd}}$; $Z_1 = \frac{1}{C_{MI}}$; $Z_2 = \frac{1}{C_{MO}}$

$\therefore$ Eqn. (1), $\frac{1}{C_{MI}} = \frac{\frac{1}{C_{gd}}}{1 + A_V} \Rightarrow C_{MI} = C_{gd}(1 + A)$

of the amplifier is very dependant on the DC biasing at the i/p of the amplifier.) This is a common problem in CMOS analog IC design, that is, determining the exact voltage on the drains of two series connected p- and n-channel MOSFETs. Feedback is normally employed to set the o/p voltage of the amplifier at some known value. This results in a DC voltage on the i/p of the amplifier. A common feedback scheme used to bias the single stage amplifier is to AC couple the i/p signal to the gate of M1 and to take the o/p signal from the drain of M2/M1. A large resistor is then placed between the o/p & the i/p. When the basic amplifier is designed properly, this forces M1 and M2 into the saturation region.

The resistance looking into the drain of M2 is simply the o/p resistance given by $r_{o2} \approx 1/\lambda_2 I_D$. This is in parallel with the resistance looking into the drain of M1, which is also $r_{o1} \approx 1/\lambda_1 I_D$.

$$\therefore \boxed{A_v = -g_{m1}(r_{o1} \| r_{o2}) = \frac{-g_{m1}}{g_{o1} + g_{o2}}}$$

To see how the DC bias current affects the small-signal gain, we rewrite above expn. by substituting for g_{m1}, g_{o1} and g_{o2}:

$$A_v = \frac{-\sqrt{2\beta_1 I_D}}{I_D\lambda_1 + I_D\lambda_2} = -\sqrt{2\beta_1 I_D} \times \frac{1}{I_D(\lambda_1+\lambda_2)} = \frac{-\sqrt{2\beta_1}}{\sqrt{I_D}\,(\lambda_1+\lambda_2)}$$

This eqn. shows that the lower the bias drain current, the larger the gain. Since λ decreases with channel length, L, by increasing L we can increase the gain.

The gain of the amplifier can be increased by using a cascode current load in place of M2. The resistance looking into the drain of the cascode current source/load is much larger than the o/p resistance of the M1. Now the gain is referred as the open ckt. gain (intrinsic gain) of a common source amplifier. This is given by:

$$\boxed{A_v = -g_{m1} r_{o1} = -\sqrt{2\beta_1 I_D} \cdot \frac{1}{I_D\lambda_1}}$$

The frequency performance of the current source load amplifier is poorer than the diode connected amplifier.

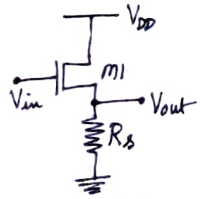
Example 22.5 : Boyce & Baker : page 503. [Estimate the B.W. of the amplifier shown above. $f = 1/\ldots$

2. Common Drain Amplifier (Source Follower):—

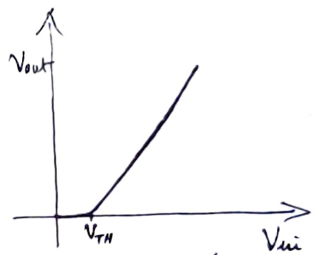
We have seen in CS amplifier that, to achieve a high voltage gain with limited supply voltage, the load impedance must be as large as possible. If such a stage is to drive a low-impedance load, then a "buffer" must be placed after the amplifier so as to drive the load with negligible loss of the signal level. The source follower can operate as a voltage buffer.

a) Source Follower with passive resistor load:—

As shown in fig ⓐ, the source follower senses the signal at the gate and drives the load at the source, allowing the source potential to "follow" the gate voltage.



a) Source follower



b) I/O characteristics.

Begining with the large-signal behavior, we note that for $V_{in} < V_{TH}$, $M1$ is off and $V_{out} = 0$. As V_{in} exceeds V_{TH}, $M1$ turns ON in saturation and I_{D1} flows through R_S. As V_{in} increases further, V_{out} follows the i/p with a difference (level shift) equal to V_{GS}.

We can express the i/p–o/p characteristic as:

$$V_{out} = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(V_{in}-V_{TH}-V_{out}\right)^2 R_s = V_{out}.$$

The small signal gain is obtained by differentiating both sides w.r.t. V_{in},

$$\frac{\partial V_{out}}{\partial V_{in}} = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\cdot 2\left(V_{in}-V_{TH}-V_{out}\right)\left(1-\frac{\partial V_{TH}}{\partial V_{in}}-\frac{\partial V_{out}}{\partial V_{in}}\right)R_s.$$

We know that, $\partial V_{TH}/\partial V_{in} = \eta\, \partial V_{out}/\partial V_{in}$.

$$\therefore \frac{\partial V_{out}}{\partial V_{in}} = \mu_n C_{ox}\frac{W}{L}\left(V_{in}-V_{TH}-V_{out}\right)\left(1-\eta\frac{\partial V_{out}}{\partial V_{in}}-\frac{\partial V_{out}}{\partial V_{in}}\right)R_s$$

or,
$$A_v = \frac{\mu_n C_{ox}\frac{W}{L}\left(V_{in}-V_{TH}-V_{out}\right)R_s}{1+\mu_n C_{ox}\frac{W}{L}\left(V_{in}-V_{TH}-V_{out}\right)R_s\left(1+\eta\right)}$$

But, $g_m = \mu_n C_{ox}\frac{W}{L}\left(V_{in}-V_{TH}-V_{out}\right)$

Consequently,
$$\boxed{A_v = \frac{g_m R_s}{1+\left(g_m+g_{mb}\right)R_s}}$$

The same result is more easily obtained with the aid of a small-signal equivalent ckt.

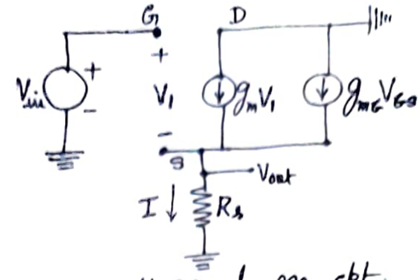


Fig © Small signal eq. ckt.

From fig. © we have,

$$V_{out} = V_{in} - V_1 \text{, or } V_1 = V_{in} - V_{out}.$$

$$V_{GS} = -V_{out}$$

$$\& \quad g_m V_1 + g_{mG} V_{GS} = V_{out}/R_s.$$

$$\text{or, } g_m V_1 - g_{mG} V_{out} = V_{out}/R_s.$$

OR, $v_{out} = I R_s$

$v_{out} = \left[g_m V_1 + g_{mG} V_{GS}\right] R_s.$

$=$

$$\text{or, } g_m (V_{in} - V_{out}) - g_{mG} V_{out} = V_{out}/R_s.$$

$$g_m V_{in} = V_{out}/R_s + g_m V_{out} + g_{mG} V_{out}.$$

$$\cancel{g_m V_{in} = \frac{V_{out}}{R_s}(1 + g_m + g_{mG}) R_s}$$

$$g_m V_{in} = \frac{V_{out}}{R_s}\left[1 + (g_m + g_{mG}) R_s\right]$$

$$A_v = \frac{V_{out}}{V_{in}} = \frac{g_m R_s}{1 + (g_m + g_{mG}) R_s}$$

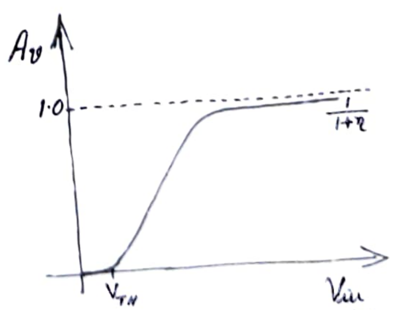


Fig ⓓ: Voltage gain of source follower versus input voltage.

c) Source-follower with diode-connected (active) load:-

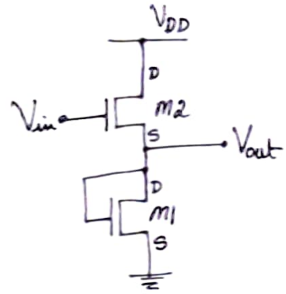


Fig (a): NMOS load

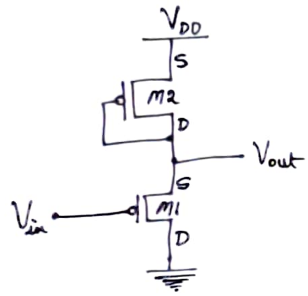
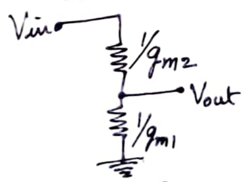


Fig (b): PMOS load.

Source-follower configurations using active loads are shown above. The M2 sources current, while M1 sinks current. A source-follower implemented in CMOS has an asymmetric drive capability; that is, the ability of the follower to source current is not equal to the ability to sink current for a given bias condition & AC input signal. Also in both configurations, the common-drain amplifier exhibits the body effect.

The small-signal gain of the NMOS source follower shown above (a) is simply determined by a voltage divider between the resistance looking into the source of M2 with the resistance of the gate-drain connected load, M1. The o/p voltage is given by,



$$V_{out} = V_{in} \cdot \frac{1/g_{m1}}{1/g_{m1} + 1/g_{m2}}$$

$$= V_{in} \cdot \frac{1}{g_{m1}} \cdot \frac{g_{m1} g_{m2}}{g_{m1}+g_{m2}} = \frac{g_{m2}}{g_{m1}+g_{m2}} \cdot V_{in}$$

or, $A_v = \frac{V_{out}}{V_{in}} = \frac{g_{m2}}{g_{m1}+g_{m2}}$ [g_{mb2} is neglected].

[Note:- For passive load: replace g_{m1} with $1/R_s$, we get the same gain as derived in the last section for passive load.]

or, $A_V = \frac{g_{m2}}{g_{m1}+g_{m2}} = \frac{g_{m2}}{g_{m2}(1+g_{m1}/g_{m2})} = \frac{1}{1+g_{m1}/g_{m2}}$

But $g_m = \sqrt{2\beta_i I_D}$,

$$\therefore A_v = \frac{1}{1+\sqrt{\frac{W_1/L_1}{W_2/L_2}}} \quad \text{——} \quad (A)$$

The o/p resistance, $R_{out} = \frac{1}{g_{m1}} \parallel \frac{1}{g_{m2}} = \underline{\underline{\frac{1}{g_{m1}+g_{m2}}}}$

The above eqn. Ⓐ reveals that the gain of source-follower is always less than one.

Note: G_m is the transconductance when output is shorted to ground.

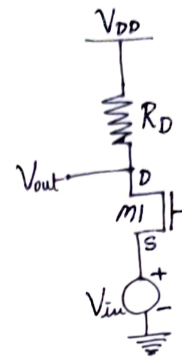
$$G_m = \frac{I_{out}}{V_{in}}$$

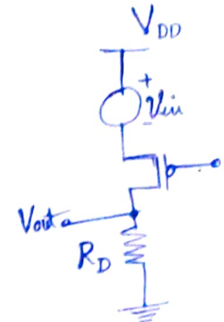
R_{out} is the output impedance when input is set to zero.

3. Common – Gate Amplifier :–

In common – source amplifiers and source – followers, the input signal is applied to the gate of a MOSFET. It is also possible to apply the signal to the source terminal.

a) With passive load :–





As shown here, a common-gate (CG) amplifier senses the i/p at the source terminal and produces the o/p at the drain. The gate is connected to a dc voltage to establish proper operating conditions. Note that the bias current of M1 flows through the input signal source.

We first study the large-signal behavior of the above ckt. For simplicity, let us assume that V_{in} decreases from a large positive value. For $V_{in} \geq V_G - V_{TH}$, M1 is off and $V_{out} = V_{DD}$. For lower values of V_{in}, we can write,

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_G - V_{in} - V_{TH})^2, \text{ if M1 is in saturation.}$$

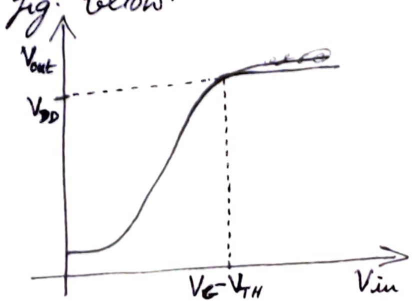
As V_{in} decreases, so does V_{out}, eventually driving M1 into the triode region if,

$$V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_G - V_{in} - V_{TH})^2 R_D = V_G - V_{TH}$$

If M1 is saturated, we can express that o/p voltage as, [neglecting λ (i.e, $\lambda = 0$)]

$$V_{out} = V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_G - V_{in} - V_{TH})^2 R_D$$

The input-o/p characteristics is as shown in fig. below.



The small-signal gain is obtained by differentiating the above eqn. w.r.t. V_{in}, we get,

$$\frac{\partial V_{out}}{\partial V_{in}} = -\mu_n C_{ox}\frac{W}{L}(V_G - V_{in} - V_{TH})\left(-1 - \frac{\partial V_{TH}}{\partial V_{in}}\right)R_D.$$

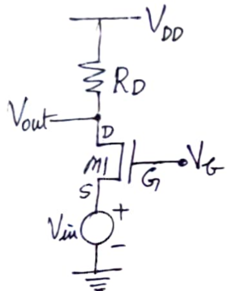
Since $\frac{\partial V_{TH}}{\partial V_{in}} = \frac{\partial V_{TH}}{\partial V_{SB}} = \eta$, we have,

$$\frac{\partial V_{out}}{\partial V_{in}} = \mu_n C_{ox} \frac{W}{L} (V_G - V_{in} - V_{TH})(1+\eta) R_D .$$

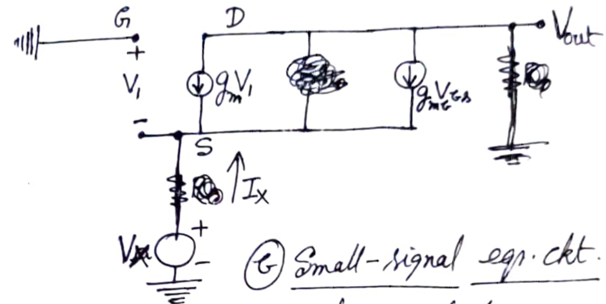
or, $$\boxed{A_v = g_m (1+\eta) R_D}$$ Here we assumed that impedance of the signal source, $R_s = 0$; and o/p impedance $r_o = \infty$.

Note that the gain is positive. Interestingly, body effect increases the equivalent transconductance of the stage.

The i/p impedance of the ckt. seen at the source of $M1$ is (for $\lambda = 0$)



$\Longrightarrow$



(b) Small-signal eqv. ckt. to find input impedance:

Here $V_1 = -V_X$

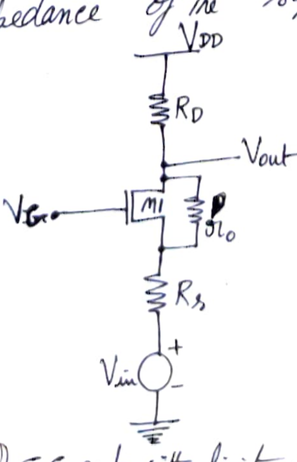
$$I_X - g_m V_X - g_{mb} V_X = 0.$$

$$I_X = V_X (g_m + g_{mb})$$

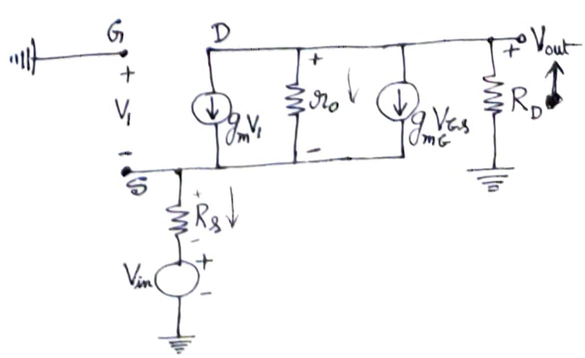
or, $$R_{in} = \frac{V_X}{I_X} = \frac{1}{g_m + g_{mb}}$$

Thus, the body effect decreases the i/p impedance of the CG amplifier. The relatively low input impedance of the CG amp. proves useful in some applications.

Now let us study the common-gate topology in a more general case, taking into account both the o/p impedance, r_o of the transistor and the impedance of the signal source. This is depicted in the fig. below:



(a) CG amp. with finite r_o.



(b) Small signal eqv. ckt.

Noting that, the current flowing through R_s is equal to $-\frac{V_{out}}{R_D}$, we have,

$$V_1 - \frac{V_{out}}{R_D} R_s + V_{in} = 0; \quad V_{GS} = V_1;$$

$$\text{or, } V_1 = \frac{V_{out}}{R_D} R_s - V_{in} \quad \text{——} \; (1)$$

Moreover, since the current through r_o is, ~~equal to~~

$$I_{r_o} = \frac{-V_{out}}{R_D} - g_m V_1 - g_{mb} V_1$$

$$\text{or, } \left(-\frac{V_{out}}{R_D} - g_m V_1 - g_{mb} V_1\right) r_o = \frac{V_{out}}{R_D} R_s + V_{in} = V_{out}.$$

Substituting value of V_1 from eqn. (1) we get,

$$\left[-\frac{V_{out}}{R_D} - (g_m + g_{mb})\left(\frac{V_{out}}{R_D} R_s - V_{in}\right)\right] r_o - \frac{V_{out}}{R_D} R_s + V_{in} = V_{out}.$$

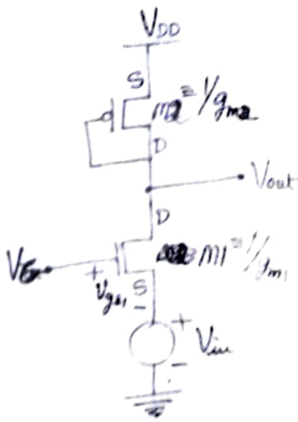
$$\text{or, } \frac{V_{out}}{V_{in}} = A_v = \frac{(g_m + g_{mb}) r_o + 1}{r_o + (g_m + g_{mb}) r_o R_s + R_s + R_D} \cdot R_D$$

The gain of the CG amplifier is slightly higher due to body effect.

In this expn. if we set $R_s = 0$ and $r_o = \infty$ we get,

$A_v = g_m (1+\eta) R_D$, which is same as we have derived earlier.

(b) With diode-connected (active) load:-



The gain of this amplifier (neglecting body effect) is,

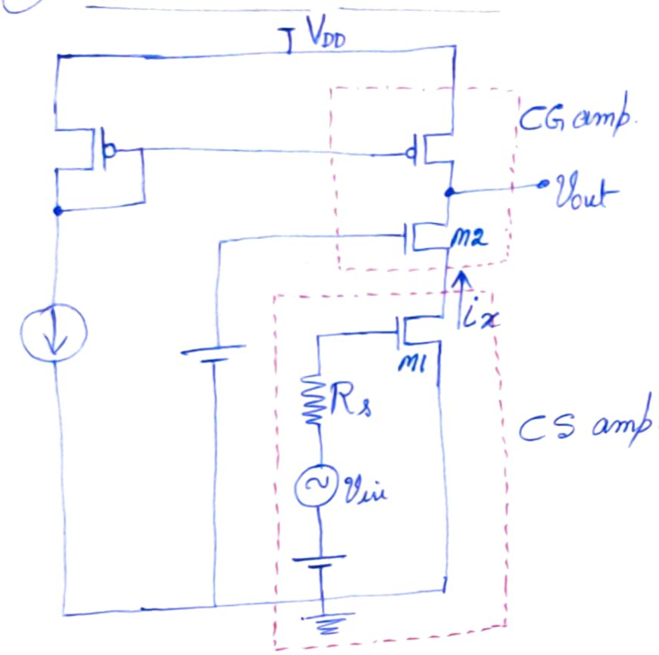
$$A_v = \frac{V_{out}}{V_{in}} = \frac{-i_d \cdot 1/g_{m2}}{-v_{gs1}} = \frac{-i_d \cdot 1/g_{m2}}{-i_d \cdot 1/g_{m1}} = \frac{1/g_{m2}}{1/g_{m1}}$$

$$\text{i.e, } A_v = \frac{g_{m1}}{g_{m2}} = \sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}}$$

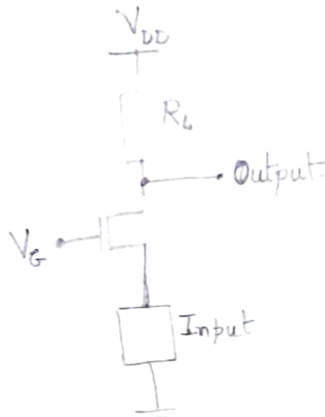
Which is the same form as the gain for the common-source amplifier with active load.

Cascode Amplifiers:-

(a) CS-CG Cascode:



$$A_v = \frac{v_{out}}{v_{in}} = \left(\frac{v_{out}}{i_x}\right) \cdot \left(\frac{i_x}{v_{in}}\right)$$



Advantages: (1) Increased gain.

(2) To achieve high output resistance.

CG amp. is used as a transimpedance amp in combinati