Chapter 3

Single-Stage Amplifiers

中科大

微电子学院

教材：拉扎维《模拟CMOS集成电路设计》

contents

- Four types of amplifiers（MOSFET，小信号输入输出公共端）：
- Common-source topology
- Common-gate topology
- Source followers
- Cascode(共源共栅) configuration

- Large-signal characteristics
- Small-signal characteristics (Low frequency behavior)

3.1 basic concepts

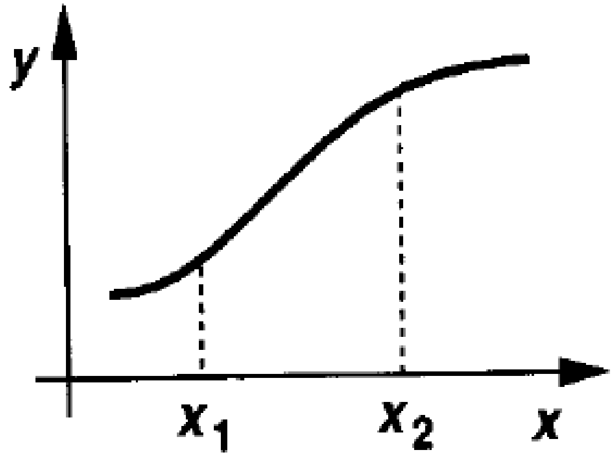
- The input-output characteristic of an amplifier:

$$y(t) = a_0 + a_1 x(t) + a_2 x^2(t) + \cdots + a_n x^n(t), \qquad x_1 \le x \le x_2$$

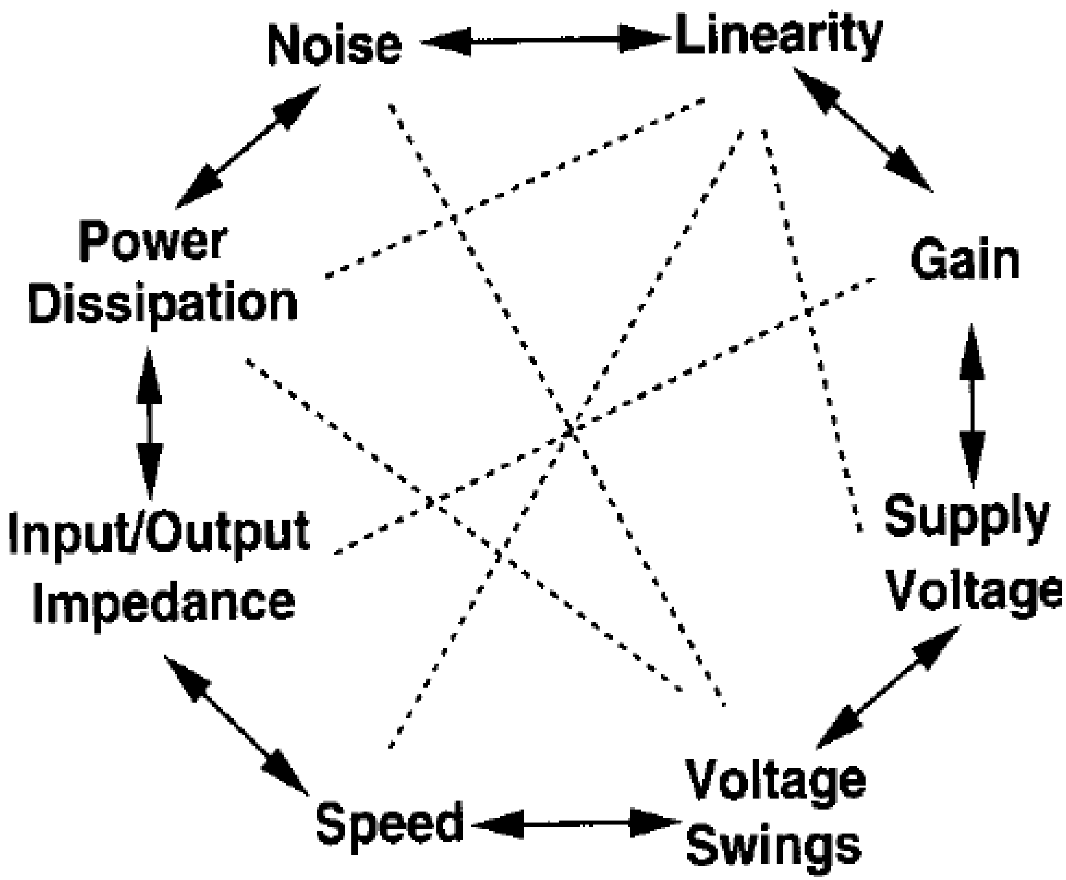
- So long as x(t) is small enough, reasonable approximation

$$y(t) \approx a_0 + a_1 x(t)$$

- when $a_1 x(t) << a_0$ the bias point （偏置点a0） is disturbed negligibly.(受到的扰动忽略不计)。



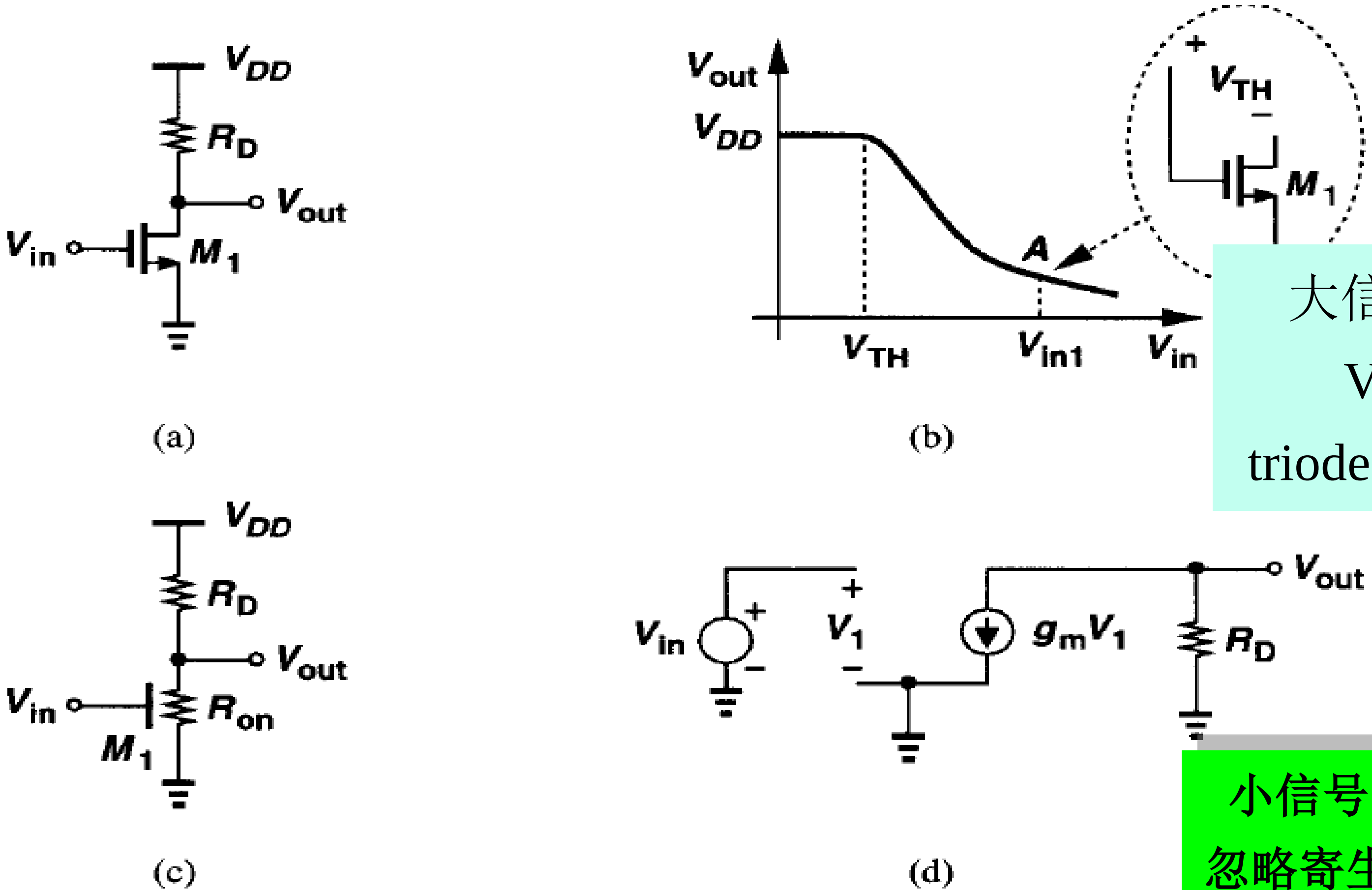
Analog circuits design octagon



trade-off (compromise)

3.2 Common-source (CS) Stage

- 3.2.1 Common-source stage with resistive load



大信号（包括直流）

$V_{in} - V_{TH} > V_{out}$

triode region： $V_{in} > V_{in1}$

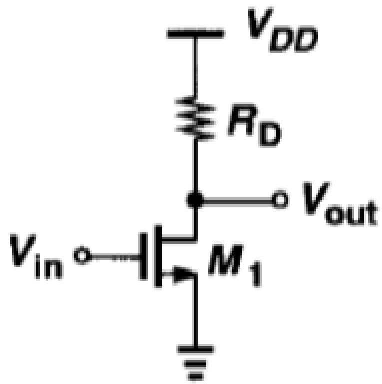
小信号：交变，无直流

忽略寄生电容，仅限低频

Figure 3.3 (a) Common-source stage, (b) input-output characteristic, (c) equivalent circuit in deep triode region, (d) small-signal model for the saturation region.

1.采用大信号分析方法：IV公式推导

- By virtue of(借助) trans-conductance，MOSFET convert variations in its gate-source voltage to a small drain current.



In saturation region : $V_{out} > V_{in} - V_{Tn}$

$$V_{out} = V_{DD} - R_D I_{Dn} = V_{DD} - R_D \frac{1}{2}\mu_n C_{OX}\frac{W}{L'}(V_{GS} - V_{Tn})^2$$

手算时忽略沟道长度调制效应，L’=Leff=Ldrawn-2LD

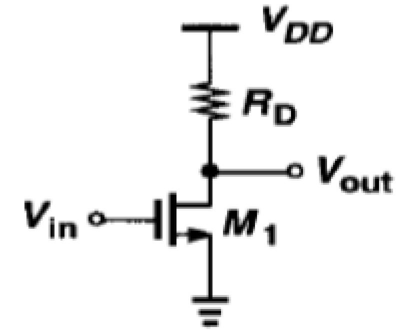
（LD=漏/源与栅极的交叠长度）.

Note that the input impedance of the circuit is very high at low frequencies .

仿真时RD可先用analogLib库中res，然后用流片工艺库中电阻代替。

大信号分析方法(续)

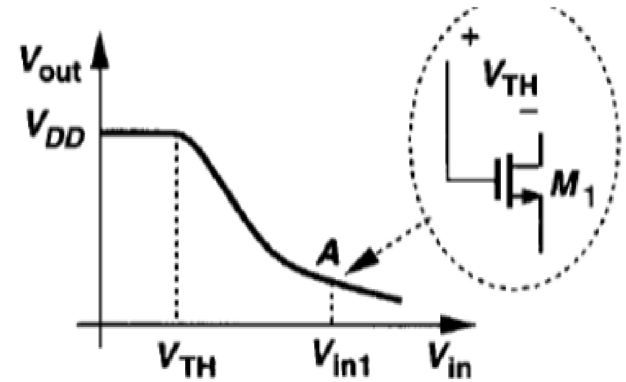
A点(饱和区最小V_{DS})V_{in1}，输入信号的上限



$$V_{out} = V_{in1} - V_{Tn} = V_{DD} - R_D \frac{1}{2} \mu_n C_{OX} \frac{W}{L'} (V_{in1} - V_{Tn})^2$$

$$= V_{DD} - R_D \frac{1}{2} \mu_n C_{OX} \frac{W}{L_{eff}} (V_{in1} - V_{Tn})^2 (1 + \lambda V_{DS})$$

解出V_{in1}。手算时$\lambda V_{DS} \approx 0$



In the triode region

Vin > Vin1, 即Vin - V_{TH} >Vout

$$V_{out} = V_{DD} - R_D \times \mu_n C_{OX} \frac{W}{L_{eff}} \left[(V_{in1} - V_{Tn}) V_{DS} - \frac{1}{2} V_{DS}^2 \right]$$

$$= V_{DD} - R_D \times \frac{1}{2} \mu_n C_{OX} \frac{W}{L_{eff}} \left[2(V_{in1} - V_{Tn}) V_{out} - V_{out}^2 \right]$$

三极管区仍能工作，
但跨导小，且与V_{DS}有关

得到small-signal gain of common source

$$V_{out} = V_{DD} - R_D \frac{1}{2} \mu_n C_{ox} \frac{W}{L_{eff}} (V_{in} - V_{Tn})^2 (1 + \lambda V_{DS}))$$

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} = -R_D \times \mu_n C_{ox} \frac{W}{L_{eff}} (V_{in} - V_{Tn})(1 + \lambda V_{DS}) = -g_m R_D$$

式中 $g_m = \dfrac{\partial I_D}{\partial V_{GS}} = \mu_n C_{ox} (W / L_{eff}) (V_{GS} - V_{Tn})(1 + \lambda V_{DS})$

The dependence of the gain upon the signal level leads to nonlineari ty

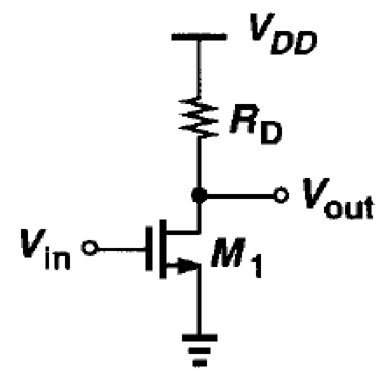
基本概念：（1）放大倍数（增益）仅是对交变信号，不包括直流部分（工作点）；

（2）CS负增益表示输入与输出中的交变信号反向；

（3）负载R_D可泛指一切有源器件（含受控电压或电流源的器件，如MOS、三极管）和无源器件(RLC）。

Example 3.1

Sketch the drain current and transconductance as a function of the input voltage.



Solution:

饱和区 $I_D = \frac{1}{2}\mu_n C_{OX}\frac{W}{L_{\text{eff}}}(V_{\text{in}} - V_{Tn})^2(1+\lambda V_{out})$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \mu_n C_{ox}(W/L_{eff})(V_{in} - V_{Tn})(1+\lambda V_{out})$$

还须满足RD上的VI关系

Vin1=Vout+Vtn

三极管（线性电阻）区： $V_{out} < V_{\text{in}} - V_{Tn}$

$$I_D = \mu_n C_{OX}\frac{W}{L_{\text{eff}}}[(V_{in} - V_{Tn})V_{out} - \frac{1}{2}V_{out}^2]$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \frac{\partial I_D}{\partial V_{in}} = \mu_n C_{ox}\frac{W}{L_{\text{eff}}}V_{out}$$

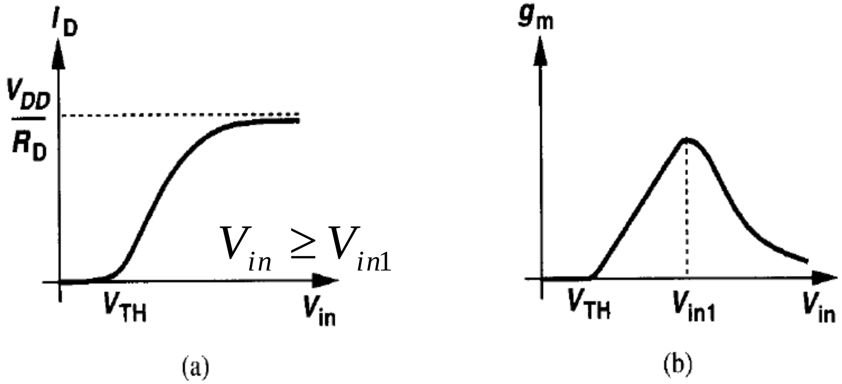


Figure 3.4

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} = -g_m R_D$$

沟道长度调制对增益的影响

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

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

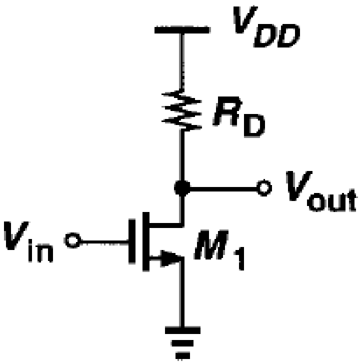
$$I_D \approx (1/2) \mu_n C_{ox} (W/L)(V_{in} - V_{TH})^2$$

$$A_v = -R_D g_m - R_D I_D \lambda A_v \qquad A_v = -\frac{g_m R_D}{1 + R_D \lambda I_D}.$$

Since $\lambda I_D = 1/r_O$ $\quad A_V = -\frac{g_m R_D}{1 + R_D \lambda I_D} = -g_m \frac{r_o R_D}{r_o + R_D} = -g_m (r_o \,||\, R_D)$

沟道长度调制效应(ro=rds)使得gain减小

2.采用交变小信号模型计算方法



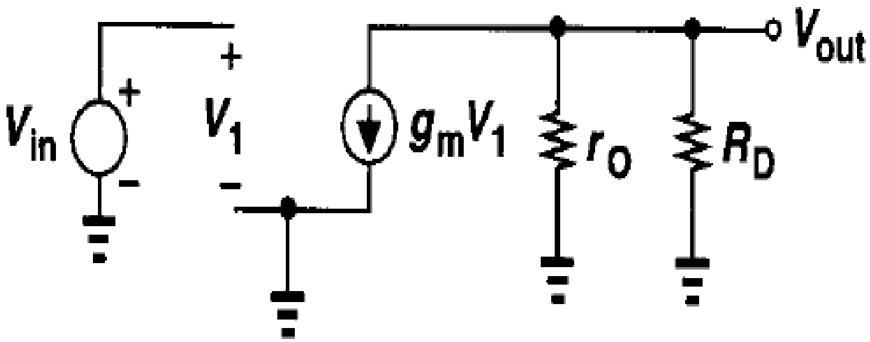


Figure 3.5 Small-signal model of CS stage including the transistor output resistance.

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

$$V_1 = V_{in}$$

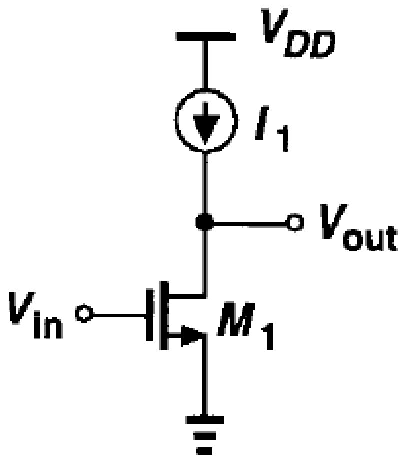
$$V_{out}/V_{in} = -g_m (r_O \| R_D)$$

ro代表了沟道长度调制效应。如负载RD较小(小于几K欧)，可忽略ro.

采用小信号模型，计算结果与大信号方法相同(Vin和Vout是交变小信号）。

信号流分析可省略小信号模型。直观方法：M1通过跨导gm将Vin转换成漏极受控电流源变化，该电流变化通过总输出阻抗转换成输出电压变化Vout

本征增益$g_m r_o$



理想电流源具有无穷大输出阻抗。

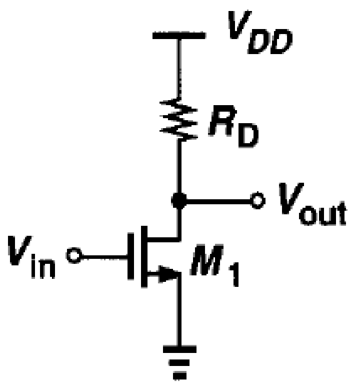
$$I_{D1} = \frac{1}{2}\mu_n C_{ox}(V_{in} - V_{TH})^2(1 + \lambda V_{out})$$
$$= I_1,$$

电压增益=$-g_m r_o$

"intrinsic gain" represent the maximum voltage gain that can be achieved using a single device.

应$g_m r_o$ >>1， thus usually $1/g_m << r_o$

例：



已知：$C_{OX}=10\times10^{-7}\,F/cm^2$，$\mu_n=300\,cm^2/(V\cdot s)$，$V_{Tn}=0.5V$

$V_{DD}=1.8V$，$V_{in,\ DC}=0.7V$，设$\lambda_n=0.1/V$，$I_D=100\mu A, V_{out}=1V$，

$0.5\mu m$工艺漏/源与栅极交叠$LD\approx0.1\mu m$，计算$\frac{W}{L_{eff}}$，R_D，$g_m r_o$，A_v。

解：$I_D=\frac{1}{2}\mu_n C_{OX}\frac{W}{L_{\text{eff}}}(V_{\text{in}}-V_{Tn})^2(1+\lambda V_{out})$，已知过驱动电压和电流可用此式得$\frac{W}{L_{\text{eff}}}$

$$100\times10^{-6}A=\frac{1}{2}\times300\,cm^2/(V\cdot s)\times10\times10^{-7}\,F/cm^2\times\frac{W}{L_{eff}}\times0.2^2V^2\times1.1$$

得到：$\frac{\text{W}}{\text{L}_{\text{eff}}}=\frac{100}{6.6}\approx\frac{4.5}{0.3}$，取$L_{drawn}=0.5\mu m$

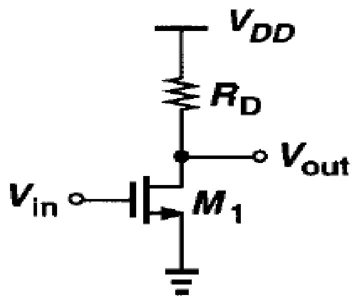
$$R_D=\frac{V_{\text{DD}}-V_{out}}{I_{\text{D}}}=\frac{1.8V-1V}{100\times10^{-6}A}=8k\Omega，\quad r_o=\frac{1}{\lambda I_{\text{D}}}=\frac{1}{0.1/V\times100\times10^{-6}A}=100\,k\Omega$$

$$g_m=\mu_n C_{ox}(W/L_{eff})(V_{in}-V_{Tn})(1+\lambda V_{out})=300\times10\times10^{-7}\times\frac{4.5}{0.3}\times0.2\times1.1\approx1\frac{mA}{V}$$

$$g_m r_o=100，\quad A_v=-g_m(R_D\,||\,r_o)=-10^{-3}\times(8\,||\,100)\times10^3=-7.4$$

饱和区$V_{in,min}=V_{Tn}$, $V_{out,max}=V_{DD}$

$V_{in,max}=V_{out,min}+V_{Tn}$,



$$V_{out,\min} = V_{DD} - R_D \frac{1}{2} \mu_n C_{OX} \frac{W}{L_{eff}} (V_{in,\max} - V_{Tn})^2 (1 + \lambda V_{out,\min})$$

$$\approx V_{DD} - R_D \frac{1}{2} \mu_n C_{OX} \frac{W}{L_{eff}} V_{out,\min}{}^2, \quad (\because \lambda V_{out,\min} \approx 0, \quad V_{out} \geq V_{in,\max} - V_{Tn})$$

$$= 1.8V - 8 \times 10^3 \Omega \times \frac{1}{2} \times 300\ cm^2 / (V \cdot s) \times 10 \times 10^{-7}\ F/cm^2 \times \frac{4.5}{0.3} \times V_{out,\min}{}^2$$

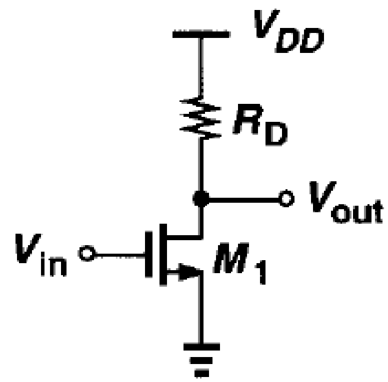
$$= 1.8V - 18V_{out,\min}{}^2 \times \Omega \bullet V \bullet F/s = 1.8V - 18V_{out,\min}{}^2$$

$$(\because Q = CV = It, \text{故量纲} FV = As, \therefore \Omega \bullet V \bullet F/s = \Omega A = V)$$

得到：$$V_{out,\min} = \frac{-1 + \sqrt{1^2 - 4 \times 18 \times (-1.8)}}{2 * 18} = \frac{10.43}{36} \approx 0.3V$$

$$V_{in,\max} \approx V_{out,\min} + V_{Tn} \approx 0.8V$$

例：改变直流电流进行比较



已知：$C_{OX}=10\times10^{-7}\,F/cm^2$，$\mu_n=300\,cm^2/(V\cdot s)$，$V_{Tn}=0.5V$

$V_{DD}=1.8V$，$V_{in}=0.7V$，设$\lambda_n=0.1/V$，改变$I_D=20\mu A, V_{out}=1V$，

采用$0.5\mu m$工艺$LD\approx0.1\mu m$，重新计算$\dfrac{W}{L_{eff}}, R_D, g_m r_o$，$A_v$。

解：$I_D=\dfrac{1}{2}\mu_n C_{OX}\dfrac{W}{L_{\mathrm{eff}}}(V_{\mathrm{in}}-V_{Tn})^2(1+\lambda V_{out})$

$$20\times10^{-6}A=\frac{1}{2}\times300\,cm^2/(V\cdot s)\times10\times10^{-7}\,F/cm^2\times\frac{W}{L_{eff}}\times0.2^2V^2\times1.1$$

得到：$\dfrac{\mathrm{W}}{\mathrm{L_{eff}}}=\dfrac{20}{6.6}\approx\dfrac{0.9}{0.3}$，取$L_{drawn}=0.5\mu m$

$$R_D=\frac{V_{\mathrm{DD}}-V_{out}}{I_{\mathrm{D}}}=\frac{1.8V-1V}{20\times10^{-6}A}=40k\Omega，\quad r_o=\frac{1}{\lambda I_{\mathrm{D}}}=\frac{1}{0.1/V\times20\times10^{-6}A}=500k\Omega$$

$$g_m=\mu_n C_{ox}(W/L_{eff})(V_{in}-V_{Tn})(1+\lambda V_{out})=300\times10\times10^{-7}\times\frac{0.9}{0.3}\times0.2\times1.1\approx0.2\frac{mA}{V}$$

$$g_m r_o=100，\quad A_v=-g_m(R_D\,||\,r_o)=-0.2\times10^{-3}\times(40\,||\,500)\times10^3=-7.4$$

该例如何提高增益？

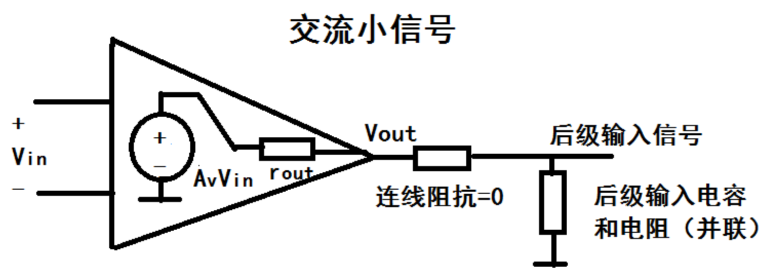
单级CS放大器：实际电阻R_D做为负载

- 增益不太。原因：R_D受到直流电流和工作点电压限制，不能很大，如在高频（需大电流）和大动态范围（直流输出电平为1/2电源）时。
- 大交变信号时，应注意V_{DS}避免进入线性区。
- 实际设计应注意考虑后级电路，避免影响工作点。
- 增益等参数仿真时须有后级电路的交流等效输入阻抗做为本级load。

负载C_L=后级电路的输入电容（高频时与后级增益和器件尺寸有关）+长连线的寄生电容。

放大电路建模：受控电压源AvVin或电流源GmVin，交流小信号！

问题：**rout=RD||ro**大或小比较好？



实际电阻（例smic18mmrf工艺）

(1) POLY电阻（栅多晶硅）：

1）高阻：1000欧/方块，例rhrpo, rpposab,…

2）普通：几欧/方块, rppo,rnpo,

(2) 扩散区电阻(N+,P+,NW)：几百欧/方块电阻

例：rpdif，rndif, rnwaa, rndifsab,…

(3) 金属(M1~MTOP)：几毫欧~几十毫欧/方块

设计时检查电流密度：金属层（顶层除外）：<0.5~1mA/微米宽

硅（POLY,扩散区）：<0.3~0.5mA/微米宽

电阻宽度有最低限制，如0.4~2um，大电阻面积大，有寄生电容

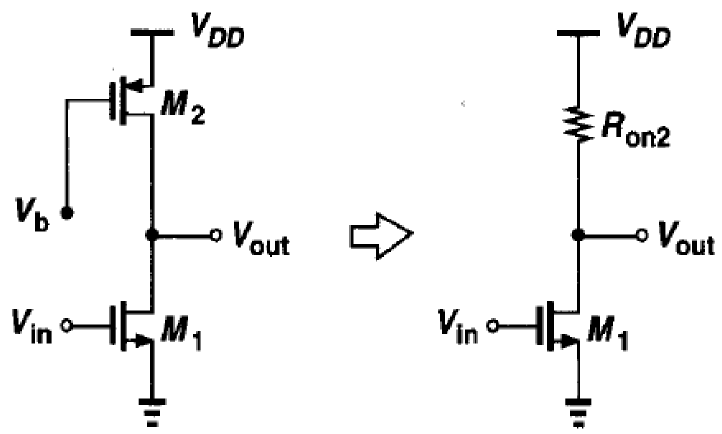
电阻值不太准确，误差达20%,且温度系数大。
设计电路时，最好是采用2个电阻成比例计算的结构，
单个电阻应用在电路时应在系统设计上采取鲁棒性设计。

3.2.2 CS with Triode MOSFET Load

- 深线性区工作的MOS器件，伏安特性为电阻。

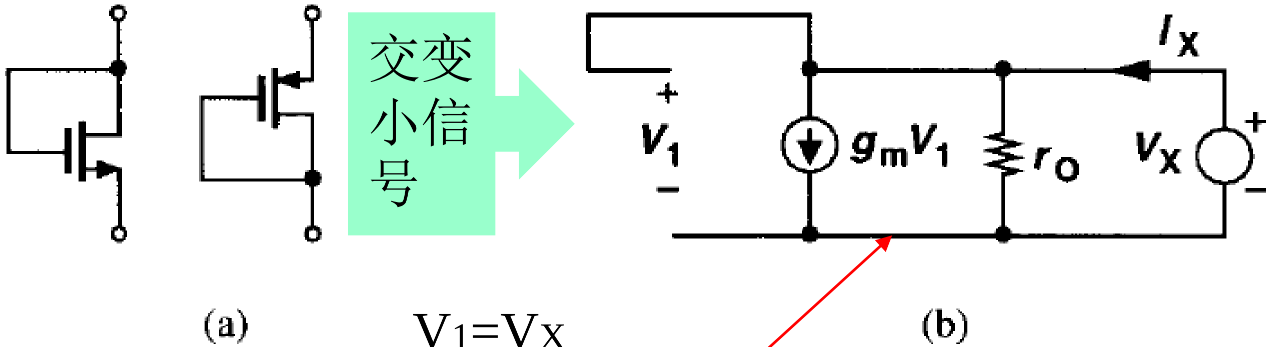
$$R_{on2} = \frac{1}{\mu_p C_{ox}\left(\frac{W}{L}\right)_2 (V_{DD} - V_b - |V_{THP}|)}$$

- 缺点：VDD-Vb噪声较大。



3.2.3 CS Stage with diode-Connected Load

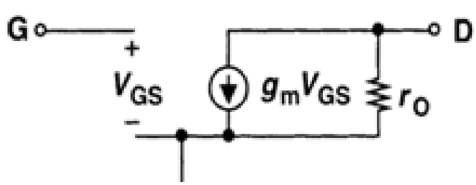
MOS 管连接成二极管做负载电阻，总是工作在饱和区，直流与交流电阻不同。先不考虑体效应（gmb)。设直流工作点由外电路确定。



输出电阻（作为信号来源电路的负载）

增益和输出电阻均是指交变小信号

Figure 3.7 (a) Diode-connected NMOS and PMOS devices, (b) small-signal equivalent circuit.

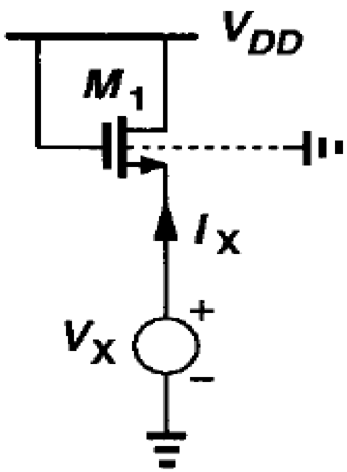


交变信号 $I_X = \dfrac{V_X}{r_0} + g_m V_X \qquad g_m r_o >> 1$

二极管作为电阻：$r_{out} = \dfrac{V_X}{I_X} = \dfrac{1}{1/r_o + g_m} = r_o \,||\, \dfrac{1}{g_m} \approx \dfrac{1}{g_m}$

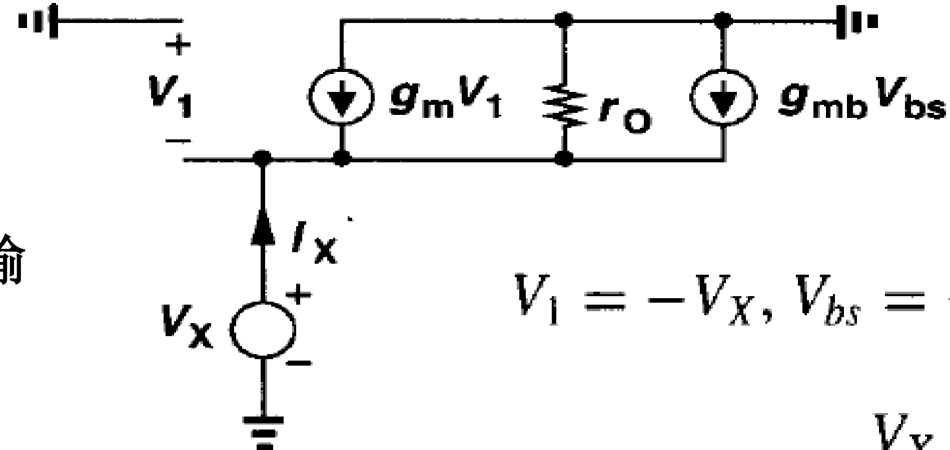
二极管负载等效于电阻1/gm，多为~几百欧。S/D无方向性

输出阻抗：If Body effect exists



(a)

输出阻抗与输入信号无关



$$V_1 = -V_X,\ V_{bs} = -V_X$$

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

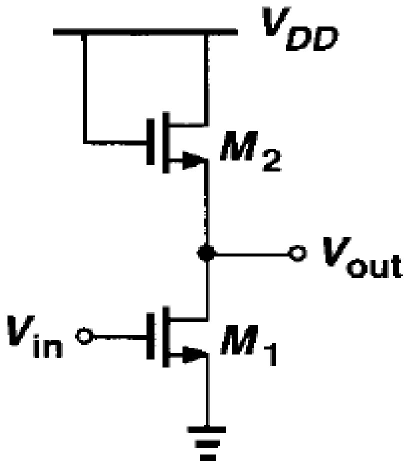
Figure 3.8 (a) Arrangement for measuring the equivalent resistance of a diode-connected MOSFET, (b) small-signal equivalent circuit.

二极管输出电阻：

$$r_{out} = \frac{V_X}{I_X} = \frac{1}{g_m + g_{mb} + 1/r_o} = r_o \,||\, \frac{1}{g_m + g_{mb}} \approx \frac{1}{g_m + g_{mb}}$$

体效应使rout减小。如何设计使得电路无体效应？

Study CS Stage with diode-Connected Load



$$A_v = -g_m R_D$$ 若可忽略沟道长度调制效应（ **L1不太小**）

$$A_v = -g_{m1}\frac{1}{g_{m2}+g_{mb2}}$$

$$= -\frac{g_{m1}}{g_{m2}}\frac{1}{1+\eta},$$ **式(3.26)**

由式2. 42:

$$\eta = \frac{\partial V_{TH}}{\partial V_{SB}} = \frac{\gamma}{2} \bullet \frac{1}{\sqrt{2\Phi_F + V_{SB}}}$$

$$g_m = \sqrt{2\mu_n C_{ox} W/L \, I_D(1+\lambda V_{DS})}$$

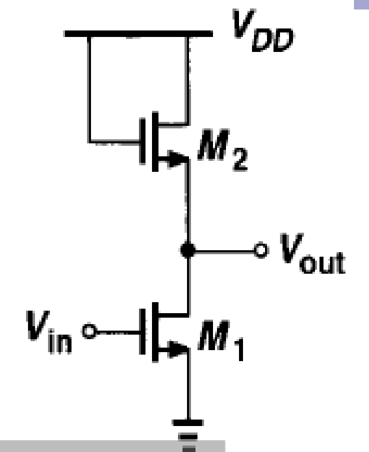
$$\longrightarrow A_v = -\frac{\sqrt{2\mu_n C_{ox}(W/L)_1 I_{D1}}}{\sqrt{2\mu_n C_{ox}(W/L)_2 I_{D2}}}\frac{1}{1+\eta},$$

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

$$A_v = -\sqrt{\frac{(W/L)_1}{(W/L)_2}}\frac{1}{1+\eta} \qquad (3.28)$$

So long as M1 stays in saturation ,the gain is independent of the bias currents and voltages .The input-output characteristic is linear.

Large-signal analysis



同样可以由大信号V/A分析，证明其具有线性特点.

Neglecting channel-length modulation for simplicity.

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

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

Vin与Vout成线性特性

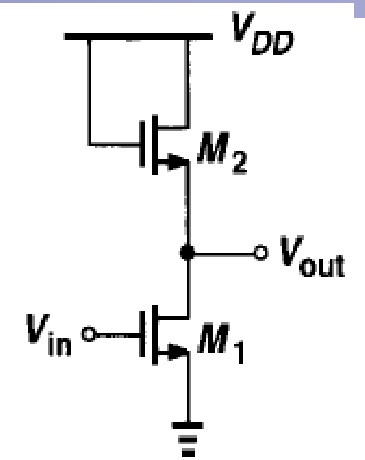
对 V_{in} 求偏导：$\sqrt{\left(\frac{W}{L}\right)_1} = \sqrt{\left(\frac{W}{L}\right)_2}\left(-\frac{\partial V_{out}}{\partial V_{in}} - \frac{\partial V_{TH2}}{\partial V_{\mathrm{in}}}\right)$

$$\because \frac{\partial V_{TH2}}{\partial V_{in}} = \frac{\partial V_{TH2}}{\partial V_{out}}\frac{\partial V_{out}}{\partial V_{in}} = \frac{\partial V_{TH2}}{\partial V_{SB2}}\frac{\partial V_{out}}{\partial V_{in}} = \eta_2\frac{\partial V_{out}}{\partial V_{in}} \qquad \therefore \sqrt{\left(\frac{W}{L}\right)_1} = -\sqrt{\left(\frac{W}{L}\right)_2}(1+\eta_2)\frac{\partial V_{out}}{\partial V_{in}}$$

$$A_v = -\sqrt{\frac{\left(W/L\right)_1}{\left(W/L\right)_2}} \times \frac{1}{1+\eta_2}$$

，与小信号分析相同。$\sqrt{\ }$ 是缺点！ $\eta_2 = \frac{g_{mb2}}{g_{m2}}$

直流工作点



$M2$始终在饱和区$I_{D2} = \frac{1}{2}\mu_n C_{OX}\frac{W_2}{L_2}(V_{GS2} - V_{Tn2})^2 = I_{D1}$

(1) $V_{in} < V_{TH1}$，则$I_{D1} = 0 = \frac{1}{2}\mu_n C_{OX}\frac{W_2}{L_2}(V_{GS2} - V_{TH2})^2$,

得$V_{GS2} = V_{TH2}$, $\therefore V_{out} = V_{DD} - V_{TH2} = V_{DD} - V_{TH2}$

(2) $V_{in} > V_{TH1}$ 时：$\frac{1}{2}\mu_n C_{OX}\frac{W_1}{L_1}(V_{in} - V_{TH1})^2 = \frac{1}{2}\mu_n C_{OX}\frac{W_2}{L_2}(V_{DD} - V_{out} - V_{TH2})^2$

$$V_{out} = V_{DD} - V_{TH2} + \sqrt{\frac{(W/L)_1}{(W/L)_2}}V_{TH1} - \sqrt{\frac{(W/L)_1}{(W/L)_2}}V_{in}$$

(3) Vin > Vout + VTH1 (beyond point A), M1 enters the triode region, the characteristic becomes nonlinear.

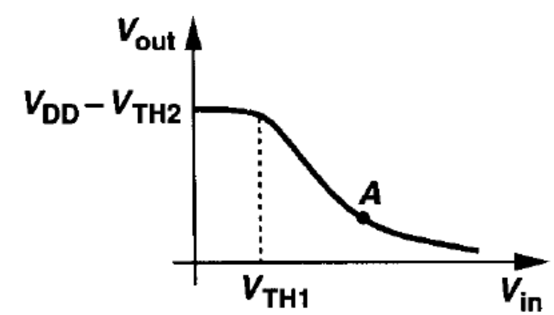


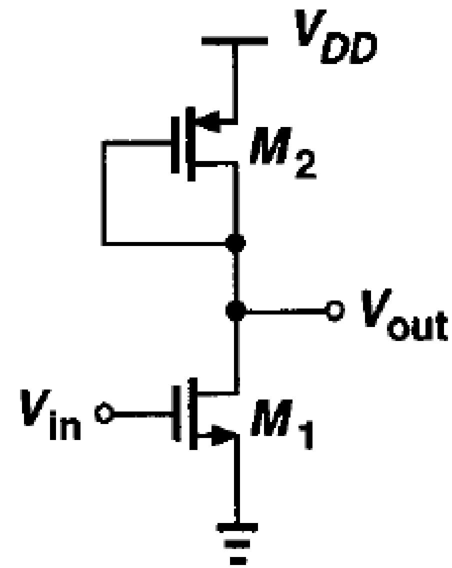
Figure 3.11 Input-output characteristic of a CS stage with diode-connected load.

PMOS as diode-connected load

This circuit is free from body effect!

$$\mu_n\left(\frac{W}{L}\right)_1(V_{GS1}-V_{TH1})^2\approx\mu_p\left(\frac{W}{L}\right)_2(V_{GS2}-V_{TH2})^2$$

channel-length modulation is neglected



$$A_v=\frac{\partial V_{out}}{\partial V_{in}}=-\frac{\partial V_{GS2}}{\partial V_{in}}=-\sqrt{\frac{\mu_n\left(W/L\right)_1}{\mu_p\left(W/L\right)_2}}$$

Fig. 3.12 CS stage with diode connected PMOS device

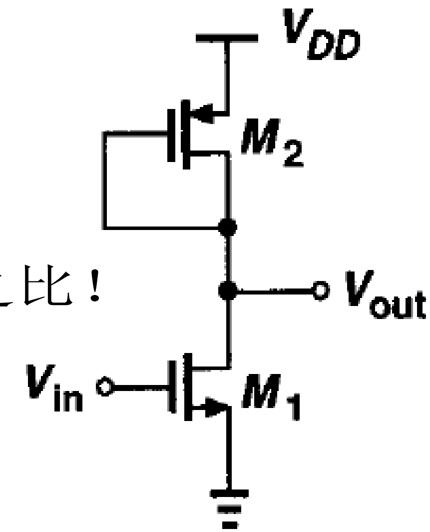
增益比NMOS二极管大！

结论：高增益需要大的输入MOS和小的负载MOS。

二极管负载的共源级放大线性度较好（大信号）.

Fault of CS with diode-connected load

$$\mu_n\left(\frac{W}{L}\right)_1(V_{GS1}-V_{TH1})^2\approx\mu_p\left(\frac{W}{L}\right)_2(V_{GS2}-V_{TH2})^2$$



$$\because\ A_v=-\sqrt{\frac{\mu_n\left(W/L\right)_1}{\mu_p\left(W/L\right)_2}}=-\frac{|V_{GS2}-V_{TH2}|}{V_{GS1}-V_{TH1}}$$，过驱动电压之比！式(3.35)

$$\therefore\ \mu_n\left(W/L\right)_1=A_v^2\mu_p\left(W/L\right)_2,\ and$$

$$|V_{GS2}-V_{TH2}|=-(V_{GS2}-V_{TH2})=-(V_{out}-V_{DD}-V_{TH2})$$

$$=-V_{out}+V_{DD}+V_{TH2}=-A_v(V_{GS1}-V_{TH1})$$

得 $V_{out}=V_{DD}+V_{TH2}+A_v(V_{GS1}-V_{TH1})$

例：$A_v=-10, V_{DD}=3V, V_{GS1}-V_{TH1}=0.2V$， $V_{TH2}=-0.7V$。求输出？

解：$V_{out}=3-0.7-10\times0.2=0.3V$，即$V_{GS2}=2.7V$很大，不好！

$V_{DS1}=0.3V>V_{GS1}-V_{TH1}=0.2V$，虽在饱和区，但输出允许摆幅很小

How to explain the paradox

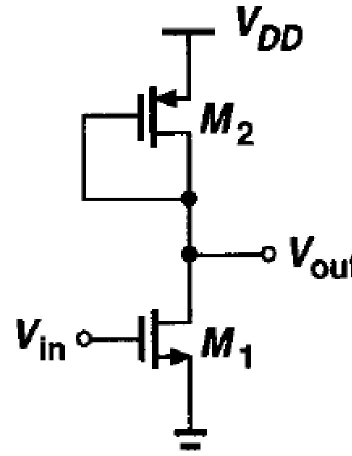
由*Equation* (3.26)

$$A_v \approx -\frac{g_{m1}}{g_{m2}} = -\frac{\mu_n\left(\frac{W}{L}\right)_1 (V_{GS1}-V_{TH1})}{\mu_p\left(\frac{W}{L}\right)_2 |V_{GS2}-V_{TH2}|}$$

(3. 35):

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

矛盾？



VGS2=VDS2不是独立变化的,而是与VGS1联动

$$\mu_n\left(\frac{W}{L}\right)_1 (V_{GS1}-V_{TH1})^2 \approx \mu_p\left(\frac{W}{L}\right)_2 (V_{GS2}-V_{TH2})^2$$

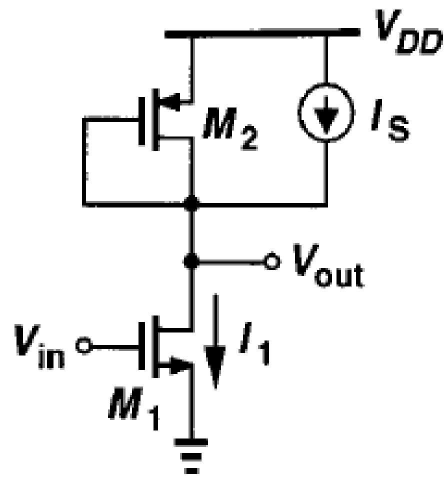
$$\frac{V_{GS1}-V_{TH1}}{|V_{GS2}-V_{TH2}|} = \sqrt{\frac{\mu_p\left(\frac{W}{L}\right)_2}{\mu_n\left(\frac{W}{L}\right)_1}}$$

$$A_v = -\frac{\mu_n\left(\frac{W}{L}\right)_1 (V_{GS1}-V_{TH1})}{\mu_p\left(\frac{W}{L}\right)_2 |V_{GS2}-V_{TH2}|}$$

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

Example 3.3： 增加gain的思路

减小负载管电流，增加负载电阻（1/gm2）。



M_2电流减小到I_1的 $1/K$，即增加电流源$I_S = \frac{K-1}{K} I_1$

$\mu_n \left(\frac{W}{L}\right)_1 (V_{GS1} - V_{TH1})^2 \approx K\mu_p \left(\frac{W}{L}\right)_2 (V_{GS2} - V_{TH2})^2$，未考虑$\lambda$

得到：$\sqrt{\frac{\mu_n \left(\frac{W}{L}\right)_1}{\mu_p \left(\frac{W}{L}\right)_2}} = \sqrt{K} \frac{|V_{GS2} - V_{TH2}|}{V_{GS1} - V_{TH1}}$

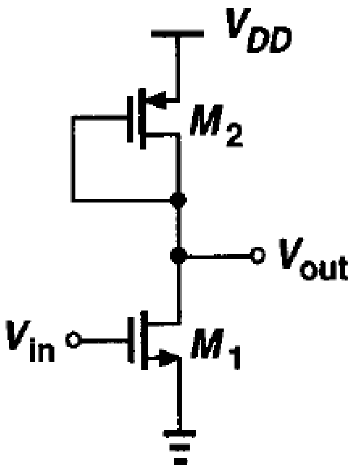
电流源输出电阻无穷大

$$A_V \approx -g_{m1} \times \frac{1}{g_{m2}} = -\frac{\sqrt{2\mu_n C_{OX} \left(\frac{W}{L}\right)_1 I_{D1}}}{\sqrt{2\mu_p C_{OX} \left(\frac{W}{L}\right)_2 I_{D2}}}$$

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

不仅增加增益，而且减小了M2过驱动电压，相当于扩大了输出动态范围

沟道长度调制效应等效于ro阻抗

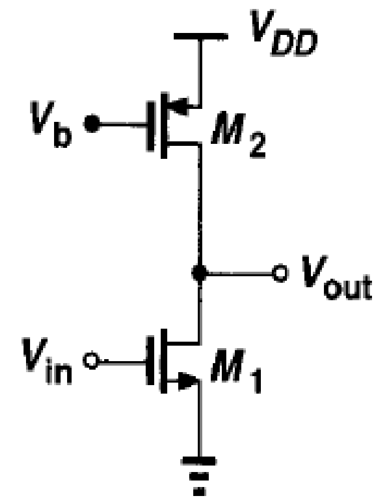


$$A_V \approx -g_{m1} \times R_D = -g_{m1}\left(\frac{1}{g_{m2}} \,||\, \mathrm{r_{o1}} \,||\, \mathrm{r_{o2}}\right)$$

3.2.4 CS Stage with current-source Load

实际电阻做负载，限制了直流电流和工作点，
一般情况下不能很大。

饱和区M2电流源（L2大）做CS负载，其等效阻抗为ro=rds很大（理想为无穷大）。



交变小信号增益：$A_v = -g_m(r_{o1} \;||\; r_{o2})$

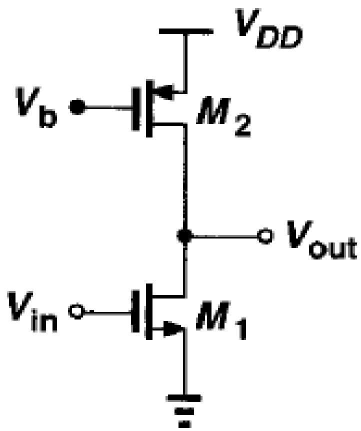
输出直流工作点电压可变化范围很大：

$$V_b + |V_{TH2}| > V_{out} > V_{in} - V_{TH1}$$

$$L \uparrow \Rightarrow r_o \uparrow \Rightarrow A_v \uparrow$$

缺点：电流源负载CS放大电路的输出DC电位需要通过其他方法确定。

Increasing Intrinsic gain by inc. L1 or dec. ID



- **Intrinsic gain(本征增益):**

$$g_{m1}r_{O1} = \sqrt{2\left(\frac{W}{L}\right)_1 \mu_n C_{ox} I_D}\,\frac{1}{\lambda I_D}$$

$$r_{01} \approx \frac{1}{\lambda I_D} \propto \frac{L_1}{I_D}$$

$$\therefore\ g_{m1}r_{o1} \propto \sqrt{2\mu_p C_{ox} \frac{W_1 L_1}{I_D}}$$

输入管（信号放大管）本征增益
会随着 I_D减小或MOSFET 栅极面积增加而增加。

3.2.5 CS with source degeneration

- 源极负反馈(源简并)

$$V_{out} = V_{DD} - I_D R_D$$

设计思想：采用负反馈，
减小I_D与Vin之间的非线性。

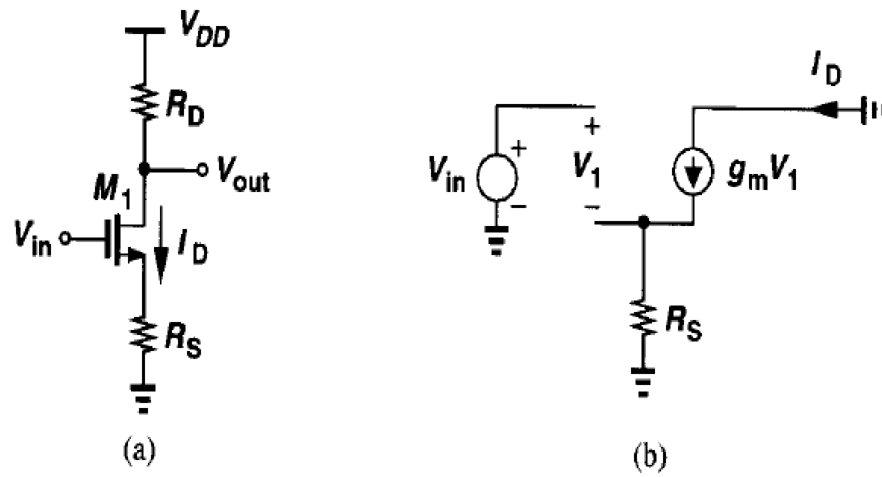


图3.16 带源极负反馈的共源极

输入电压Vin一部分损失在R_S上，

V_{GS}减小，使得 I_D非线性项减小。

$R_S = 0$时：$A_V = -g_m R_D$，g_m定义为*MOSFET*的跨导

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \mu_n C_{ox}(W/L)(V_{GS} - V_{TH}) = \sqrt{2\mu_n C_{OX}\frac{W}{L}I_D}$$

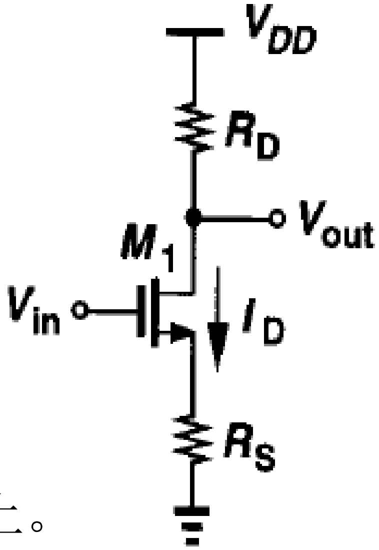
小信号电压增益与电路等效跨导

$R_S \neq 0$时：$V_{GS} = V_{in} - I_D R_S$，设$I_D = f(V_{GS})$

定义电路跨导$G_m = \dfrac{\partial I_D}{\partial V_{in}} = \dfrac{\partial I_D}{\partial V_{GS}} \dfrac{\partial V_{GS}}{\partial V_{in}} = \dfrac{\partial I_D}{\partial V_{GS}}(1 - \dfrac{\partial I_D}{\partial V_{in}} R_S) = g_m(1 - G_m R_S)$

得到：$G_m = \dfrac{g_m}{1 + g_m R_S} = \dfrac{1}{\dfrac{1}{g_m} + R_S}$

不考虑λ时$A_V = \dfrac{\partial V_{out}}{\partial V_{in}} = -\dfrac{\partial I_D}{\partial V_{in}} R_D = -G_m R_D = \dfrac{-g_m R_D}{1 + g_m R_S}$



当$g_m R_S >> 1$时，$A_V = -G_m R_D \approx \dfrac{-R_D}{R_S}$，$\therefore G_m \approx \dfrac{1}{R_S}$

此时$\Delta V_{in} = \Delta I_D / G_m \approx R_S \Delta I_D$，表明输入的变化信号基本都加在$R_S$上。

交变输出电流ΔI_D以及电压ΔV_{out}的线性变好，代价是增益变小。

求有源极反馈电组CS增益的简便方法

$$A_V = -G_m R_D = -\frac{g_m}{1+g_m R_S} R_D = -\frac{R_D}{\frac{1}{g_m}+R_S}$$

忽略沟道长度调制效应

- **增益=漏极通路电阻/源极通路从地向上看到的电阻**

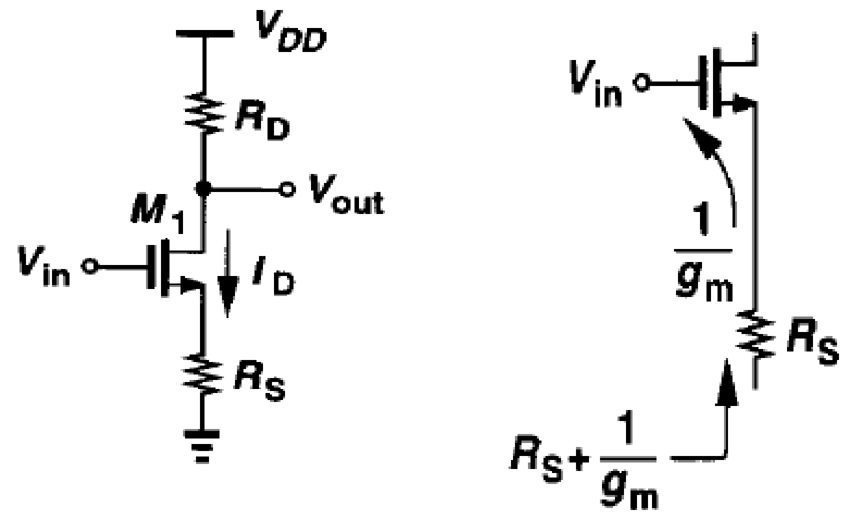
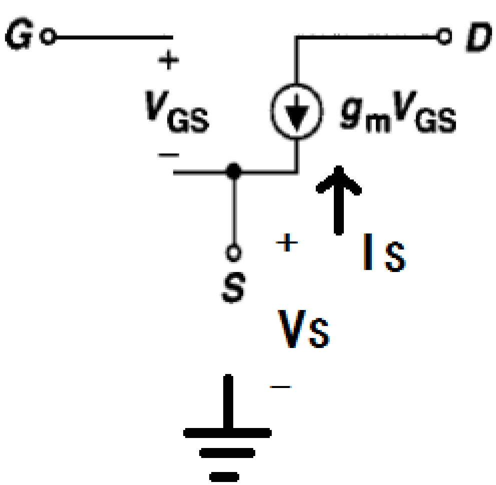


Figure 3.20 resistance seen in the source path

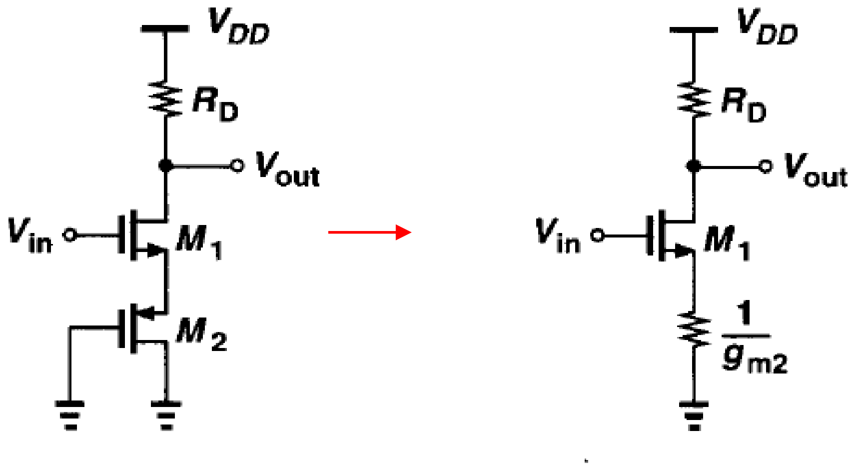


$$MOS的S极小信号电阻 = \frac{V_S}{I_S} = \frac{-V_{GS}}{-g_m V_{GS}} = \frac{1}{g_m}$$

（∵ 求输出阻抗时输入电压 = 0，即G交流接地）

Example 3.5

- Assuming $\lambda = \gamma = 0$



(a) (b)

- solution

$$A_v = -\frac{R_D}{\dfrac{1}{g_{m1}} + \dfrac{1}{g_{m2}}}.$$

辅助定律（电路基本理论）

在线性电路中，开路电压增益等于-GmRout，其中Gm表示**输出与地短接（恒压）**时的**电路跨导**（Gm=Iout/Vin）。Rout表示当输入电压为零时（Iout= GmVin=0 ）电路的输出电阻。

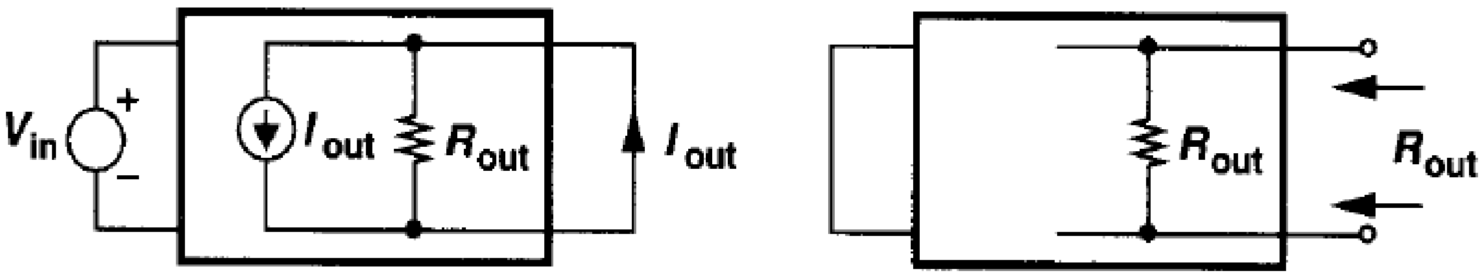


Figure 3.25 Modeling output port of an amplifier by a Norton equivalent.

输出开路（不接负载）电压：$V_{out} = -I_{out}R_{out} = -G_m V_{in} R_{out}$

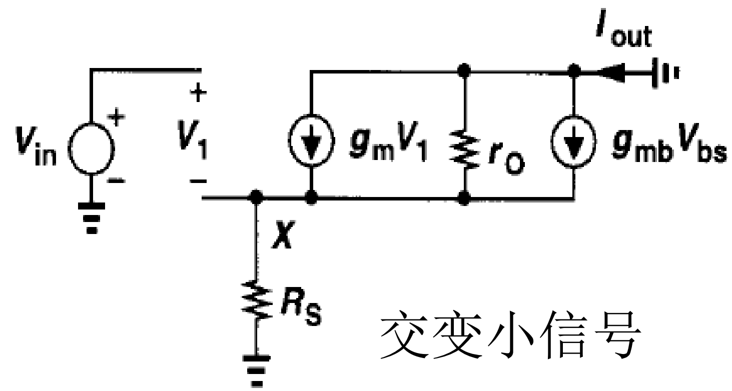
带负载R_D时输出电压：$V_{out} = -I_{out}(R_{out} \,||\, R_D)$

$= -V_{in}G_m(R_{out} \,||\, R_D)$

负载可后加计算

Gm with body effect and channel-length modulation

前例图*3.16*： $V_{in} = V_1 + I_{out}R_S$



考虑沟道长度调制和体效应。

交变小信号

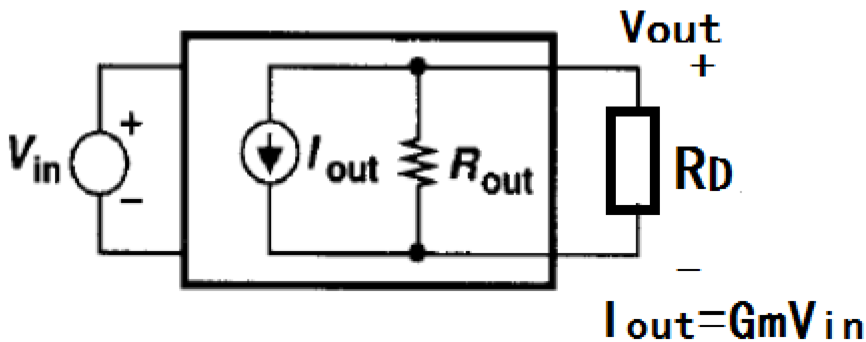
如何求电路交变小信号跨导： $G_m = \dfrac{I_{out}}{V_{in}}$?

$$\text{输出短路交变电流} I_{out} = g_m V_1 + g_{mb} V_{bs} - \frac{V_X}{r_o} = g_m V_1 - g_{mb} V_X - \frac{V_X}{r_o}$$

$$= g_m V_1 - g_{mb} I_X R_S - \frac{I_X R_S}{r_o} = g_m (V_{in} - I_{out} R_S) - g_{mb} I_{out} R_S - \frac{I_{out} R_S}{r_o}$$

$$\longrightarrow \quad G_m = \frac{I_{out}}{V_{in}} = \frac{g_m r_o}{R_S + [1 + (g_m + g_{mb}) R_S] r_o}$$ ，式（3. 55）

$$A_v = -G_m (R_{out} \parallel R_D) \approx -G_m R_D$$



Rout见式（3.60）

Output resistance of degenerated CS

$$R_{out} = \frac{V_X}{I_X} \quad 令 V_{in} = 0，先不考虑 R_D$$

$$V_{bs} = V_1 = -I_X R_S$$

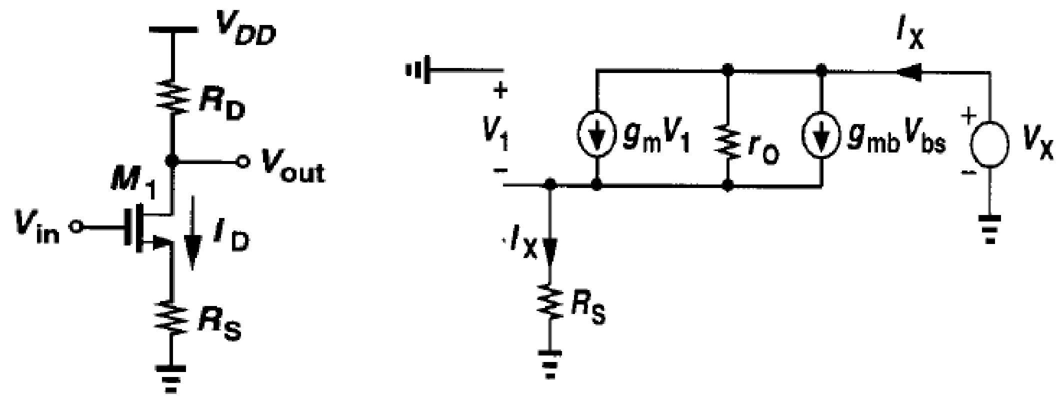


Figure 3.22 Equivalent circuit for calculating the output resistance of a degenerated CS stage.

$$V_X = r_0[I_X - g_m V_1 - g_{mb} V_{bs}] + I_X R_S$$
$$= r_0[I_X + (g_m + g_{mb}) I_X R_S] + I_X R_S$$

$$R_{out} = \frac{V_X}{I_X} = [1 + (g_m + g_{mb})R_S]r_0 + R_S = [1 + (g_m + g_{mb})r_0]R_S + r_0，式(3.60)$$

$$\because (g_m + g_{mb})r_0 >> 1$$

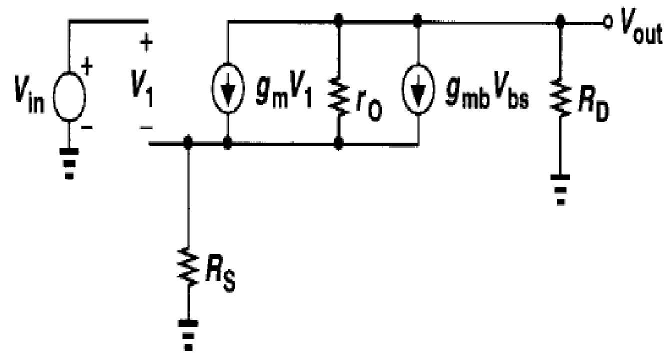
$$\therefore R_{out} \approx (g_m + g_{mb})r_0 \times R_S + r_0$$

源极负反馈:增大共源级的输出电阻。

验证：包含RD的公式推导

$$I_{RS} = -I_{RD} = -\frac{V_{out}}{R_D}$$

$$V_{bs} = -V_{R_S}$$



$$I_{ro} = -I_{RD} - (g_m V_1 + g_{mb} V_{bs}) = -\frac{V_{out}}{R_D} - [g_m (V_{in} - V_{RS}) - g_{mb} V_{RS}]$$

$$= -\frac{V_{out}}{R_D} - [g_m (V_{in} + \frac{V_{out}}{R_D} R_S) + g_{mb} \frac{V_{out}}{R_D} R_S]$$

$$V_{out} = I_{ro} r_o + V_{RS} = I_{ro} r_o - \frac{V_{out}}{R_D} R_S$$

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

$$A_V = \frac{V_{out}}{V_{in}} = -\frac{g_m R_D r_o}{R_D + R_S + r_o + (g_m + g_{mb}) R_S r_o} = -\frac{g_m R_D r_o}{R_D + \{[1 + (g_m + g_{mb}) r_o] R_S + r_o\}}$$

(3.71)

验证：包含RD的公式推导（续）

分析Av分母大括号项（=带负反馈共源极的输出电阻Rout）。

$$R_{out}=[1+(g_m+g_{mb})r_0]R_S+r_0，\text{式}(3.60)$$

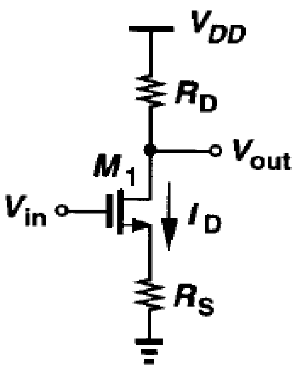
$$A_V=-\frac{g_mR_Dr_o}{R_D+\{[1+(g_m+g_{mb})r_o]R_S+r_o\}}\bullet\frac{[1+(g_m+g_{mb})r_o]R_S+r_o}{[1+(g_m+g_{mb})r_o]R_S+r_o}$$

$$=-\frac{g_mr_o}{[1+(g_m+g_{mb})r_o]R_S+r_o}\bullet\frac{R_D\{[1+(g_m+g_{mb})r_o]R_S+r_o\}}{R_D+[1+(g_m+g_{mb})r_o]R_S+r_o}$$

$$=-G_m(R_D\ ||\ Rout)$$

与先不计算**RD**，再将**RD**加入电路的计算方法结果一致

Gain of a degenerated CS



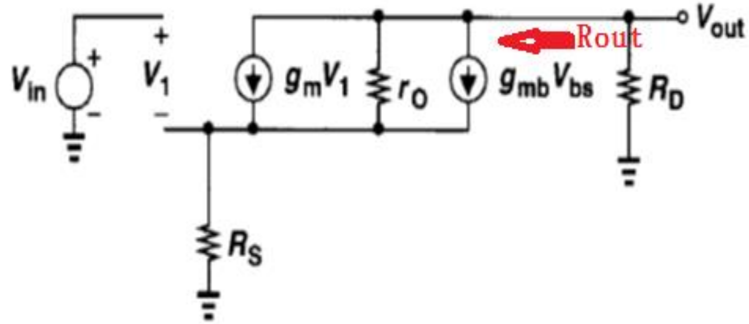
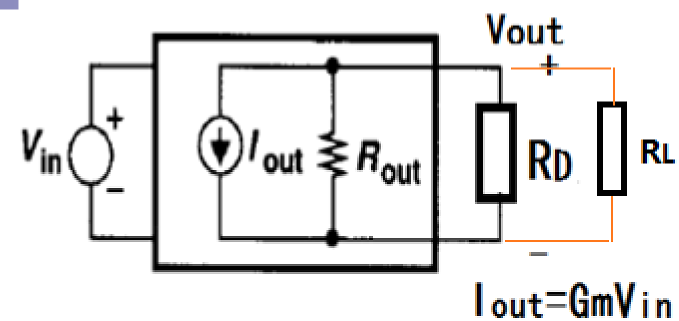
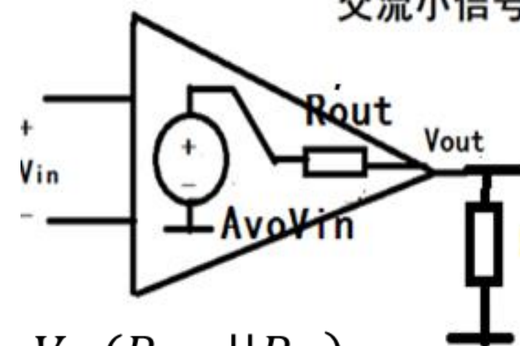


Figure 3.24 Small-signal model of degenerated CS stage with finite output resistance.



交流小信号



负载R_L开路输出：$V_{outO} = -I_{out}(R_{out}||R_D) = -G_mV_{in}(R_{out}||R_D)$

R_{out}是向MOS漏极看进去的输出阻抗。

Gm为输出短路电路跨导, 式3.55

总输出阻抗为$R_{out}' = R_{out}||R_D$

记负载开路电压增益$A_{vo} = -G_m(R_{out}||R_D) = -G_mR_{out}'$

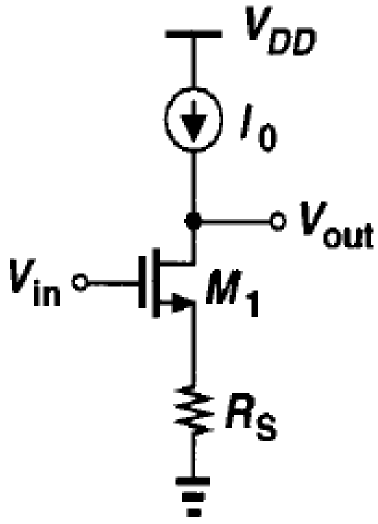
带负载R_L？ RL与RD并联，或$A_v = A_{vo}\dfrac{R_L}{R_{out}' + R_L} = -G_m(R_{out}'||R_L)$）

例 3.6 calculate the voltage gain

Solution

$$A_v = -\frac{g_m R_D r_o}{R_D + \{[1 + (g_m + g_{mb}) r_o] R_S + r_o\}}$$

$$= -g_m r_o, \text{式}(3.71)\text{中} R_D = \infty$$



$$\text{或} A_v = -G_m (Rout \,||\, R_D) = -G_m Rout$$

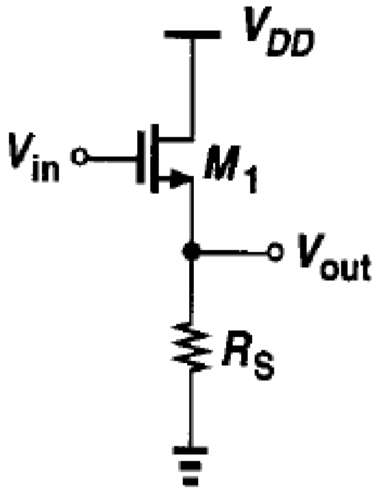
$$= -\frac{g_m r_o}{[1 + (g_m + g_{mb}) r_o] R_S + r_o} \bullet \{[1 + (g_m + g_{mb}) r_o] R_S + r_o\}$$

$$= -g_m r_o$$

与Rs无关。原因是电流不变，Rs上的小信号（变化）压降=0

3.3 source follower 源跟随器

common-drain stage



- 特点:

小信号电压近似等于输入电压；

低频输入阻抗大，

输出阻抗小（$1/g_m)||R_s$。

作用:

当两级放大器级联时，若前级输出电阻大，或后级输入电阻小，则在两级放大器之间插入源跟随器，可以提高多级放大器的电压增益。

Large-signal & Small-signal behavior

忽略沟道长度调制效应

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

$$\text{小信号}A_v = \frac{\partial V_{out}}{\partial V_{in}}$$

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

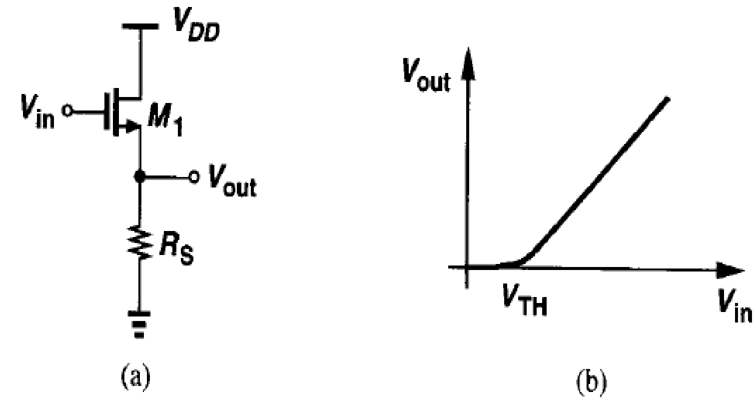


Figure 3.27 (a) Source follower, and (b) its input-output characteristic.

$$\text{由式(2. 42)得}\ \frac{\partial V_{TH}}{\partial V_{in}} = \frac{\partial V_{TH}}{\partial V_{SB}}\frac{\partial V_{SB}}{\partial V_{in}} = \eta\frac{\partial V_{out}}{\partial V_{in}} = \eta A_v$$

$$\therefore\ A_v = \mu_n C_{ox}\frac{W}{L}(V_{in} - V_{out} - V_{TH})(1 - A_v - \eta A_v)R_S$$

$$\eta = \frac{g_{mb}}{g_m} = \frac{\partial I_D}{\partial V_{BS}} \Bigg/ \frac{\partial I_D}{\partial V_{GS}}$$

Gain of Source Follower

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} == \frac{\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{out} - V_{TH})R_S}{1 + \mu_n C_{ox}\frac{W}{L}(V_{in} - V_{out} - V_{TH})R_S(1+\eta)} = \frac{g_m R_S}{1 + (g_m + g_{mb})R_S} < 1$$

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

$$g_{mb} = \eta g_m = \frac{\gamma}{2\sqrt{2\Phi_F + V_{SB}}} g_m$$，式（2.44）

小信号等效电路得到相同结果

$$V_1 = V_{in} - V_{out}$$

$$V_{bs} = -V_{sb} = -V_{out}$$

为何ro可忽略?

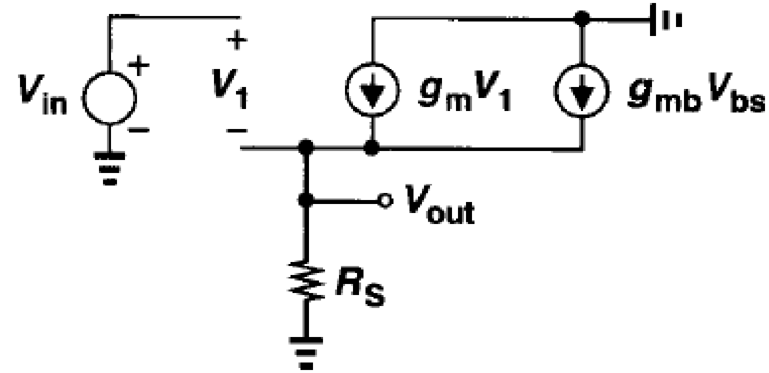


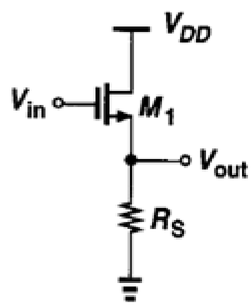
Figure 3.28 Small-signal equivalent circuit of source follower.

$$\therefore \quad g_m V_1 + g_{mb} V_{bs} = g_m (V_{in} - V_{out}) - g_{mb} V_{out} = \frac{V_{out}}{R_S}$$

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} \text{(即等效电路图中} \frac{V_{out}}{V_{in}}\text{)}$$

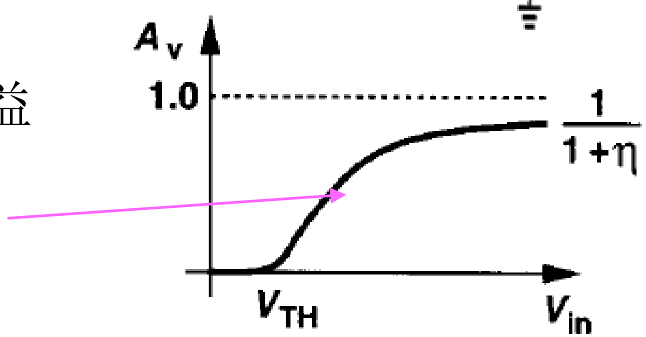
$$= \frac{g_m R_S}{1 + (g_m + g_{mb}) R_S} < 1$$

$$\eta = \frac{g_{mb}}{g_m}$$ 可取0. 2估算

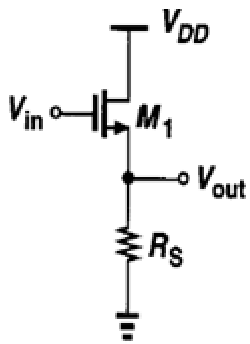


当$g_m R_s >> 1$时

$$A_V \approx \frac{g_m R_S}{(g_m + g_{mb}) R_S} = \frac{1}{1+\eta}$$ 达到“不变”的最大增益



采用电流源的源跟随器



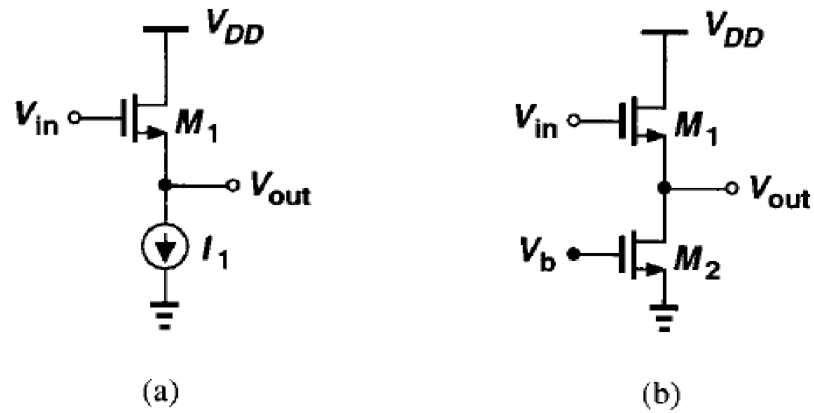


Figure 3.30 Source follower using an NMOS transistor as current source.

$R_s = \propto$ 时

$$A_V = \frac{g_m R_S}{1+(g_m+g_{mb})R_S} = \frac{1}{1+\eta}$$

最大增益=本征增益

- **保证M2处于饱和区。** $V_b - V_{out} < V_{TH2}$

设计注意事项：

1。体效应导致VTH变化，导致输入输出之间的非线性（大输入范围时）；

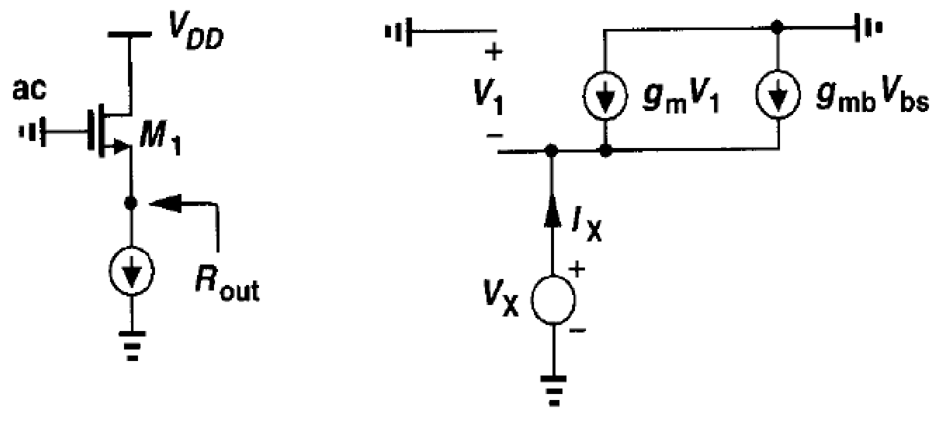
2。直流电平移动导致电压余度(headroom)减少。

源跟随器的输出电阻

$$\sup pose : \lambda \approx 0$$

$$V_1 = -V_X,$$

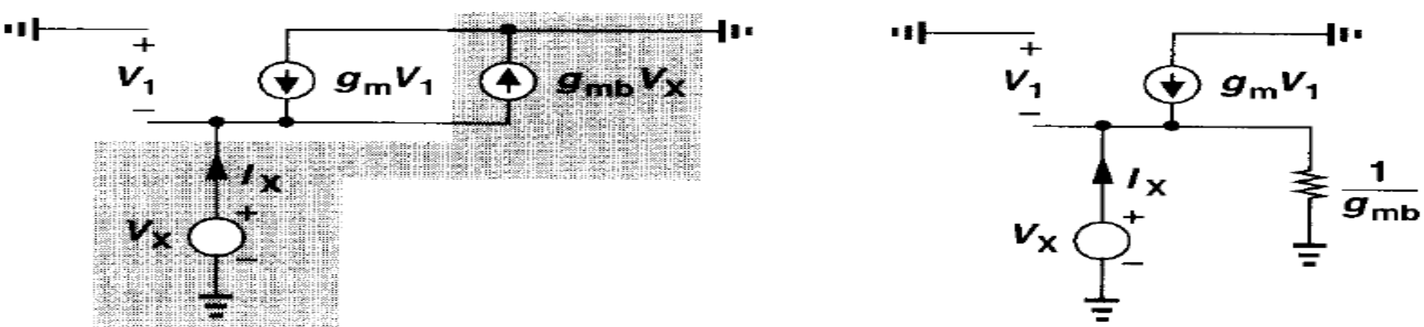
$$I_X + g_m V_1 + g_{mb} V_{bs}$$
$$= I_X - g_m V_X - g_{mb} V_X = 0$$



(a) (b)

$$R_{out} = \frac{V_X}{I_X} = \frac{1}{g_m + g_{mb}} = \frac{1}{g_m} \,||\, \frac{1}{g_{mb}}$$

与二极管相同!几十欧~几千欧.



(a) (b)

Figure 3.32 Source follower including body effect.

另一种方法：用辅助定理求Intrinsic gain of SF

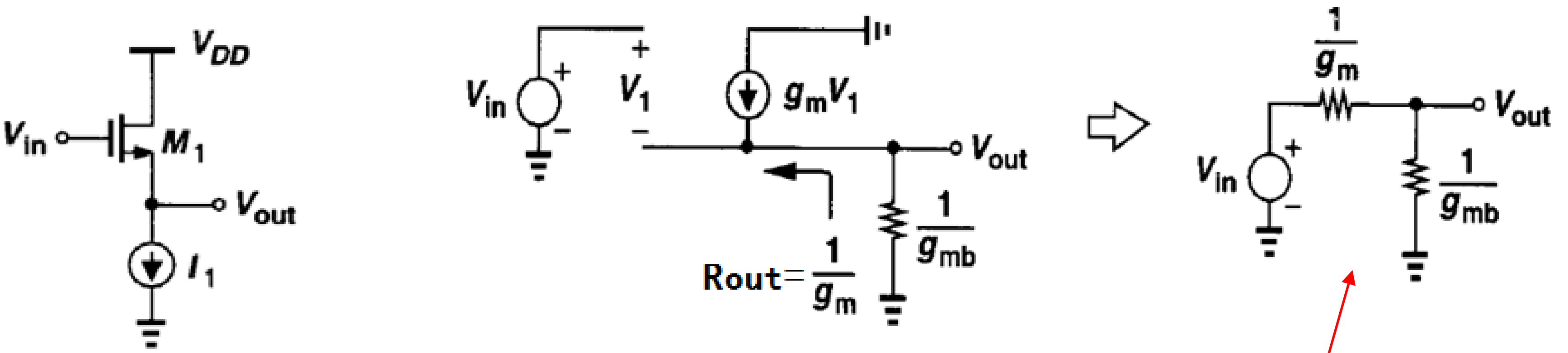


Figure 3.33 Representation of intrinsic source follower by a Thevenin equivalent.

总结：源跟随器MOS看成有一内阻1/gm 串联1/gmb。
若有负载阻抗，则与1/gmb并联。

求R_{out}：令$V_{in}=0$，则$V_{out}=-V_1$

$$R_{out}=\frac{V_{out}}{-g_m V_1}=\frac{1}{g_m}$$

交流输出短路（$V_1=V_{in}$）电流：　$I_{SC}=g_m V_1=g_m V_{in}$

开路输出电压：　$V_{OC}=R_{out}I_{SC}=\frac{1}{g_m}g_m V_{in}=V_{in}$

$$A_V=\frac{V_{out}}{V_{in}}=\frac{1/g_{mb}}{1/g_m+1/g_{mb}}=\frac{g_m}{g_m+g_{mb}}=\frac{1}{1+\eta}$$，无负载时

考虑体效应和沟道长度调制效应的SF

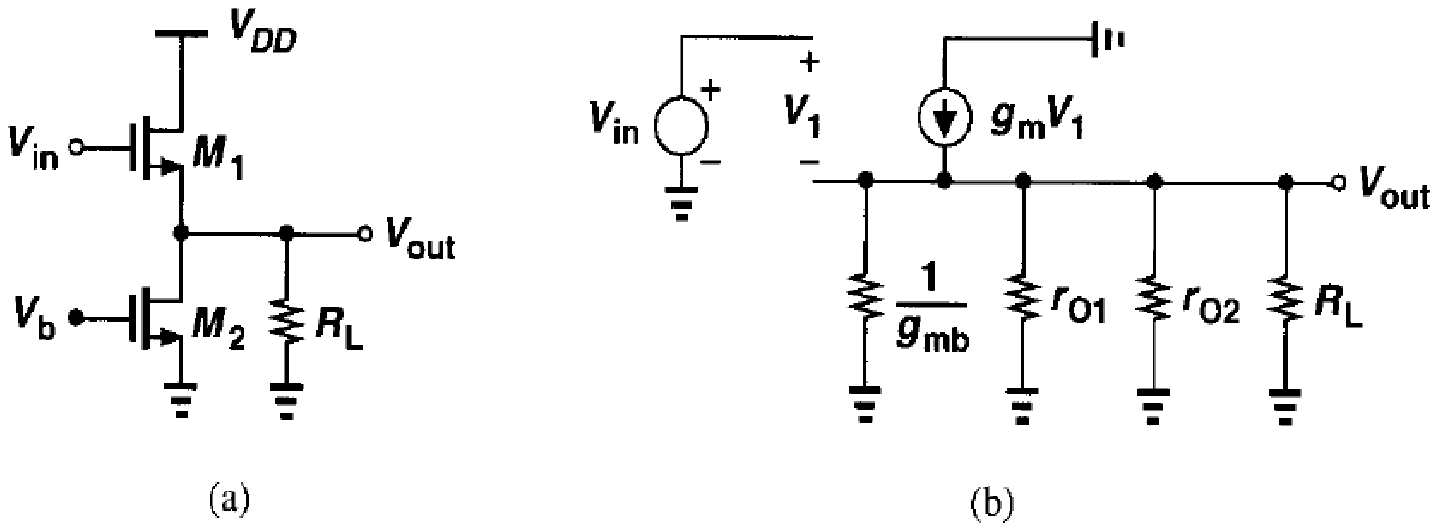


Figure 3.34 (a) Source follower driving load resistance, (b) small-signal equivalent circuit.

gmb指gmb1

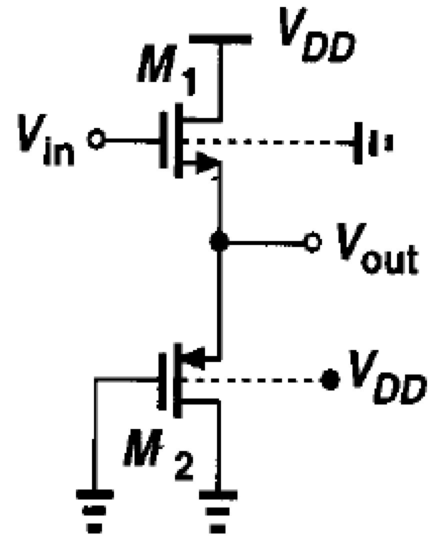
(Vin>VGS1+ Vb-VTH2)

$$A_v = \frac{\frac{1}{g_{mb}} \| r_{O1} \| r_{o2} \| R_L}{\frac{1}{g_{mb}} \| r_{O1} \| r_{o2} \| R_L + \frac{1}{g_m}} \tag{3.90}$$

公式中的gm和gmb是M1参数.

例3.8

- M2为二极管，由式（3.24）和（3.90）



$$A_v = \frac{\frac{1}{g_{mb1}} \,||\, r_{o1} \,||\, r_{o2} \,||\, \frac{1}{g_{m2}+g_{mb2}}}{\frac{1}{g_{mb1}} \,||\, r_{o1} \,||\, r_{o2} \,||\, \frac{1}{g_{m2}+g_{mb2}} + \frac{1}{g_{m1}}}$$

$$= \frac{\frac{1}{g_{mb1}} \,||\, \frac{1}{g_{m2}+g_{mb2}}}{\frac{1}{g_{mb1}} \,||\, \frac{1}{g_{m2}+g_{mb2}} + \frac{1}{g_{m1}}}$$，M2电阻小，不好

NMOS SF输出电平低于输入至少VTH

PMOS的SF:消除体效应的非线性

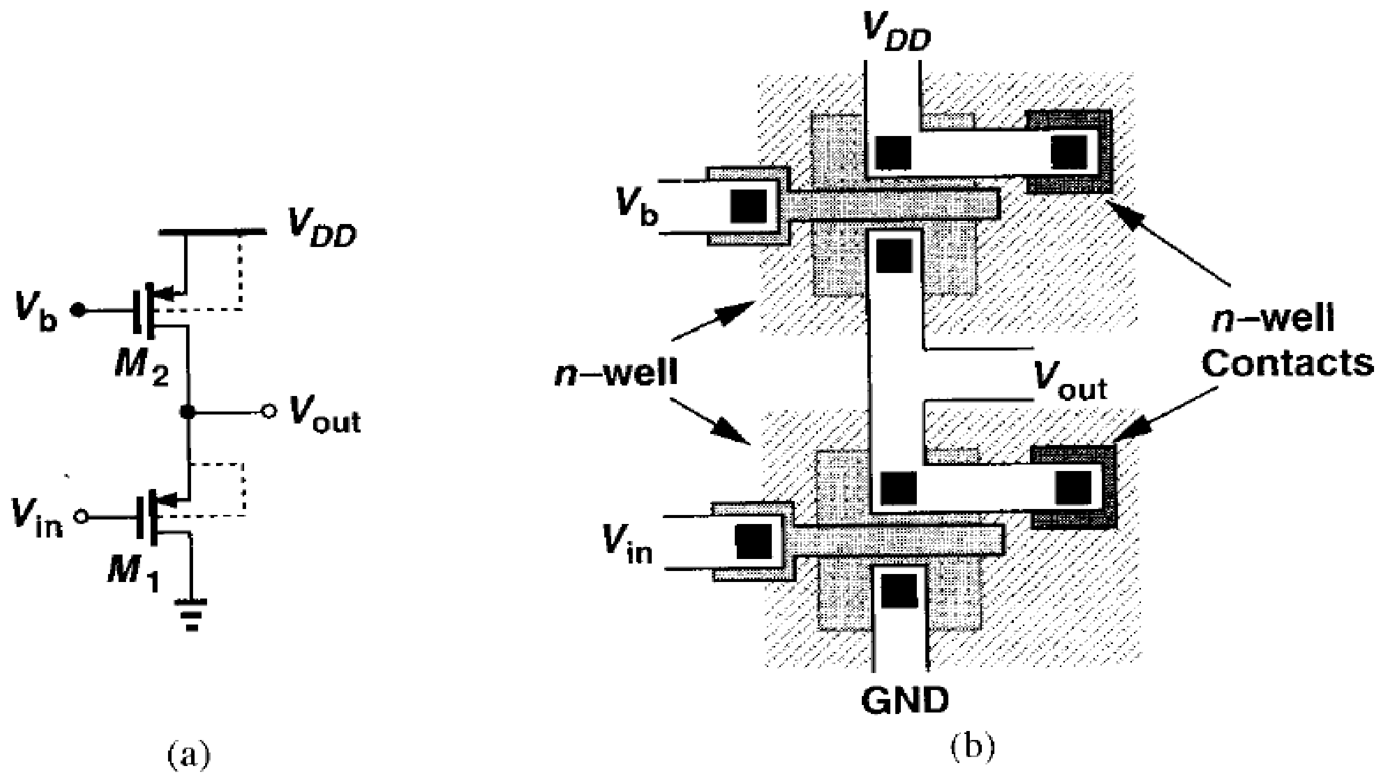
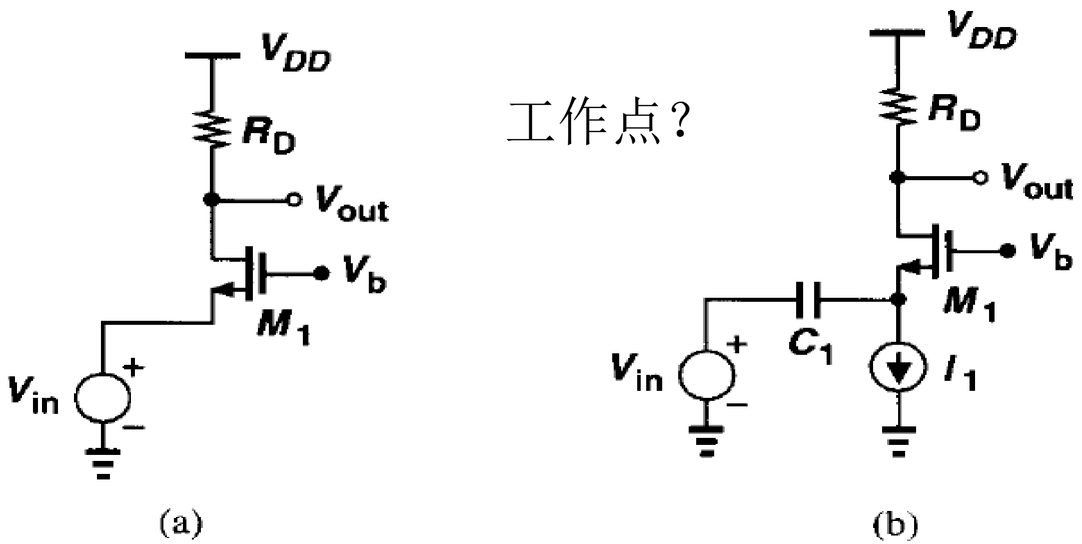


Figure 3.36 PMOS source follower with no body effect.

PMOS SF输出电平高于输入至少VTH

3.4 Common-gate stage



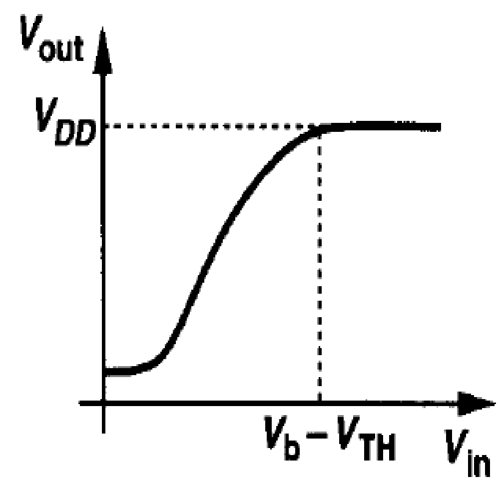
工作点？

Figure 3.40 (a) Common-gate stage with direct coupling at input, (b) CG stage with capacitive coupling at input.

信号从源端输入，从漏极输出。小信号输入输出共栅

M1可用恒流源偏置，
高频信号通过C耦合；
也可直接低频输入信号源。

图(a): For Vin>Vb-VTH
M1 is off, Vout=VDD

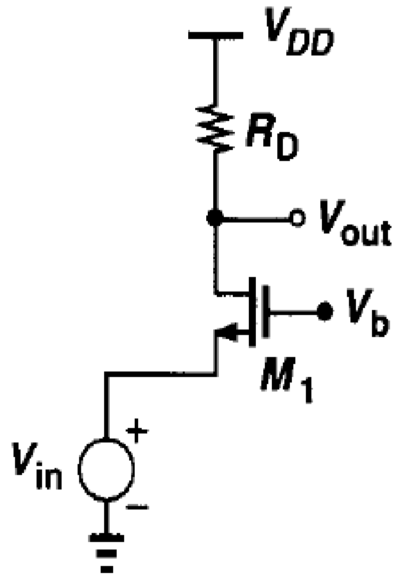


饱和区 $V_{in}\downarrow \Rightarrow I_D \uparrow \Rightarrow V_{out}\downarrow$，同向放大

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

降到 $V_b - V_{TH}$ 时进入线性区

Gain of CG



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

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

$$\because \frac{\partial V_{TH}}{\partial V_{in}} = \frac{\partial V_{TH}}{\partial V_{SB}} = \eta$$

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

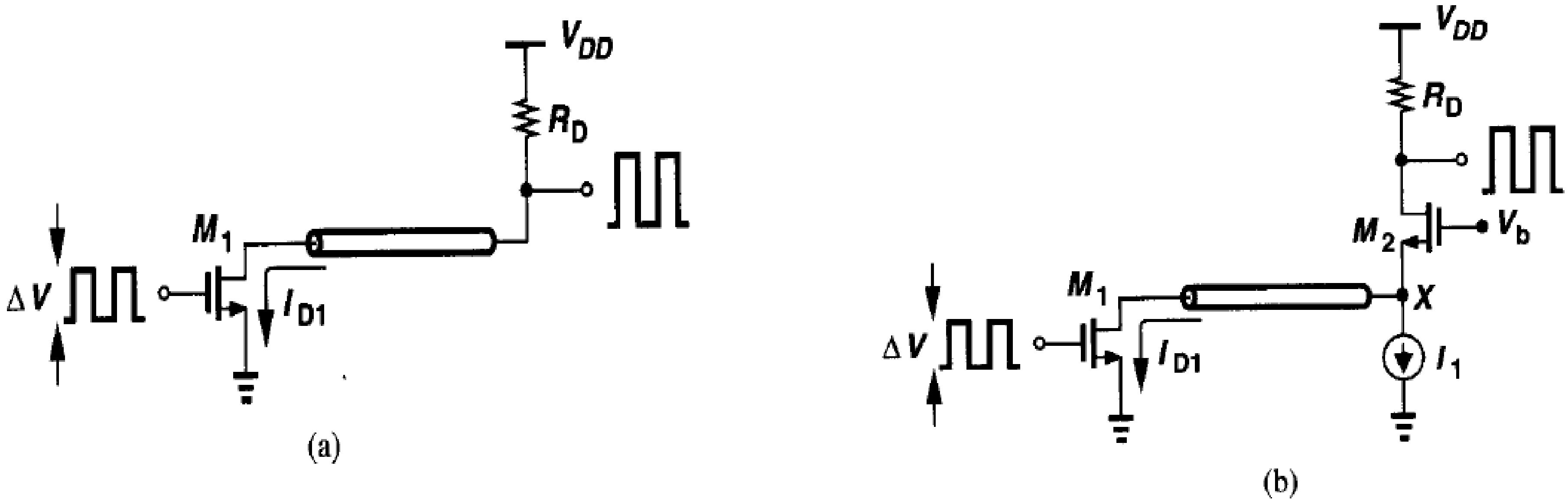
虽然体效应使共栅级的等效跨导变大，但并不会带来好的性能。

设$\lambda \approx 0$即ΔV_{DS}对I_D交变无影响，

或R_D不太大(I_D不恒定)时：

输入阻抗 $\approx \dfrac{1}{g_m + g_{mb}}$，即 $\dfrac{1}{g_m} \,||\, \dfrac{1}{g_{mb}}$

共栅级电路应用在低输入阻抗的场合



传输线电阻50欧，RD=50欧，忽略沟道长度调制 和体效应 。

- (a) 两种接法的低频增益（忽略MOS寄生电容）。
- (b) X点反射最小的条件？(即M2源极看到的交流电阻为 50欧)

例3.10 续

解：(a)两种接法输入Vin变化ΔV，$\Delta I_{D1}=g_m\Delta V$ 相同，

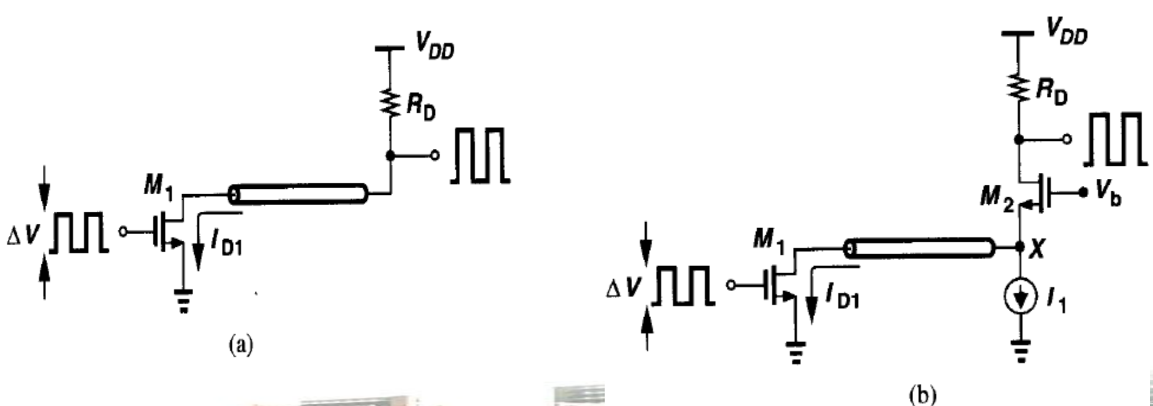
$$A_v=\frac{\partial V_{out}}{\partial V_{in}}=-g_mR_D$$ 公式形式相同，（a）图RD为50欧。

注意：图(b)共栅级电路M2的输入信号采用的是电流，而不是电压。

(b) 图X点看M2源极的小信号（交流）电阻 $\frac{\Delta V_X}{\Delta I_X}=\frac{1}{g_{m2}+g_{mb2}}=50\Omega$

$$g_m=\sqrt{2\mu_nC_{ox}\frac{W}{L}I_D}$$

图（b）由于M2的隔离，
RD可远大于 50欧，
提供更高的增益。

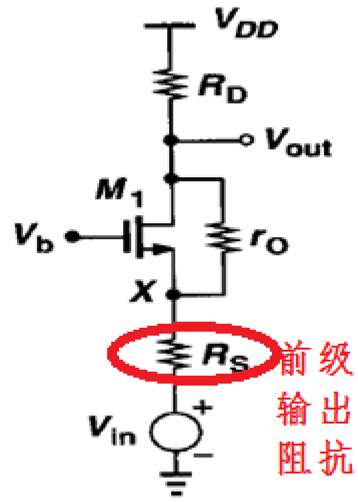


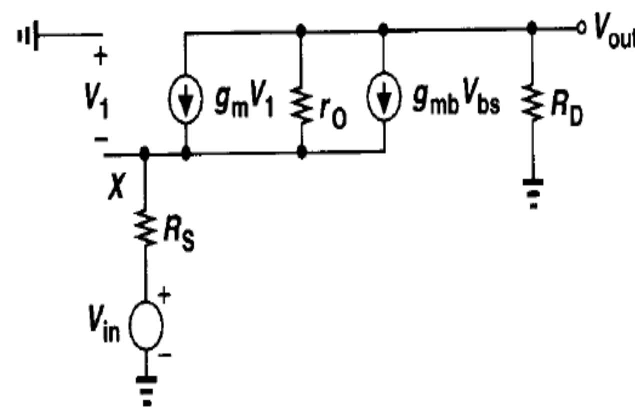
考虑MOS输出阻抗和信号源内阻情况下的CG

以下V表示交流小信号。

$$V_{out}=$$

$$r_o\left(\frac{-V_{out}}{R_D}-g_mV_1-g_{mb}V_{bs}\right)-\frac{V_{out}}{R_D}R_S+V_{in}$$

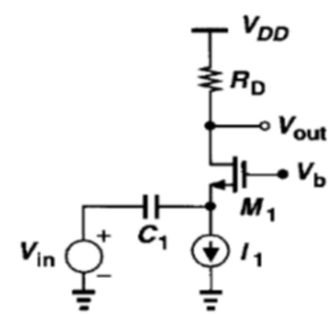




$$V_{bs}=V_1$$

图3.43 输出电阻有限的共栅级

$$V_1-\frac{V_{out}}{R_D}R_S+V_{in}=0$$



$$A_v=\frac{V_{out}}{V_{in}}=\frac{r_o(g_m+g_{mb})+1}{r_o+(g_m+g_{mb})r_oR_S+R_S+R_D}R_D \qquad (3.104)$$

若 $r_o=\infty,\quad g_m+g_{mb}\approx g_m$ 则 $A_V=\dfrac{R_D}{\dfrac{1}{g_m}+R_S}$ (3.999)

CG 比较CS

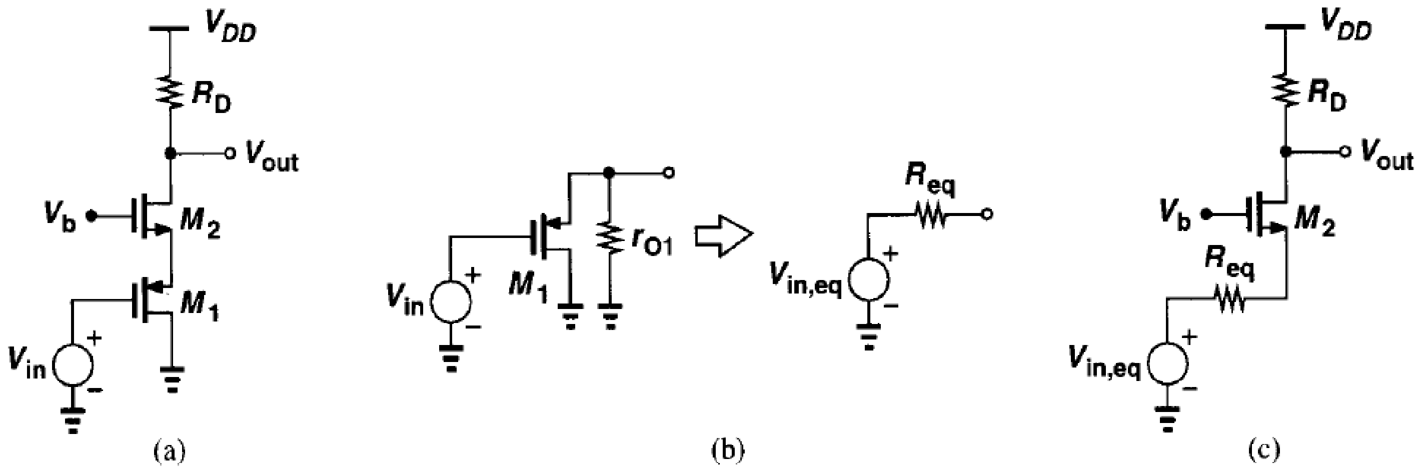
- CS $$A_v = \frac{V_{out}}{V_{in}} = \frac{-r_o g_m}{r_o + (g_m + g_{mb}) r_o R_S + R_S + R_D} R_D \quad (3.71)$$

- CG $$A_v = \frac{V_{out}}{V_{in}} = \frac{r_o (g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb}) r_o R_S + R_S + R_D} R_D \quad (3.104)$$

CG增益略高一些，CG噪声比较大。

例3.11求电压增益

$\lambda \neq 0, \quad \gamma \neq 0$



M1为PMOS源跟随器。将M1等效成戴维南电路，化为标准电路形式

求M1的戴维南电路的等效电阻时，Vin=0，则M1成二极管，由式（3.24）

$$R_{eq} = \frac{1}{g_{m1}} \, || \, r_{o1}$$

M1输出短路电流为 gm1Vin,
PMOS管gmb1一般为0

跟随器

$$M_1\text{的开路输出电压}\, V_{in,eq} = (g_{m1}V_{in})\, R_{eq} = g_{m1} \frac{\frac{1}{g_{m1}} \times r_{o1}}{\frac{1}{g_{m1}} + r_{o1}} V_{in} \approx V_{in}$$

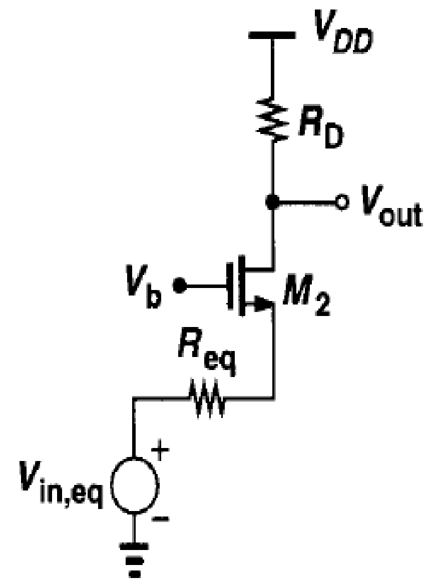
例3.11续

$$R_{eq}=\frac{1}{g_{m1}}\,||\,r_{o1}=\text{公式中}R_S$$

由CG: $A_V=\dfrac{V_{out}}{V_{in}}=\dfrac{r_o(g_m+g_{mb})+1}{r_o+(g_m+g_{mb})r_oR_S+R_S+R_D}R_D$ (3. 104)

得：$A_V=\dfrac{V_{out}}{V_{in}}=\dfrac{V_{out}}{V_{in,eq}}\times\dfrac{V_{in,eq}}{V_{in}}$

$$=\frac{r_{o2}(g_{m2}+g_{mb2})+1}{r_{o2}+[1+(g_{m2}+g_{mb2})r_{o2}](\frac{1}{g_{m1}}\,||\,r_{o1})+R_D}R_D$$



(c)

共栅级输入电阻

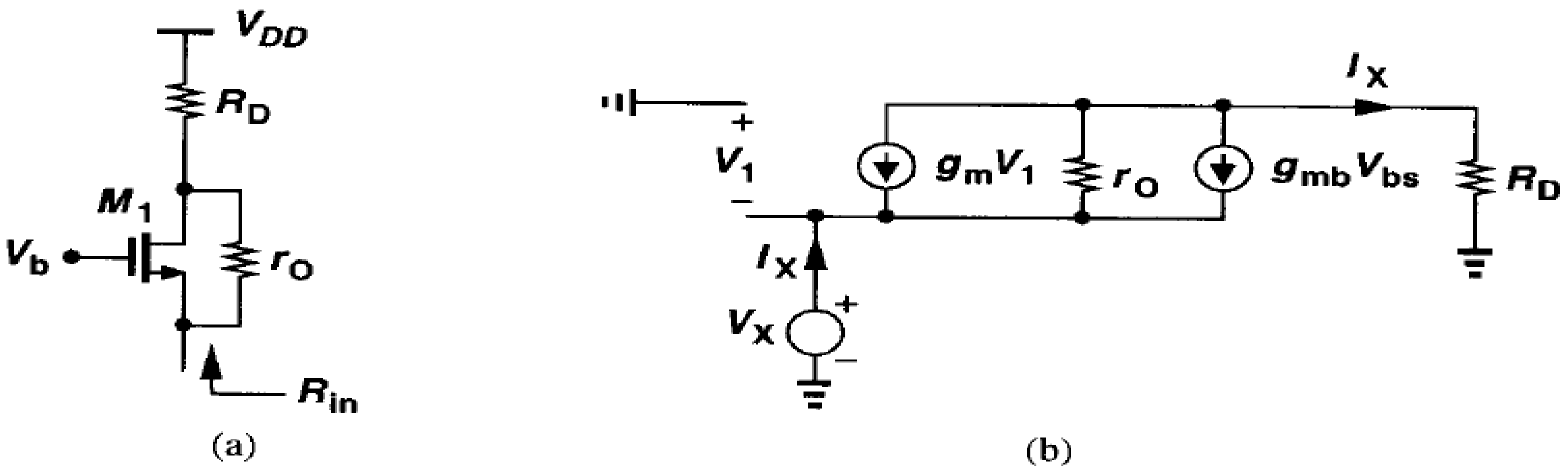


Figure 3.45 (a) Input resistance of a CG stage, (b) small-signal equivalent circuit.

$$V_1 = V_{bs} = -V_X$$ （V小信号）

从下往上流过ro的电流：

$$I_X + g_m V_1 + g_{mb} V_{bs} = I_X + g_m V_1 + g_{mb} V_1$$

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

$$V_X = [I_X - (g_m + g_{mb}) V_X] r_o + I_X R_D$$

共栅级输入电阻（从源极端往MOS看）

$$R_{in} = \frac{V_X}{I_X} = \frac{R_D + r_o}{1 + (g_m + g_{mb}) r_o} \qquad \textbf{(3.109)}$$

CG极端情况:RD=0和无穷大

$$R_{in}=\frac{V_X}{I_X}=\frac{R_D+r_o}{1+(g_m+g_{mb})r_o}$$

（1）$R_D=0$或r_o很大时$R_{in}\approx\dfrac{1}{g_m+g_{mb}}$

比较源跟随器输出阻抗（3.85）式相同

（2）图3.46，$R_D=\infty$时$R_{in}=\infty$

原因：交变小信号$I_X=0$

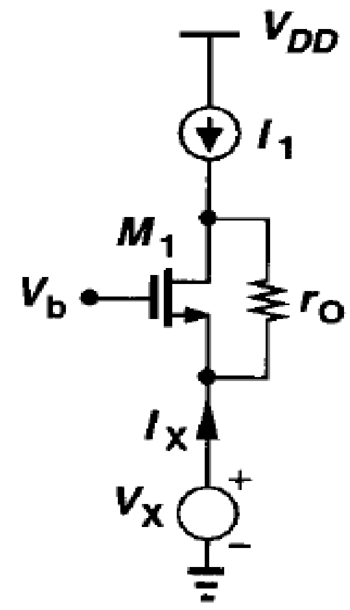


Figure 3.46 Input resistance of a CG stage with ideal current source load.

只有RD不很大情况下，
共栅级输入电阻才会较小。

直接耦合的共栅极。 $R_D = \infty$

由式（3.104 ）

$$A_v = \frac{V_{out}}{V_{in}}$$

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

$$= r_o(g_m + g_{mb}) + 1$$

与**Rs**无关。

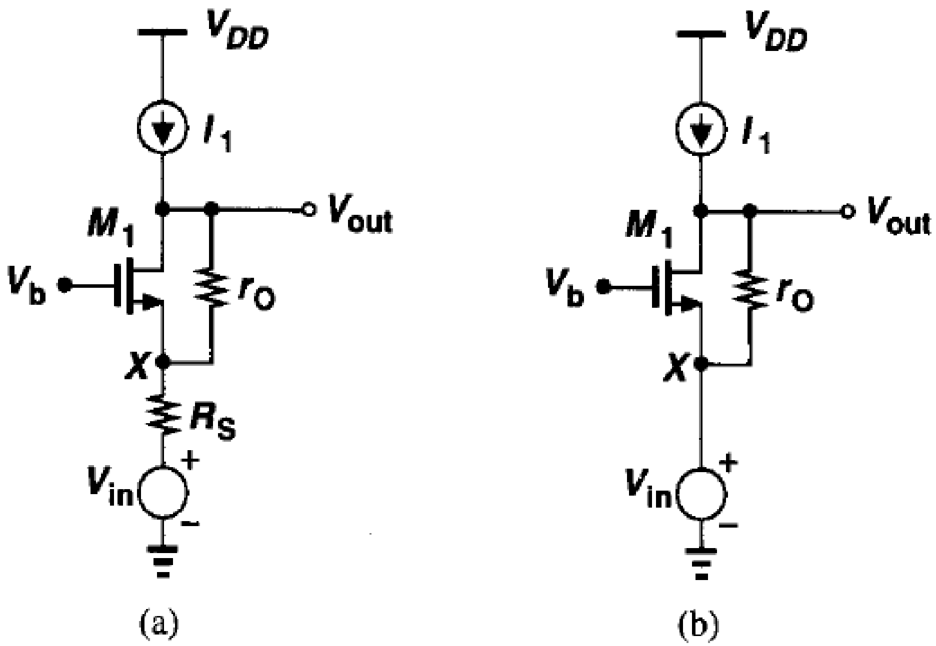
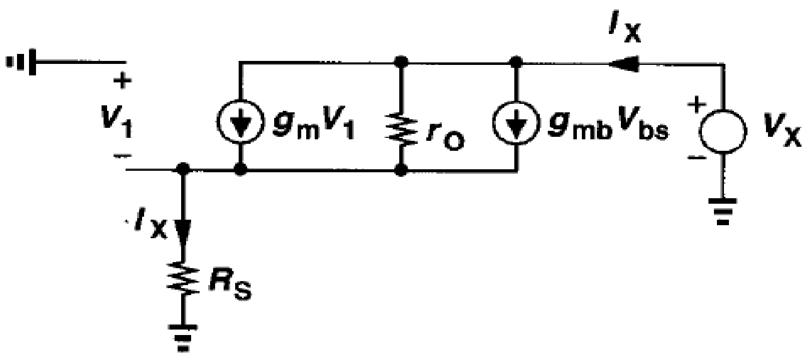


Figure 3.47

the small-signal voltage at node X is equal to Vin

与图3.22相同。



式(3. 62)： $R_{out} = [1 + (g_m + g_{mb})r_o]R_S + r_o$

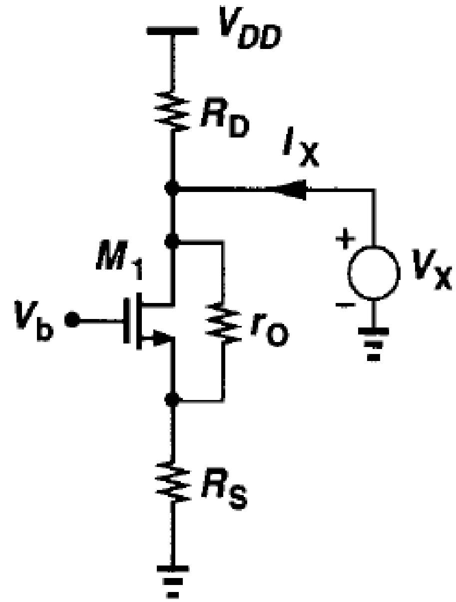
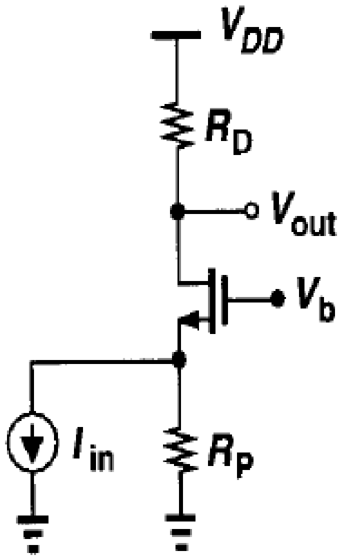


Figure 3.48 Calculation of output resistance of a CG stage.

Same as Degenerated CS

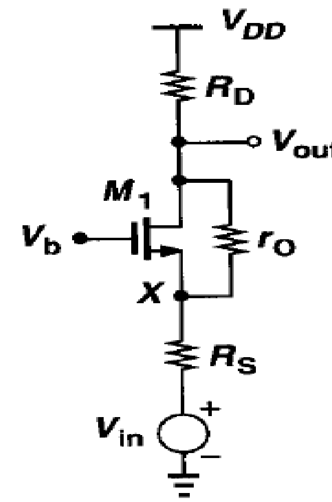
$$\text{共栅}CG: R_{out} = \frac{V_X}{I_X} = \{[1 + (g_m + g_{mb})r_o]R_S + r_o\} \,||\, R_D$$

例3.13 Calculate V_{out}/I_{in} and the output impedance



已知右图，有式（3.104）：$A_V = \frac{V_{out}}{V_{in}}$

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



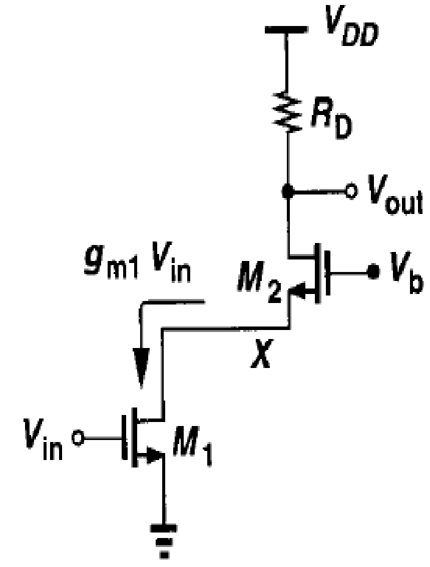
(a)

$$\therefore \frac{V_{out}}{I_{in}} = \frac{V_{out}}{V_{in}} \frac{V_{in}}{I_{in}} = \frac{r_o(g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb})r_o R_P + R_P + R_D} R_D \times R_P \qquad (3.114)$$

$$R_{out} = \{[1 + (g_m + g_{mb})r_o]R_P + r_o\} \parallel R_D$$

3.5 Cascode Stage(共源共栅)

输入器件M1将Vin转换成漏电流。M2什么作用？
M2增大了从Vout向下看的输出阻抗！

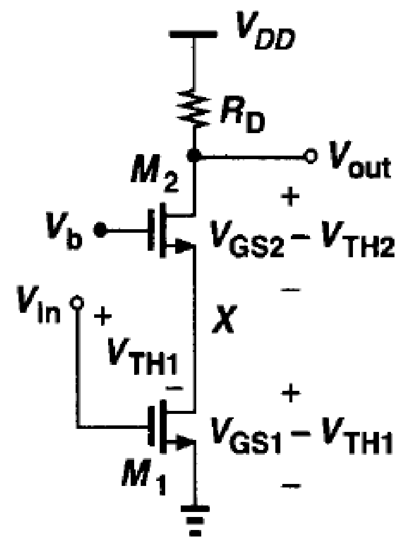


设计偏置Vb，保证M1和M2（cascode器件）工作在饱和区：

$$V_X = V_b - V_{GS2}，且 V_X \geq V_{in} - V_{TH1}$$

$$即\quad V_b - V_{GS2} \geq V_{in} - V_{TH1}$$

$$\therefore\quad V_b \geq V_{in} - V_{TH1} + V_{GS2}$$

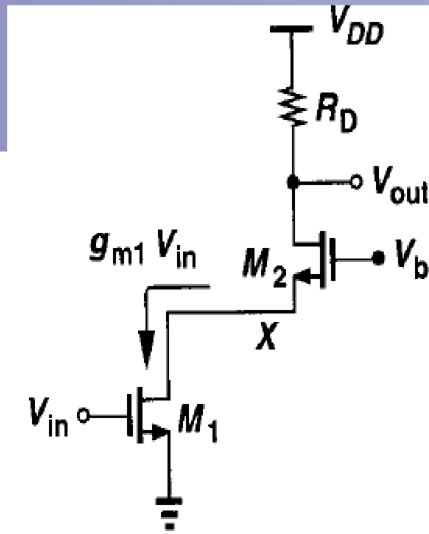


$$\therefore V_{out} \geq V_b - V_{TH2} \geq V_{in} - V_{TH1} + V_{GS2} - V_{TH2}$$

即 $V_{out} \geq M1$过驱动电压 + M2过驱动电压

直筒式CASCODE.
输入器件和cascode器件是同一类型MOS

分析：Vin从0=>VDD



(1) $V_{in} \le V_{TH1}$，M_1和M_2截止，$V_{out} = V_{DD}$

$I_D = 0, \quad V_{GS2} - V_{TH} = 0$

$\therefore \quad V_X = V_b - V_{TH2}$

(2) $V_{in} \ge V_{TH1}$，当$V_{in} \uparrow \Rightarrow I_D \uparrow \Rightarrow V_{out} \downarrow \& V_{GS2} \uparrow \Rightarrow V_X \downarrow$

M_1和M_2在饱和区时：$A_V = -g_m R_D$，与CS相同

$$I_D = \frac{\mu_n C_{ox}}{2}\frac{W}{L}(V_{GS} - V_{TH})^2$$

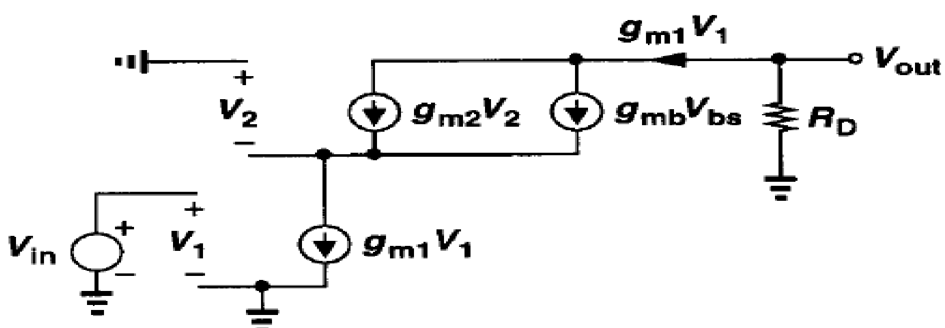
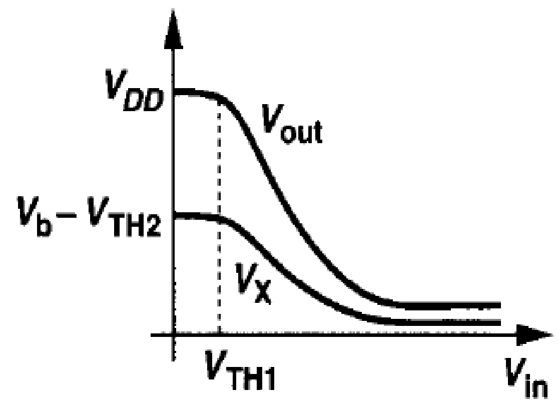


Figure 3.53 Small-signal equivalent circuit of cascode stage.



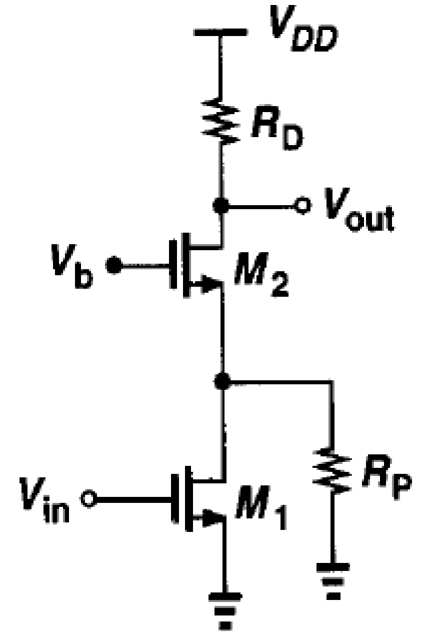
(3)Vin增加到很大时，M1或M 2 进入线性区。
Vb比较小时，M1先进入线性区。

M_2进入线性区时$V_{out} \approx V_X$

例3.14 calculate the voltage gain

M_1的小信号漏电流（受控电流源）$g_{m1}V_{in}$，被R_p（例如r_{o1}，或节点寄生电容）和向M2源极看进去的阻抗$1/(g_{m2}+g_{mb2})$分流。

$$小信号 I_{D2}=g_{m1}V_{in}\frac{R_P}{\frac{1}{g_{m2}+g_{mb2}}+R_P}=g_{m1}V_{in}\frac{(g_{m2}+g_{mb2})R_P}{1+(g_{m2}+g_{mb2})R_P}$$



$$A_V=\frac{-I_{D2}(R_D\,||\,R_{out})}{V_{in}}\approx\frac{-I_{D2}R_D}{V_{in}}$$

$$=-g_{m1}\frac{(g_{m2}+g_{mb2})R_PR_D}{1+(g_{m2}+g_{mb2})R_P},$$

V_{out}向下看的输出阻抗R_{out}很大

cascode结构的重要特性：输出阻抗大

式（3. 60）：

$$R_{out} = [1 + (g_{m2} + g_{mb2})r_{o2}]r_{o1} + r_{o2}$$

$$\approx (g_{m2} + g_{mb2})r_{o2}r_{o1}$$

将共源M_1输出电阻r_{o1}提高了$(g_{m2} + g_{mb2})r_{o2}$

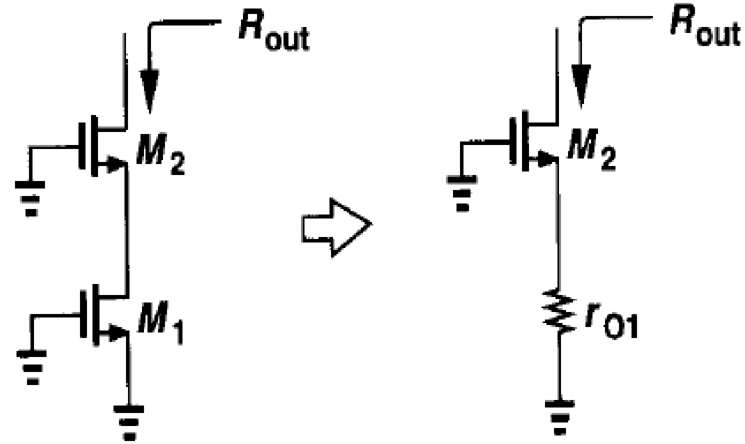
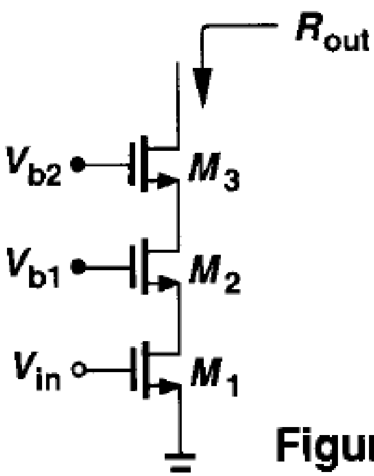


Figure 3.55 Calculation of output resistance of cascode stage.



将共源M_1输出电阻r_{o1}提高

$(g_{m3} + g_{mb3})r_{o3}(g_{m2} + g_{mb2})r_{o2}$

但现代工艺电源电压很低，此结构难以应用

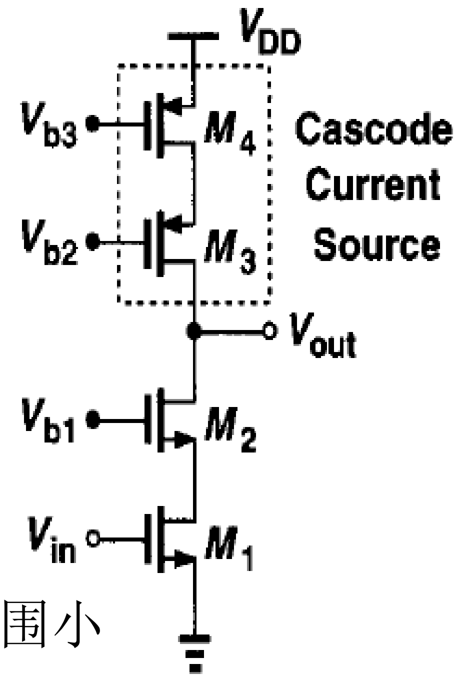
Figure 3.56 Triple cascode.

放大电路要上下阻抗对称

$$R_{out} = \{[1+(g_{m2}+g_{mb2})r_{o2}]r_{o1}+r_{o2}\} \,||\, \{[1+(g_{m3}+g_{mb3})r_{o3}]r_{o4}+r_{o3}\}$$

$$\approx (g_{m2}+g_{mb2})r_{o2}r_{o1} \,||\, (g_{m3}+g_{mb3})r_{o3}r_{o4} \approx g_{m2}r_{o2}r_{o1} \,||\, g_{m3}r_{o3}r_{o4}$$

$$A_V \approx -g_{m1}R_{out} = -g_{m1}[g_{m2}r_{o2}r_{o1} \,||\, g_{m3}r_{o3}r_{o4}]$$



直流输入电压动态范围小

Figure 3.60 NMOS cascode amplifier with PMOS cascode load.

Shielding property

设M_1M_2相同，M_3M_4相同，都是饱和区。

图（a）：$V_X - V_Y = \Delta V$

$$I_{D1} - I_{D2} = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{TH})^2(\lambda V_{DS1} - \lambda V_{DS2})$$

$$= \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{TH})^2 \lambda \Delta V_{XY}$$

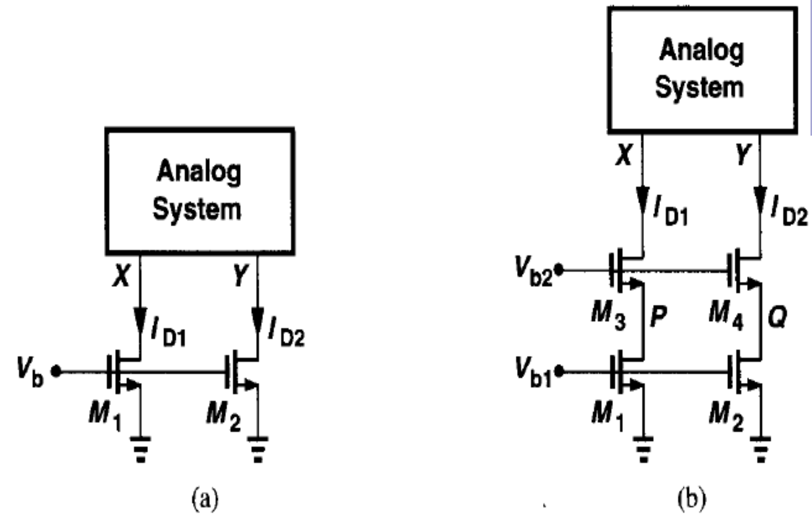


Figure 3.61

图（b）：$\Delta V_{PQ} = \Delta V_{XY}\dfrac{r_{o1}}{R_{Xout}} = \Delta V_{XY}\dfrac{r_{o1}}{[1+(g_{m3}+g_{mb3})r_{o3}]r_{o1}+r_{o3}}$

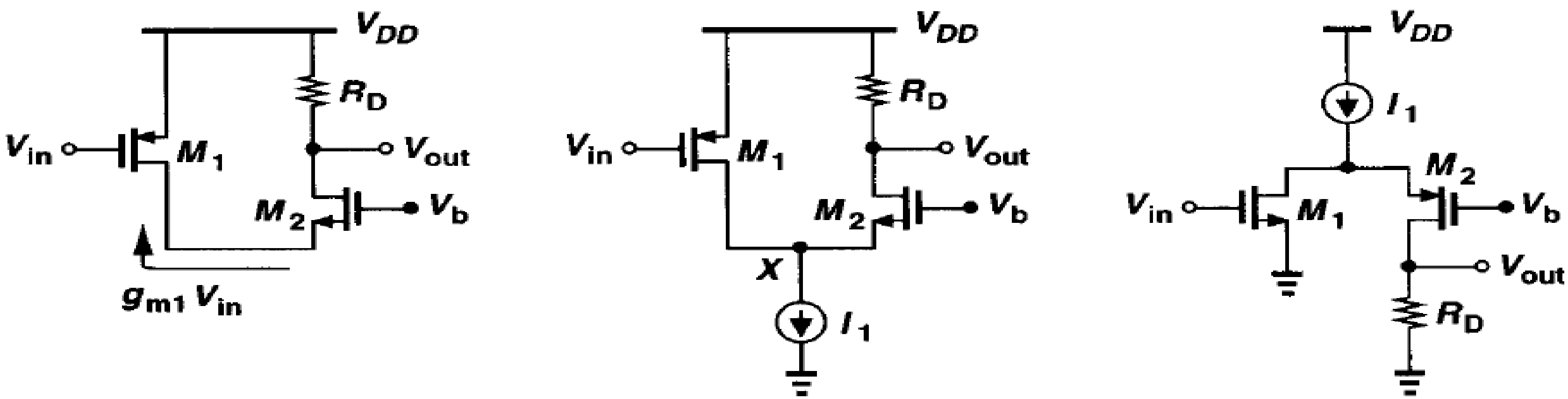
$\approx \Delta V_{XY}\dfrac{1}{(g_{m3}+g_{mb3})r_{o3}}$，PQ变化小，即$M_3$屏蔽了$X$点对$P$点的影响。

$$I_{D1} - I_{D2} = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{TH})^2 \lambda \Delta V_{PQ}$$

$$= \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{TH})^2 \lambda \Delta V_{XY}\frac{1}{(g_{m3}+g_{mb3})r_{o3}}$$

CASCODE结构使电流失配大大减少。

折叠式共源共栅 Folded cascode

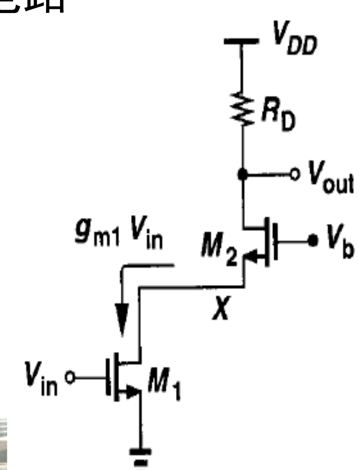


(a) (b) (c)

图3.63 (a) PMOS和NMOS组成折叠CASCODE（小信号电流回路）
(b)有直流通路的 PMOS作输入器件的折叠式共源共栅电路
(c)有直流通路的NMOS作输入器件的折叠式共源共栅电路

输入器件和cascode器件不是同一类型。
由于I1电流源，若ID1增大则ID2减小。

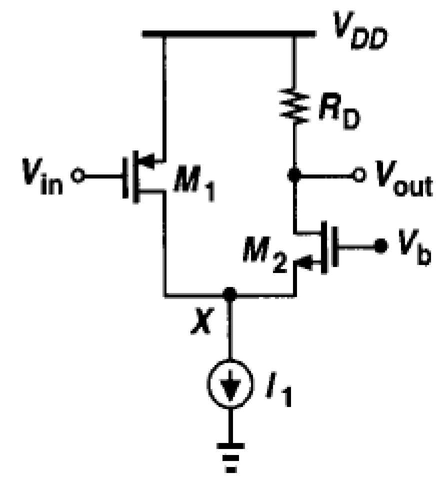
输入电压范围比直筒式cascode大。



Folded cascade 大信号分析

（1） $V_{in} > V_{DD} - |V_{TH1}|$，$M_1$截止。电流源$I_1$全部流过$R_D$，$V_{out}$最低。$V_{out} = V_{DD} - I_1 R_D$

I1RD不能过大，否则M2易进入线性区



(b)

（2） $V_{in} < V_{DD} - |V_{TH1}|$时，$M_1$工作在饱和区。

$$I_{D2} = I_1 - \frac{1}{2}\mu_p C_{ox}\left(\frac{W}{L}\right)_1 (V_{DD} - V_{in} - |V_{TH1}|)^2$$

$$V_{out} = V_{DD} - I_{D2} R_D$$

$$= V_{DD} - I_1 R_D - \frac{1}{2}\mu_p C_{ox}\left(\frac{W}{L}\right)_1 (V_{DD} - V_{in} - |V_{TH1}|)^2 R_D$$

$$I_{D1} + I_{D2} = I_1$$

当V_{in}下降使$I_{D2} = 0$时，$V_{out} = V_{DD}$

$$I_1 - \frac{1}{2}\mu_p C_{ox}\left(\frac{W}{L}\right)_1 (V_{DD} - V_{in1} - |V_{TH1}|)^2 = 0$$

最小输入$V_{in1} = V_{DD} - \sqrt{\dfrac{2I_1}{\mu_p C_{ox}\left(\frac{W}{L}\right)_1}} - |V_{TH1}|$

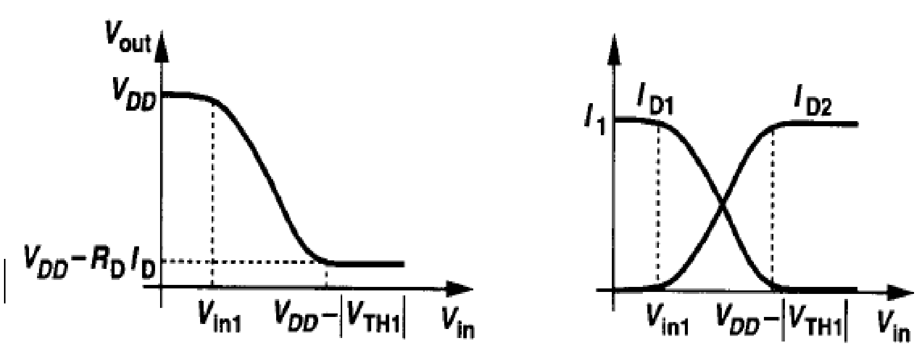


Figure 3.64 Large-signal characteristics of folded cascode.

折叠管的输出阻抗

Calculate the output impedance of the folded cascode shown in Fig. 3.65 where M_3 operates as a current source.

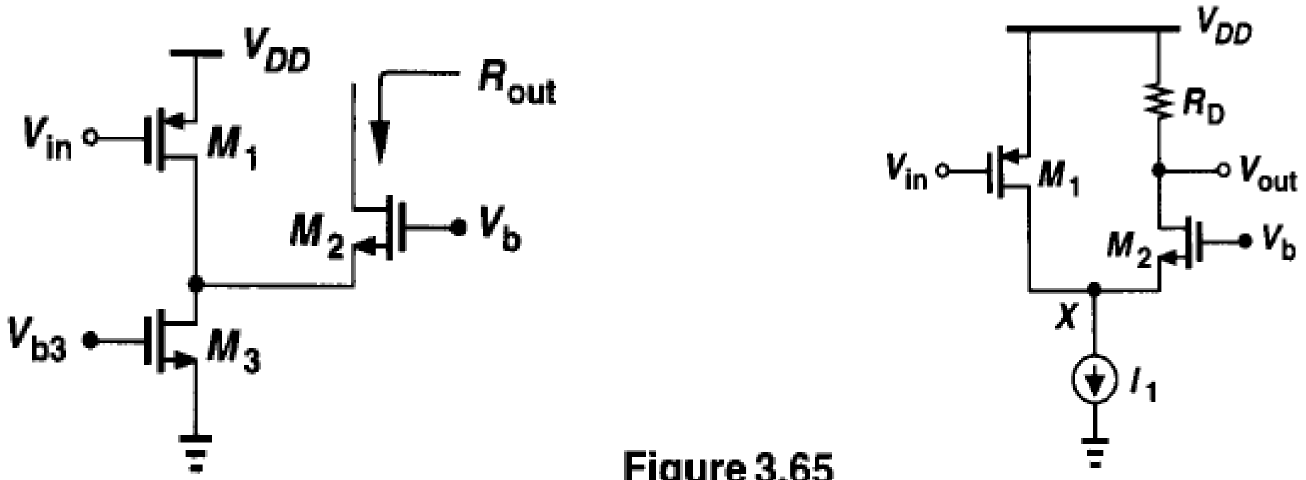


Figure 3.65

(b)

Solution

Using (3.60), we have

$$R_{out} = [1 + (g_{m2} + g_{mb2})r_{O2}](r_{O1} \| r_{O3}) + r_{O2}. \quad (3.136)$$

Thus, the circuit exhibits an output impedance lower than that of a nonfolded cascode.

The load of a folded cascode

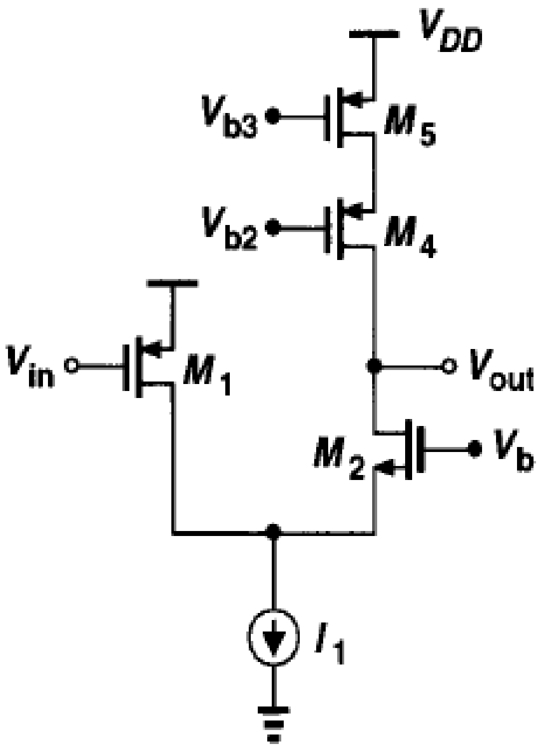


Figure 3.66 Folded cascode with cascode load.

记$M3$组成尾电流源：I_1

$$R_{out} = \{[1+(g_{m2}+g_{mb2})r_{o2}](r_{o1}\,||\,r_{o3})+r_{o2}\}$$
$$||\,\{[1+g_{m4}r_{o4}]r_{o5}+r_{o4}\}$$
$$\approx (g_{m2}+g_{mb2})r_{o2}(r_{o1}\,||\,r_{o3})\,||\,g_{m4}r_{o4}r_{o5}$$

$$A_v = -g_{m1}R_{out}$$
$$= -g_{m1}【(g_{m2}+g_{mb2})r_{o2}(r_{o1}\,||\,r_{o3})\,||\,g_{m4}r_{o4}r_{o5}】$$

本章知识要点

- 共 源 CS 、 源 跟 随 器 SF 、 CASCODE的低频增益和输出电阻
- 共栅CG的低频增益、输入与输出电阻
- MOS二极管的等效电阻
- CS各种负载情况的优缺点