Chapter 3

单级放大器
Single-Stage Amplifiers

中科大微电子学院

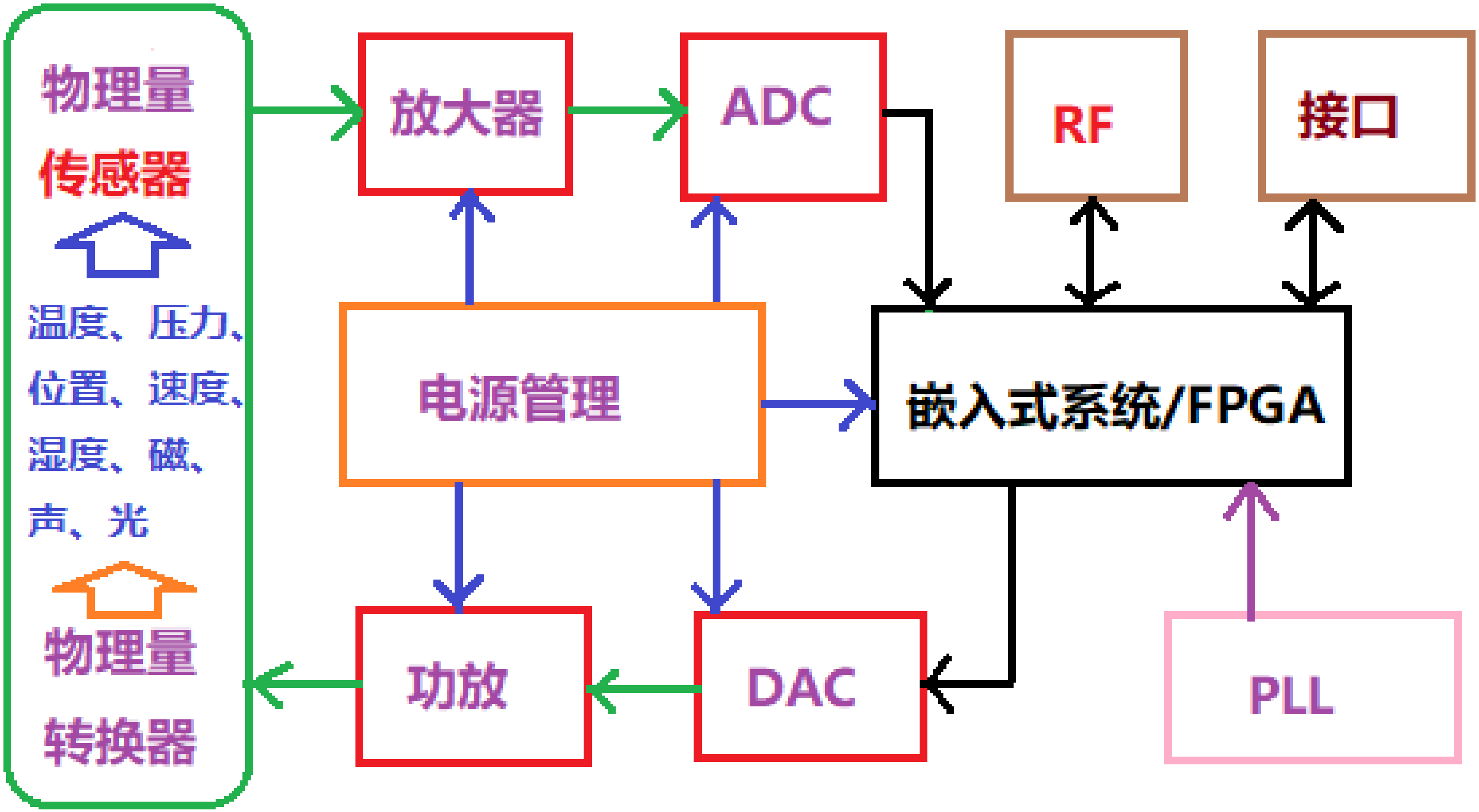
黄鲁、程林

教材：拉扎维《模拟CMOS集成电路设计》

第3章内容 contents

3.1 放大器应用例

信号处理链路中几乎所有电路模块都包含不同性能的放大器：



应用例：射频前端中的放大器

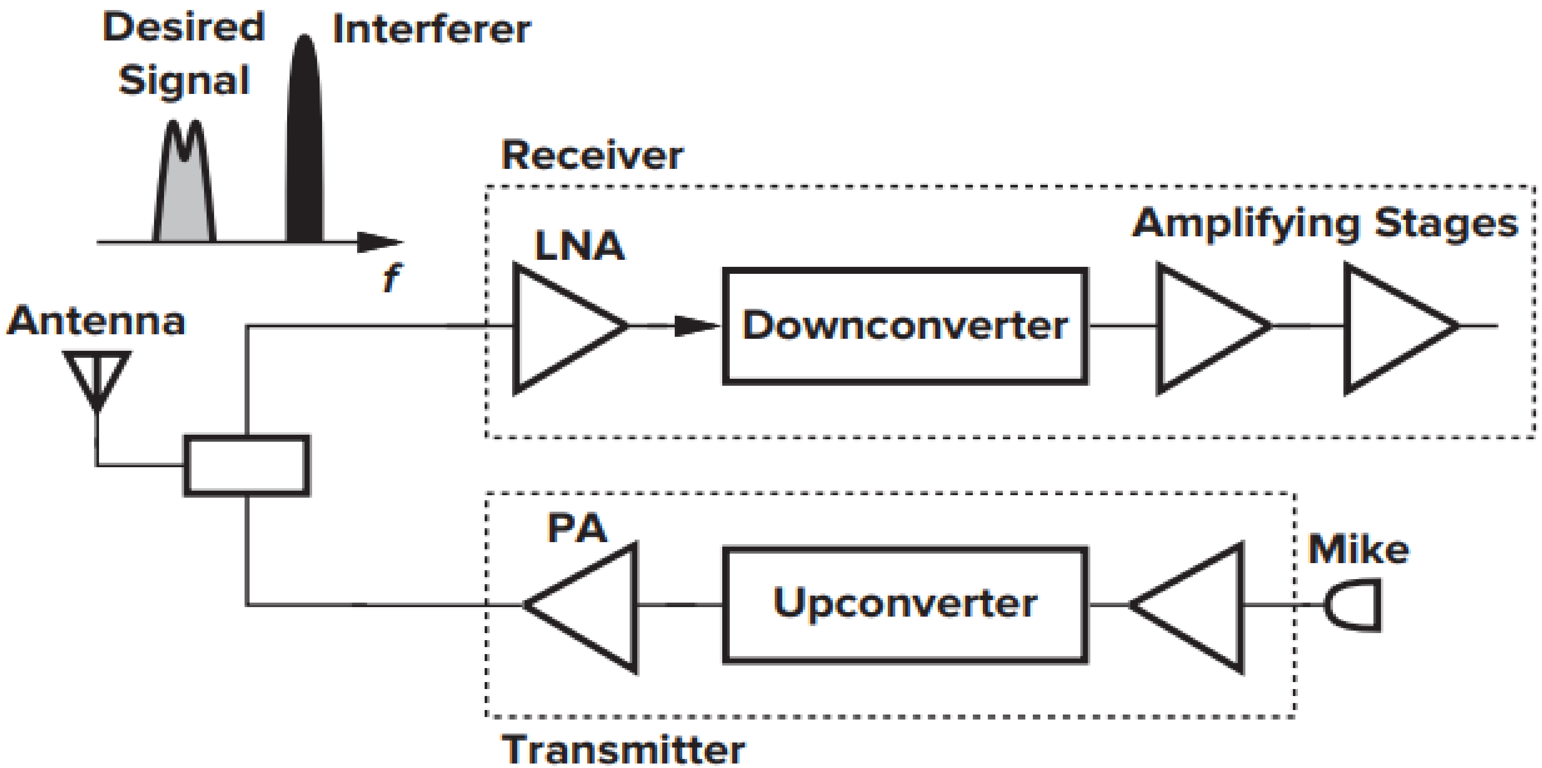


Figure 3.1 General RF transceiver.

各个放大器的中心频率、带宽、增益、线性度等指标要求不同，对应实现的电路结构不同。

没有设计指标就没有电路（结构与器件）设计！

例：集总参数还是分布参数电路?

本课程学习知识仅适用于集总参数电路，即器件（包括线路）长度远小于信号波长。

线长l $<<$ 0.1λ，工程上l $<$ 0.05λ 视为集总系统。

其中$\lambda = c/f \approx 3\times10^{10}cm/f$，芯片内实际波长 $\lambda/\sqrt{\varepsilon_r} \approx \lambda/2$

如何判断是否应采用射频传输线?

例（1）：2.4GHz信号，波长约12.5cm，芯片内部电路和连线尺寸不超过3mm(一般均满足)时，按集总系统设计。

例（2）：10GHz信号（波长3cm）及更高频率，芯片内部电路（模块间连线距离大750um时）一般需要按分布参数设计。

分布参数模拟/射频电路进行阻抗匹配，最重要目的是进行无反射的信号传输，避免信号失真，其次是功率传输或滤波。

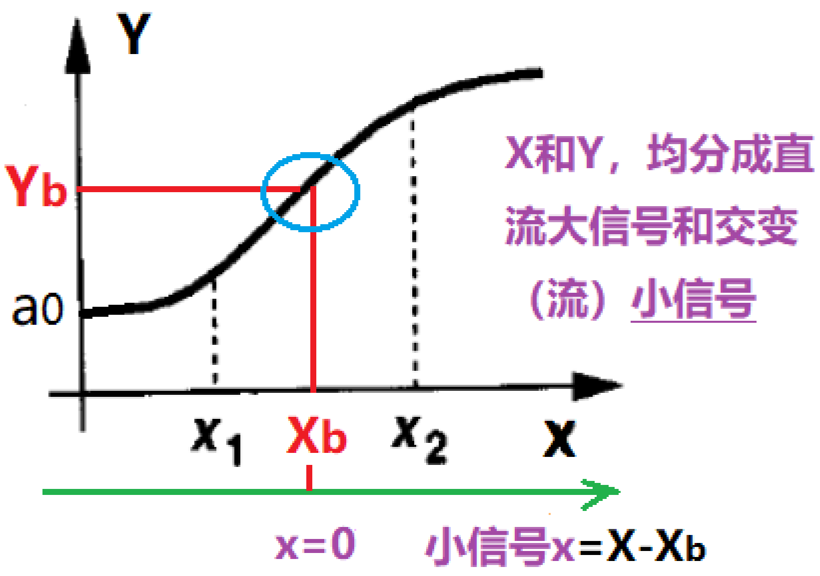
3.2 模拟电路设计基本概念：直流工作点

放大器输入X和输出Y:

$$Y(t) = a_0 + a_1 X(t) + a_2 X^2(t) + \cdots + a_n X^n(t), \qquad X_1 \le X \le X_2$$

当$X_2 \sim X_1$在X_b附近变化时，记x=X−X_b，$Y(t) \approx Y_b + a_1 x(t)$

$a_1 x(t) << Y_b$，直流偏置点为X_b和Y_b



模拟电路前端设计2类主要问题：

（1）直流工作点涉及：

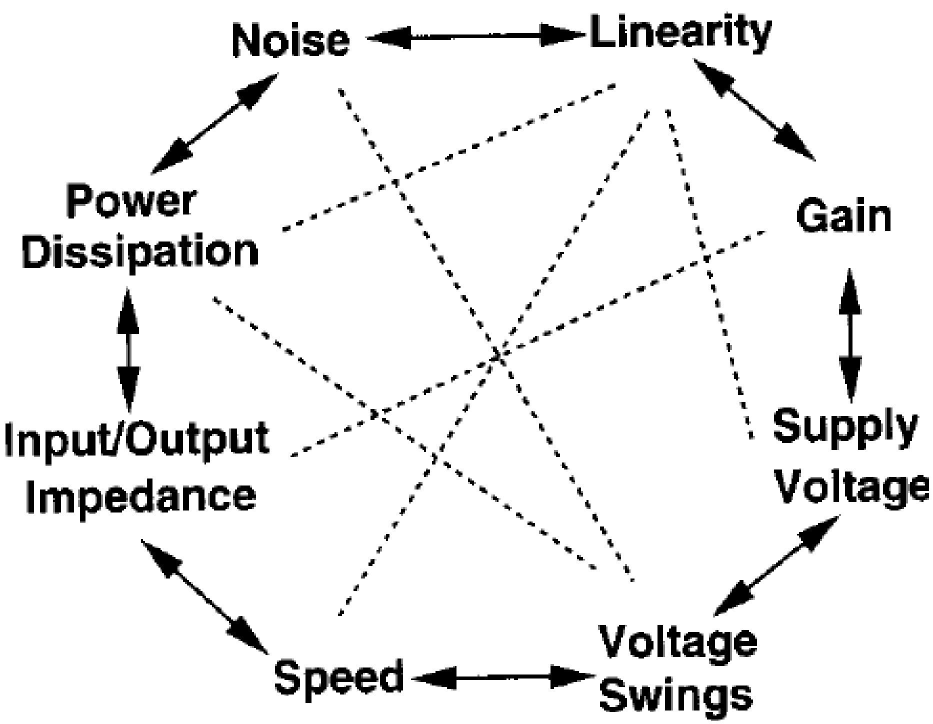
直流输入、直流输出电压或电流（工作点），最大与最小值（一般指电压摆幅，或线性度），功耗；

（2）交流小信号涉及：

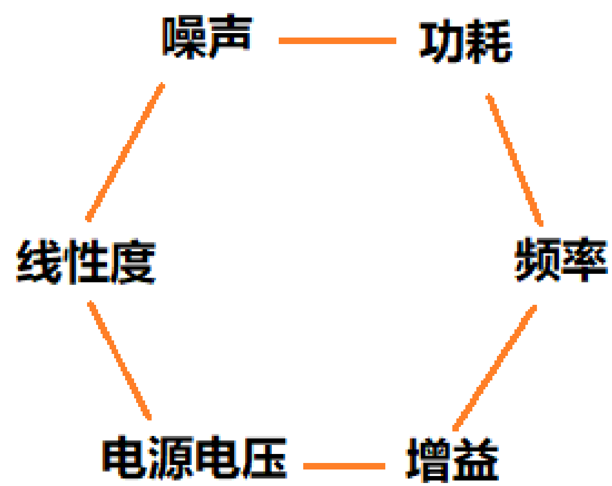
增益（传递函数与零极点）、带宽（速度）、输入输出阻抗，噪声（信噪比）、稳定性。

模拟电路设计基本概念：折中

Analog circuits design octagon：**trade-off**



比较：射频设计六边形原则



比模拟电路设计缺少：
*电压摆幅：与工作点有关，射频易设工作点
*输入输出阻抗：真正的射频电路一般是50欧

*满足重要性能指标（与应用系统有关）；
*一般地，频率（速度）指标最重要；
*负反馈系统应确保稳定性（不属于八边形原则）；
*任何指标提升必有代价，适用即可。

- 3.3.1 Common-source stage with resistive load

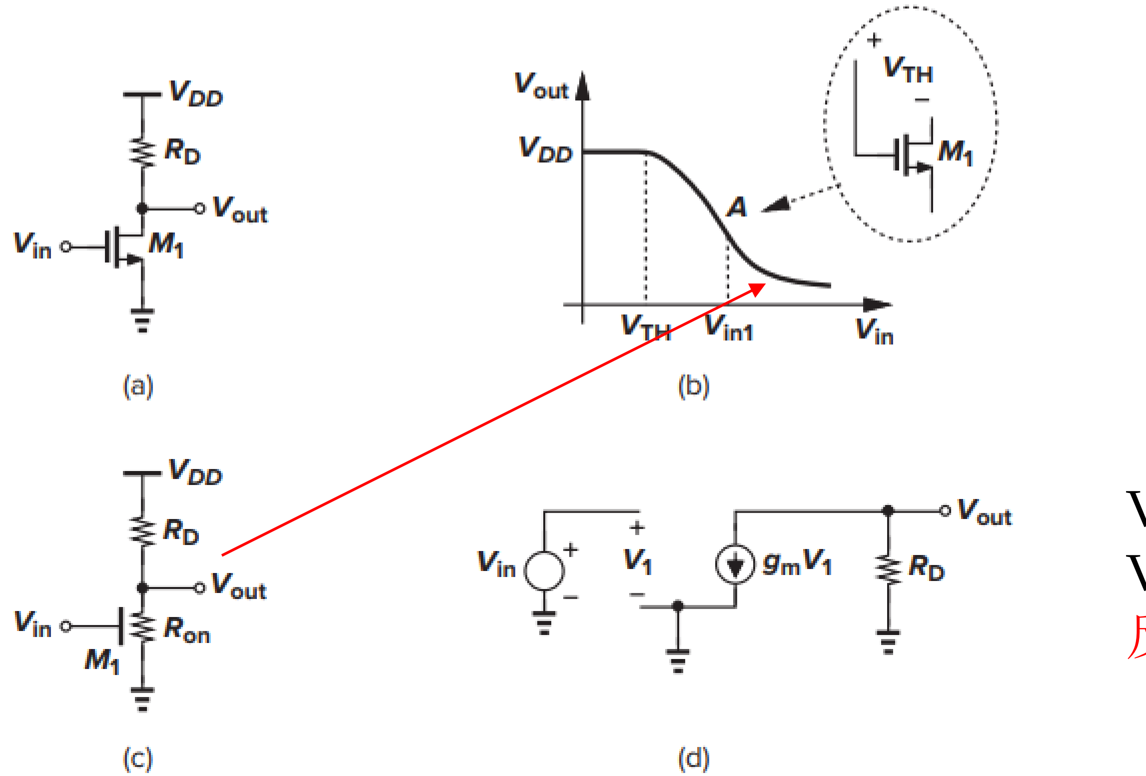


Figure 3.4 (a) Common-source stage, (b) input-output characteristic, (c) equivalent circuit in the deep triode region, and (d) small-signal model for the saturation region.

饱和与线性区

临界点A点：

当Vin >Vin1（A点）

则Vin - V_{TH} >Vout

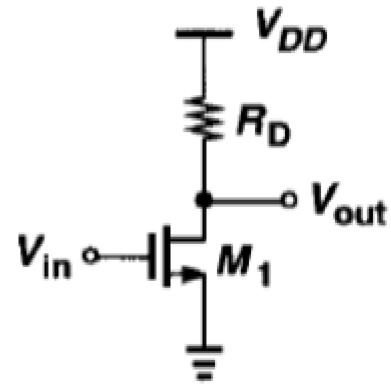
triode region

Vin增加->I_D增加->V_{RD}增加->Vout下降。反向放大

(d)图为小信号模型：交变信号（无直流）幅度小；R_D远小于ro；忽略寄生电容，仅限低频。

1.采用大信号分析方法：应用IV公式

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



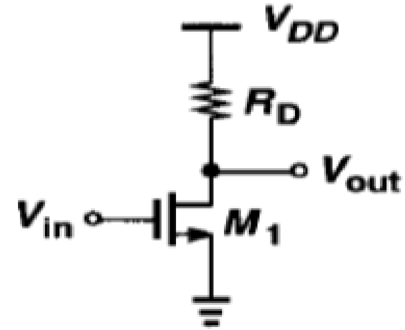
$$V_{out} = V_{DD} - R_D I_{Dn} = V_{DD} - R_D \frac{1}{2}\mu_n C_{OX} \frac{W}{L'}(V_{GS} - V_{Tn})^2$$

提示：电阻负载一般阻值较小，几k~几十 k欧，因此手算时可忽略ro(沟道长度调制效应引起，模拟电路一般在几十k~几百k欧数量级)，并取L’=Leff=Ldrawn-2LD， 这里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 .

大（总）信号在饱和区与线性区的输出

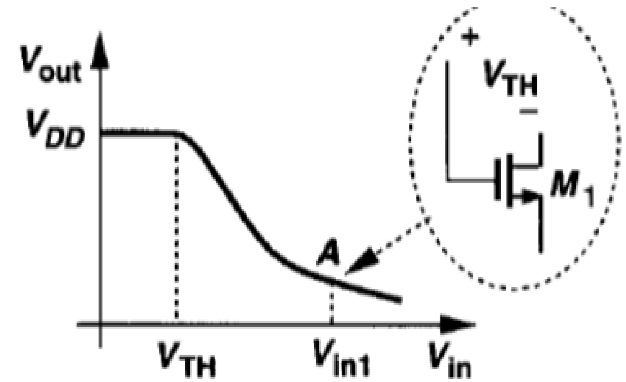
（1）饱和区：临界点A输入信号Vin1



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

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

解出 V_{in1}和对应的V_{outA}。手算时$\lambda V_{DS} \approx 0$



（2）线性区：

$V_{in} > V_{in1}$, 即$V_{in} > V_{out} + V_{TH}$

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

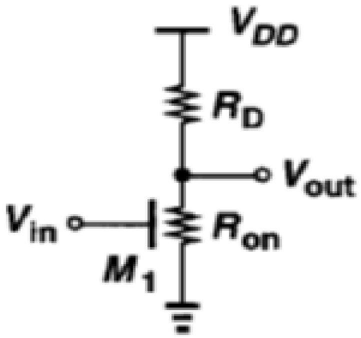
$$= V_{DD} - R_D \times \frac{1}{2} \mu_n C_{OX} \frac{W}{L_{eff}} \left[2(V_{in1} - V_{TH}) V_{out} - V_{out}^2 \right]$$

三极管区仍能工作，
但跨导小，且与VDS有关

大(总)信号在深线性区的输出电压

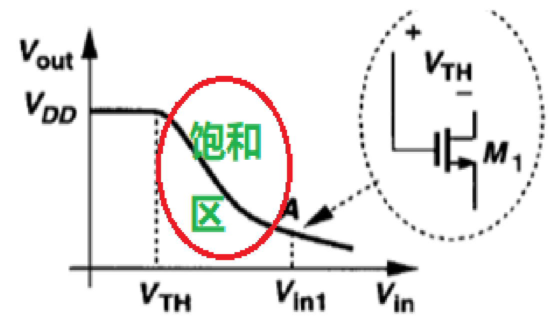
$$V_{DS} = V_{out} << 2(V_{GS} - V_{TH})\text{时}$$

$$R_{DS} = R_{on} = \frac{1}{\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})}$$

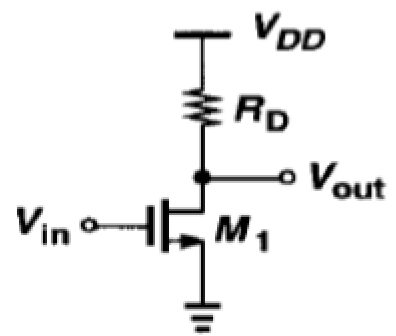


$$V_{out} = \frac{R_{on}}{R_{on} + R_D} V_{DD}$$

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



模拟电路一般应工作在饱和区，$V_{TH}<V_{in}<V_{in1}$, $V_{outA}<V_{out}<V_{DD}$,
避免工作在线性区（跨导与V_{DS}有关，即与变化较大的输出有关）。



饱和区small-signal gain of common source

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

忽略沟道长度调制效应

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

基本概念：

（1）放大倍数（增益）仅与交变信号相关，不包括直流信号电平（工作点）；

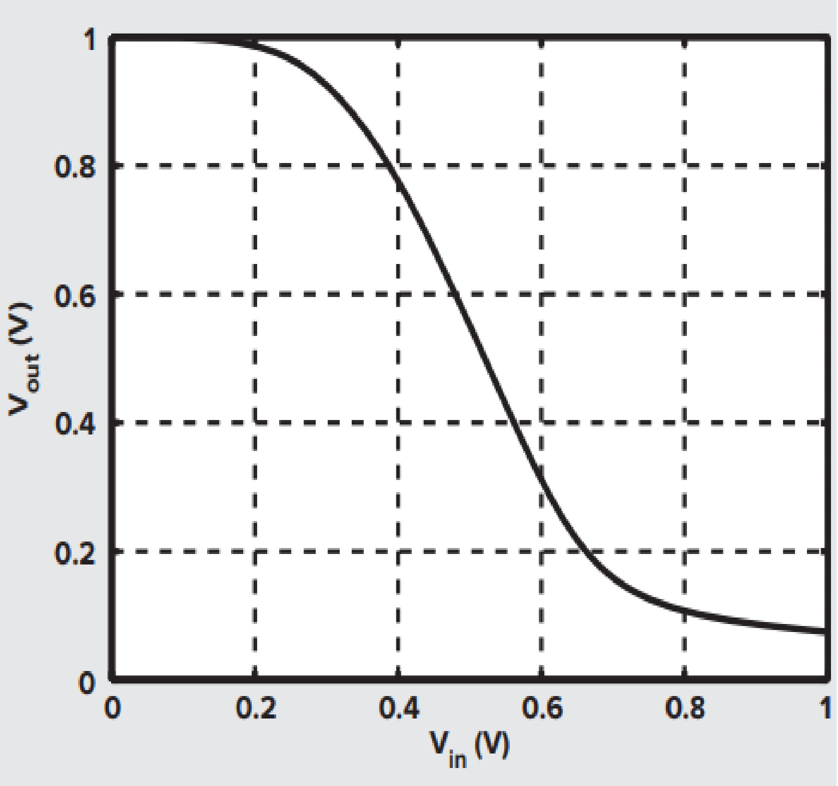
（2）CS负增益表示输入与输出中的交变信号反向；

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

（4）本电路中，因gm与Vin (工作点+变化量)有关，故输入变化较大时，增益不恒定，导致非线性失真。

从DC（直流）VTC仿真图获得gain

How does the CS stage behave in nanometer technologies? The figure plots the simulated input-output characteristic for $W/L = 2\ \mu m/40\ nm$, $R_D = 2\ k\Omega$, and $V_{DD} = 1\ V$. We observe that the circuit provides a gain of about 3 in the input range of 0.4 V to 0.6 V. The output swing is limited to about 0.3 V–0.8 V for the gain not to drop significantly.



VTC,电压传输特性

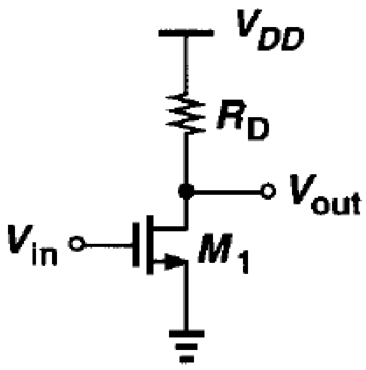
$$A_V = \frac{\Delta V_{out}}{\Delta V_{in}} \approx \frac{0.8-0.3}{0.6-0.4} = 2.5$$

Example 3.1：I_D和g_m随输入的变化

Solution:

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

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



Vout由RD压降确定

饱和与线性区临界点输入Vin1=Vout+VTn

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

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

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

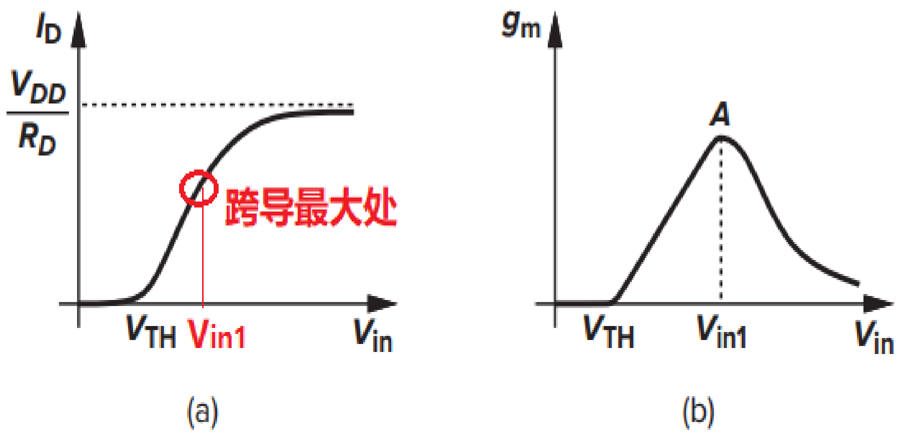


Figure 3.5

沟道长度调制效应对增益的影响

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

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

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

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

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

$$\because \lambda I_D = \frac{1}{r_o} \qquad \Rightarrow A_V = -\frac{g_m R_D}{1 + \frac{1}{r_o}R_D} = -\frac{g_m R_D r_o}{r_o + R_D} = -g_m\left(R_D \,||\, r_o\right)$$

沟道长度调制效应(ro=rds)使gain减小

2.采用交变小信号模型计算方法

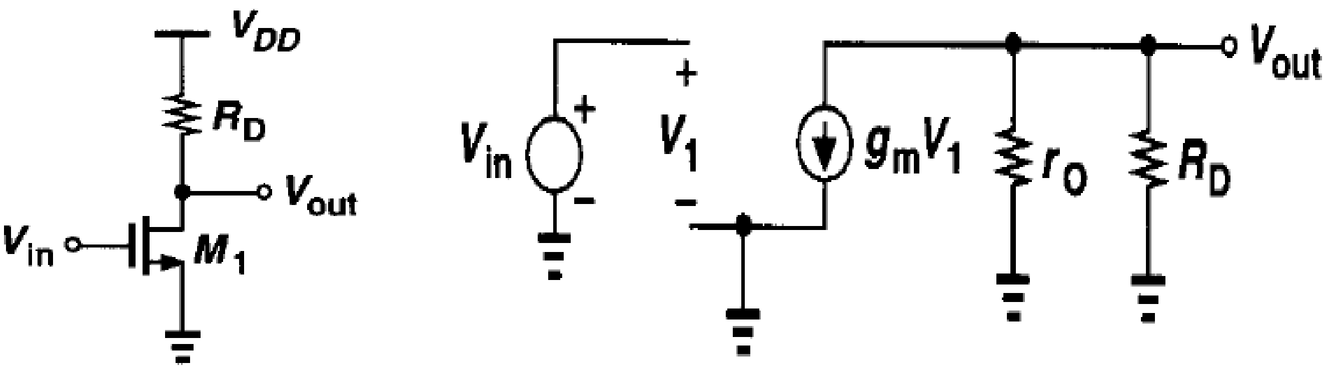


Figure 3.7 Small-signal model of CS stage including the transistor output resistance.

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.

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

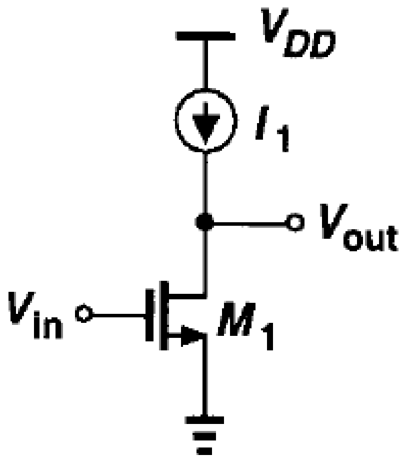
Vin和Vout是交变小信号

$$A_v = \frac{V_{out}}{V_{in}} = -g_m (r_o || R_D)$$

采用小信号模型，计算简捷，结果与大信号方法相同

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**变化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$g_m r_o$：MOS管的增益极限



理想电流源具有无穷大输出阻抗。

即电流源负载等效于$R_D = \infty$

$$A_v = -g_m(r_o||R_D) = -g_m r_o$$

MOS极限电压增益为$g_m r_o$，称为本征增益

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

这里的所谓"负载"是指交流上拉（或下拉）电阻，电路输出阻抗与此有关，并非电路真实驱动的负载。

重要推论：应gmro >>1， thus usually 1/gm<<ro

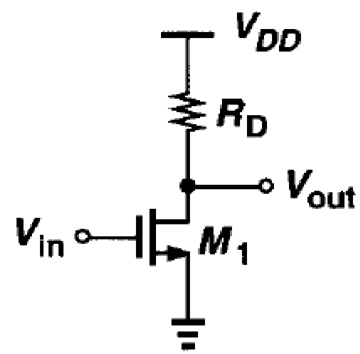
亚微米工艺gmro约几十~几百，纳米工艺gmro约几~几十。

例：CS性能参数的数量级概念

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

$V_{DD} = 1.8V$，$V_{in,\ DC} = 0.7V$，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$LD \approx 0.1\mu m$，计算$\dfrac{W}{L_{eff}}, R_D, g_m r_o$，$A_v$。



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

$$100 \times 10^{-6} A = \frac{1}{2} \times 300\ cm^2/(V \cdot s) \times 10 \times 10^{-7}\ F/cm^2 \times \frac{W}{L_{eff}} \times 0.2^2 V^2 \times 1.1$$

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

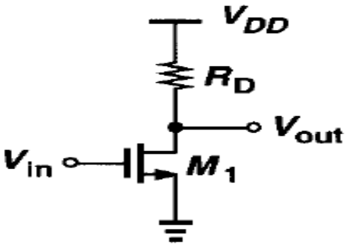
$$R_D = \frac{V_{DD} - V_{out}}{I_D} = \frac{1.8V - 1V}{100 \times 10^{-6} A} = 8k\Omega,\quad r_o = \frac{1}{\lambda I_D} = \frac{1}{0.1/V \times 100 \times 10^{-6} A} = 100k\Omega$$

$$g_m = \mu_n C_{ox}(W / L_{eff})(V_{in} - V_{Tn})(1 + \lambda V_{out}) = 300 \times 10 \times 10^{-7} \times \frac{4.5}{0.3} \times 0.2 \times 1.1 \approx 1\frac{mA}{V}$$

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

饱和区动态范围（用大信号公式计算）

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



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

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

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

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

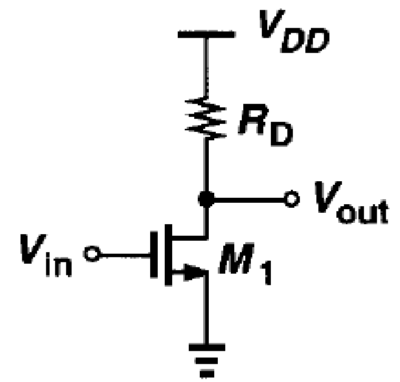
$(\because Q=CV=It$, 故量纲$FV=As$, $\therefore \Omega\bullet V\bullet F/s=\Omega A=V)$

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

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

最小输出电平的近似计算：在增益较大时$V_{out,min}\approx V_{in}-V_{Tn}$

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$\lambda_n = 0.1/V$，$V_{out} = 1V$，改变$I_D = 20\mu A$，

采用0.5μm工艺$LD \approx 0.1\mu m$，重新计算$\dfrac{W}{L_{eff}}$，R_D，$g_m r_o$，A_v。

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

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

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

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

$$g_m = \mu_n C_{ox}(W/L_{eff})(V_{in} - V_{Tn})(1 + \lambda V_{out}) = 300\times10\times10^{-7}\times\frac{0.9}{0.3}\times0.2\times1.1 \approx 0.2\frac{mA}{V}$$

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

本例如何提高增益？增大W/L！即gm。为维持电流不变，需相应地降低直流Vin

R_D电阻负载CS放大器的设计注意事项

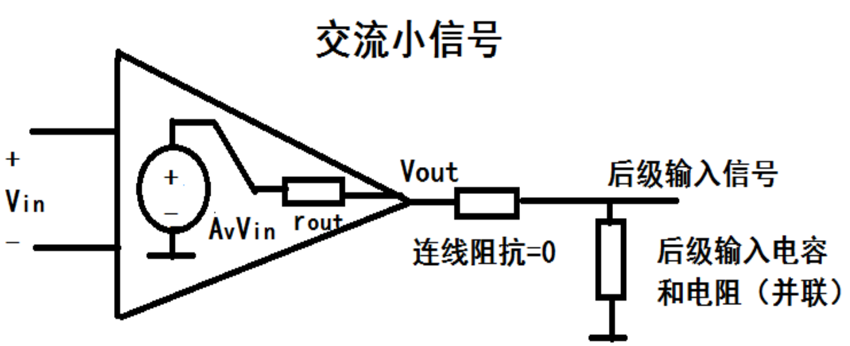
- 增益不大。原因：R_D受到直流电流和工作点电压限制，不能很大，如在高频（需大电流）和大动态范围（直流输出电平为1/2电源）时。
- 大幅度交变输入输出信号时，应注意V_{DS}避免进入线性区。
- 实际设计中应注意考虑前后级电路，避免不合适的工作点。
- 增益等参数仿真时须有后级电路的交流等效输入阻抗做为本级load。

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？

放大级大好（增益大），输出级小好



电压放大器

实际电阻（例smic18mmrf工艺）

电阻名称r开头

(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**，大电阻的面积大，有寄生电容。

电阻值不准确，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。

3.3.2 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。

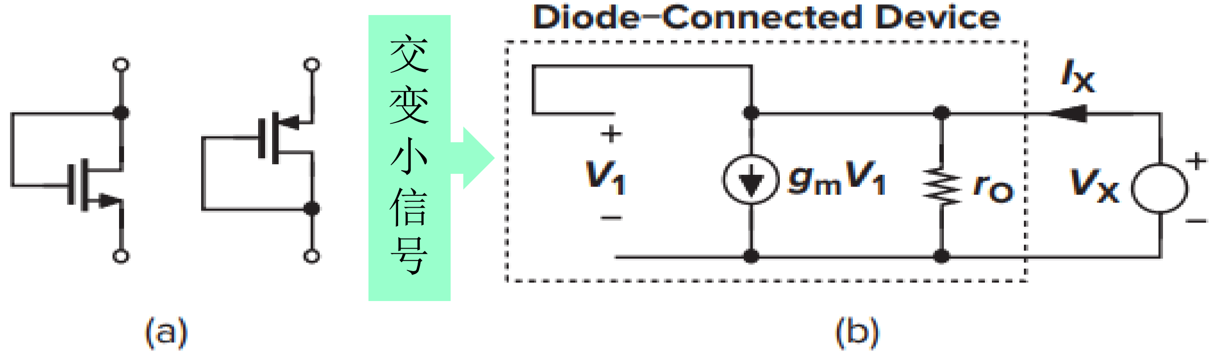


Figure 3.10 (a) Diode-connected NMOS and PMOS devices; (b) small-signal equivalent circuit.

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

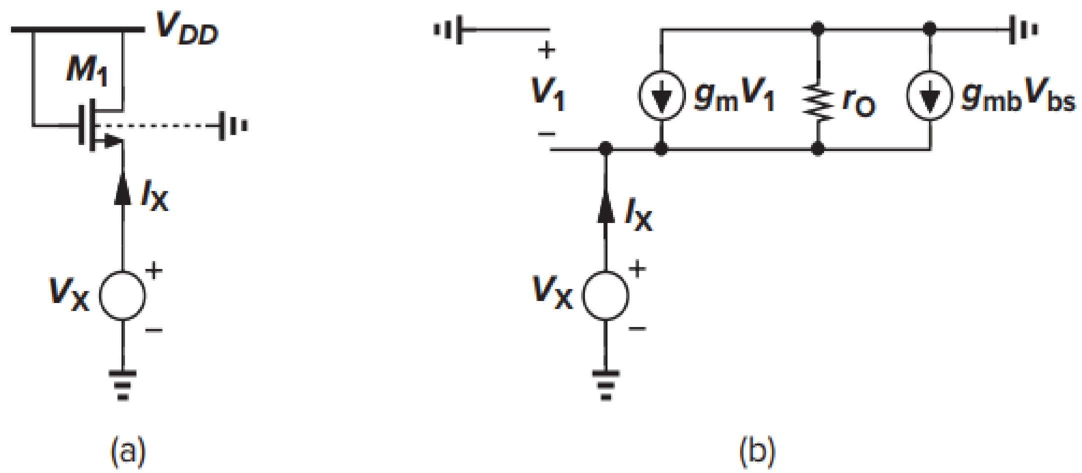
$V_1=V_X$

交变信号 $I_X = \dfrac{V_X}{r_0} + g_m V_X$

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

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



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

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

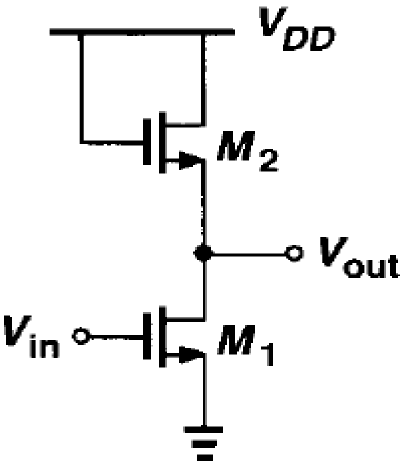
二极管输出电阻：

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

$$\approx \frac{1}{g_m} \,||\, \frac{1}{g_{mb}} = \frac{1}{g_m + g_{mb}}$$

体效应使rout减小，导致增益减小。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(3.28)**

由第2章

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

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

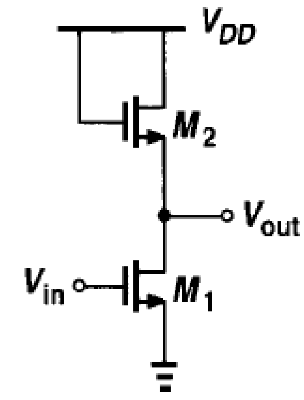
$$\longrightarrow A_v = -\frac{\sqrt{2\mu_n C_{ox}(W/L)_1 I_{D1}}}{\sqrt{2\mu_n C_{ox}(W/L)_2 I_{D2}}}\frac{1}{1+\eta}$$

since $I_{D1} = I_{D2}$ $$A_v = -\sqrt{\frac{(W/L)_1}{(W/L)_2}}\frac{1}{1+\eta}$$

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

Neglecting channel-length modulation for simplicity.



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

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

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

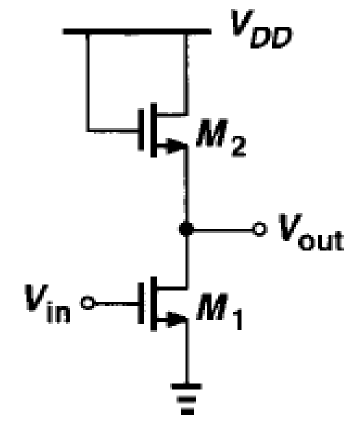
$\because \frac{\partial V_{TH2}}{\partial V_{in}}=\frac{\partial V_{out}}{\partial V_{in}}\frac{\partial V_{TH2}}{\partial V_{out}}=\frac{\partial V_{out}}{\partial V_{in}}\frac{\partial V_{TH2}}{\partial V_{SB2}}=A_v\eta_2$ $\therefore \sqrt{\left(\frac{W}{L}\right)_1}=-\sqrt{\left(\frac{W}{L}\right)_2}(1+\eta_2)A_v$

$\rightarrow$ $A_v=-\sqrt{\frac{(W/L)_1}{(W/L)_2}}\times\frac{1}{1+\eta_2}$，与小信号分析相同。 $\eta_2=\frac{g_{mb2}}{g_{m2}}$

Vout与Vin成线性特性

直流工作点

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



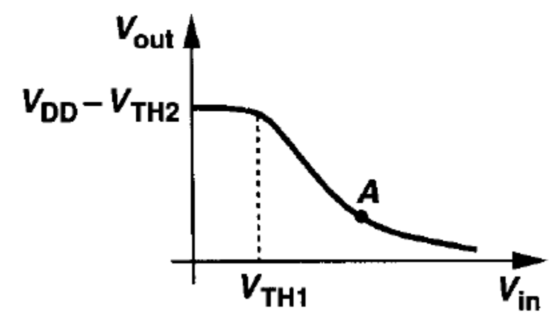
（1） $V_{in}<V_{TH1}$， 则 $I_{D1}=0=I_{D2}=\frac{1}{2}\mu_n C_{OX}\frac{W_2}{L_2}(V_{GS2}-V_{TH2})^2$,

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

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

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

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



PMOS as diode-connected load

This circuit is free from body effect!

channel-length modulation is neglected

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

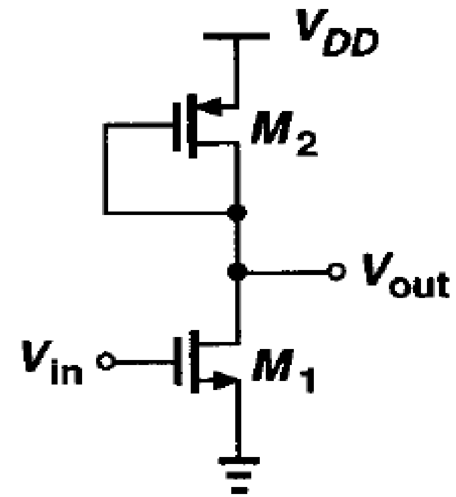


Fig. 3.16 CS stage with diode connected PMOS device

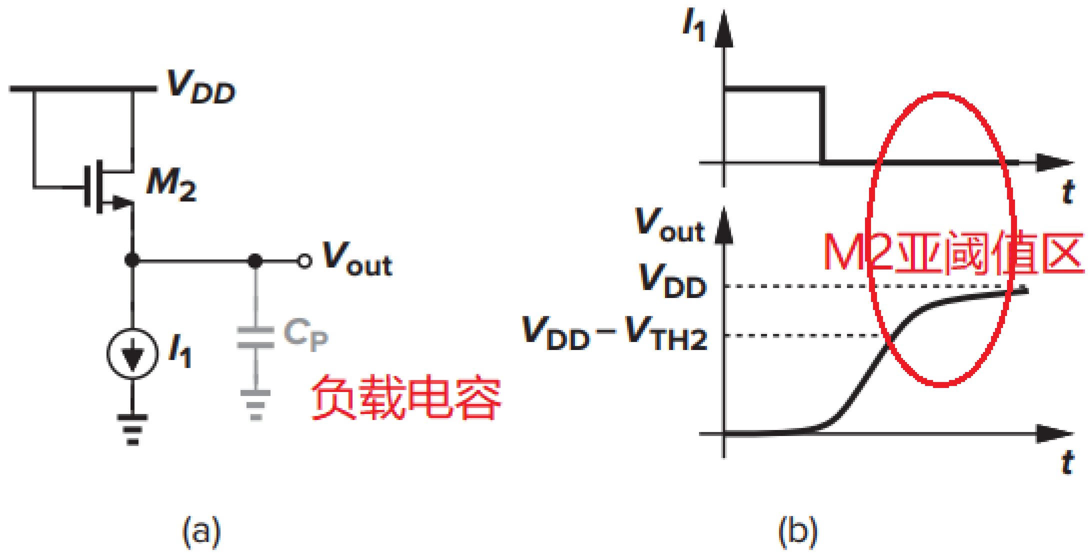
$$A_v = \frac{\partial V_{out}}{\partial V_{in}} = -\frac{\partial V_{GS2}}{\partial V_{in}} = -\sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}}$$

PMOS二极管负载的CS放大器增益比NMOS二极管负载增益大

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）。

二极管ID=0的再讨论：动态和静态的区别



若$M2$始终在饱和区，且$I_{D2} = \frac{1}{2}\mu_n C_{OX}\frac{W_2}{L_2}(V_{GS2}-V_{Tn2})^2 = 0 \Rightarrow V_{GS2}=V_{Tn2}$

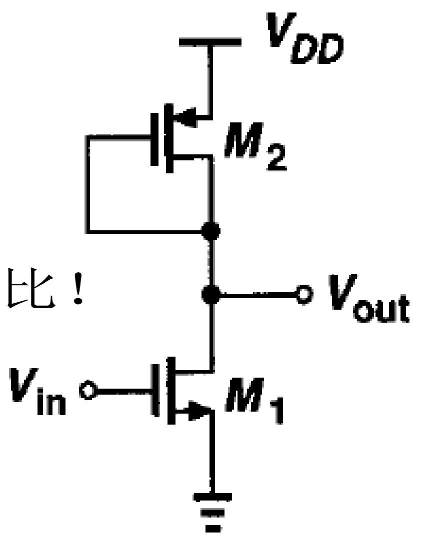
负载电容充电，使得VGS2小于VTH，进入亚阈值区。

ID长时间 = 0，静态，Vout ->VDD

ID不断变化，动态， ID短暂时间=0时， Vout ->VDD - VTH 。

Fault of CS with diode-connected load

$$\frac{1}{2}\mu_n C_{OX}\left(\frac{W}{L}\right)_1 (V_{\text{in}} - V_{TH1})^2 = \frac{1}{2}\mu_p C_{OX}\left(\frac{W}{L}\right)_2 (V_{\text{GS2}} - V_{TH2})^2$$



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

式(3.37)

$$\therefore \ \mu_n \left(W/L\right)_1 = \mathrm{A}_{\mathrm{v}}^2 \mu_p \left(W/L\right)_2 \text{ , } and$$

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

$$= -V_{out} + V_{DD} + V_{TH2} = -\mathrm{A}_{\mathrm{v}}(V_{GS1} - V_{TH1})$$

得 $V_{out} = V_{DD} + V_{TH2} + \mathrm{A}_{\mathrm{v}}(V_{GS1} - V_{TH1})$

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

解：$\mathrm{V}_{\text{out}} = 3 - 0.7 - 10 \times 0.2 = 0.3\mathrm{V}$，即$\mathrm{V}_{\text{GS2}} = 2.7V$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

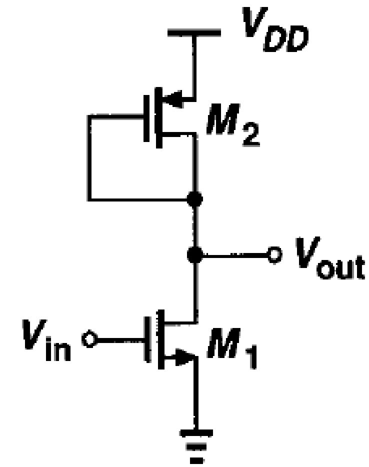
How to explain the paradox

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

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

矛盾？

大信号 $-V_{GS2}=-V_{DS2}=V_{DD}-V_{DS1}$与$V_{GS1}$联动



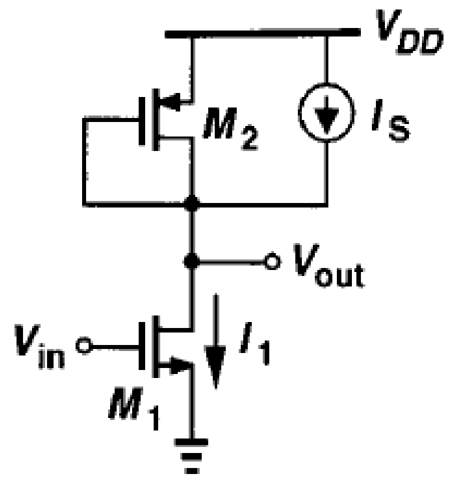
$$\frac{1}{2}\mu_n C_{OX}\left(\frac{W}{L}\right)_1 (V_{GS1}-V_{TH1})^2 = \frac{1}{2}\mu_p C_{OX}\left(\frac{W}{L}\right)_2 (V_{GS2}-V_{TH2})^2$$

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

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

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

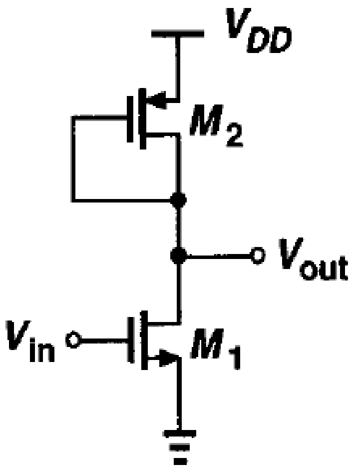
设电流源输出电阻极大

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

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

增加了$\sqrt{K}$增益，而且减小了K倍M2过驱动电压，扩大了输出动态范围

沟道长度调制效应：计入并联ro阻抗

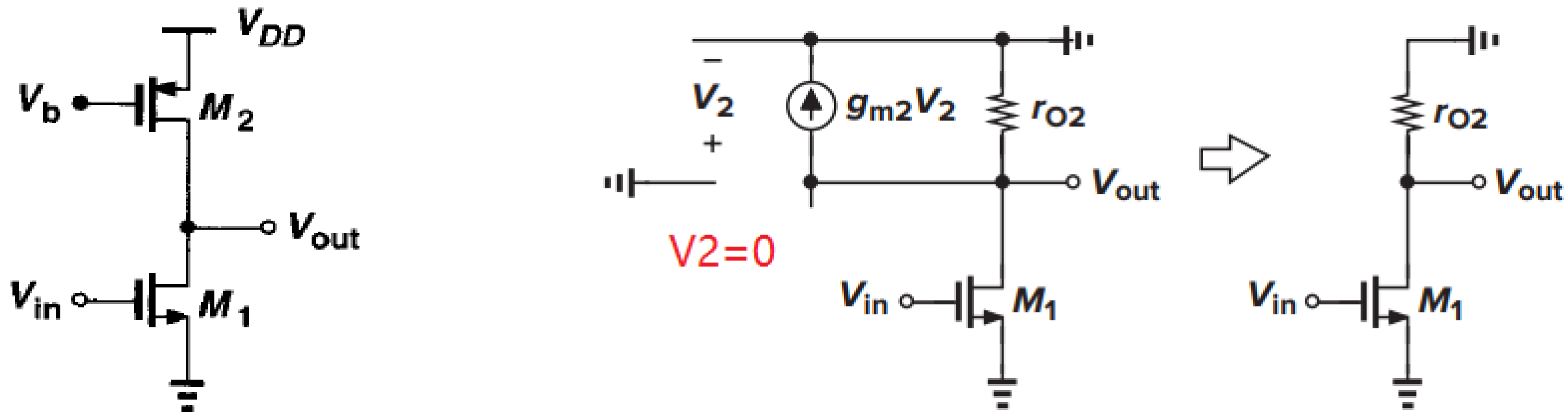


$$A_V \approx -g_{m1} \times R_D = -g_{m1}\left(\frac{1}{g_{m2}} \,||\, \mathrm{r}_{o1} \,||\, \mathrm{r}_{o2}\right)$$

3.3.3 CS Stage with current-source Load

若实际电阻做放大器的上拉负载，则直流和交流阻抗相同。由于输出工作点电压和直流电流的限制，一般情况下**RD**不能很大，故增益较小。

增大小信号RD的方法：饱和区M2电流源（L2大）做CS负载，ro2很大。



交变小信号增益： $A_v = -g_m(r_{o1} \,||\, r_{o2})$ $L\uparrow \Rightarrow r_o\uparrow \Rightarrow A_v\uparrow$

输出直流工作点电压可变化范围很大：

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

缺点：电流源负载CS放大电路的输出DC电位需要通过其它方法确定。

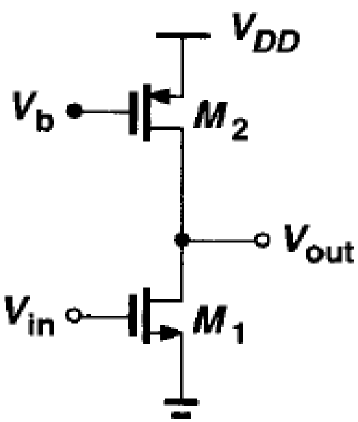
Increasing gain

（1）Intrinsic gain(本征增益):

$$g_m r_o = \sqrt{2\mu C_{ox}\frac{W}{L}I_D}\times\frac{1}{\lambda I_D} \propto \sqrt{2\mu C_{ox}\frac{W}{L}I_D}\times\frac{L}{I_D}$$

$$\therefore\ g_m r_o \propto \sqrt{2\mu C_{ox}\frac{WL}{I_D}}$$ 面积换增益

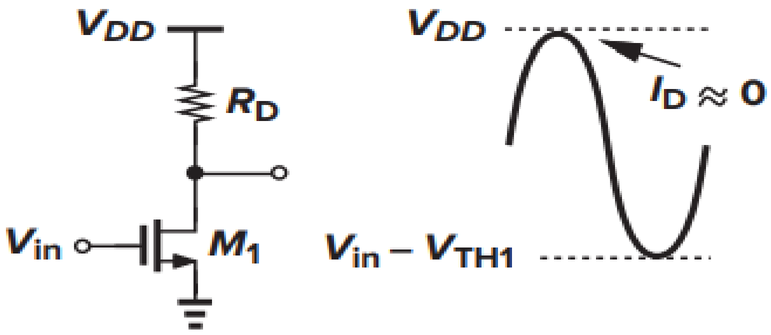
（2）电流源负载CS增益:



不妨设2管λ和L相同， $A_v = \sqrt{2\mu_n C_{ox}\left(\frac{W}{L}\right)_1 I_D}\times\frac{1}{2\lambda I_D}$

$$\propto \sqrt{2\mu_n C_{ox}\left(\frac{W}{L}\right)_1 I_D}\times\frac{L_1}{2I_D} = \sqrt{\mu C_{ox}\frac{(WL)_1}{2I_D}}$$

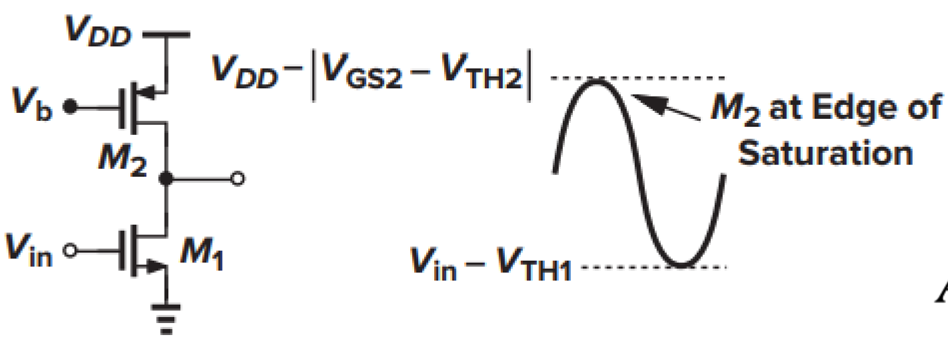
电阻负载和电流源负载CS输出电压比较



(a)

输出上限VDD

$$A_v = -g_{m1}(R_D || r_o)$$



(b)

输出上限 Vb+|VTH2|

= VDD −|VGS2-VTH2|

$$A_v = -g_{m1}(r_{o1} || r_{o2})$$

3.3.4 有源负载的CS

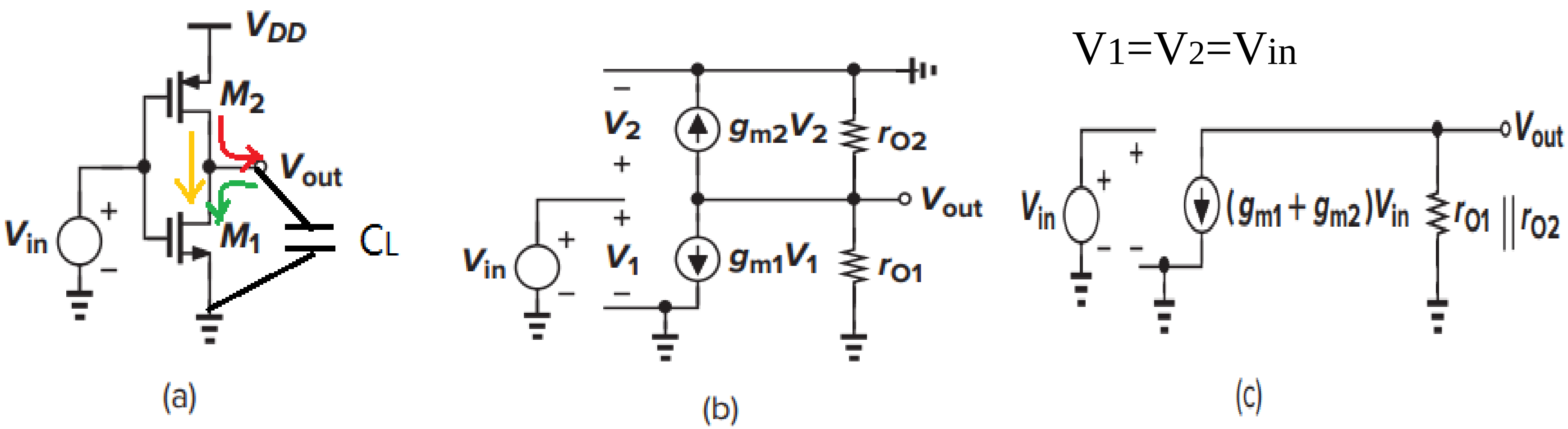


Figure 3.20 (a) CS stage with active load, (b) small-signal model, and (c) simplified model.

电路结构特点：输入管M1的负载管M2上加信号，M2成为“有源”负载；

同样理解，输入管M2的负载管M1上加信号，M1成为“有源”负载。

输出信号：2管相互增强，跨导增大！

$$\Delta V_{out} = (I_{D2} - I_{D1})\, {}^{t}/_{C_L}$$

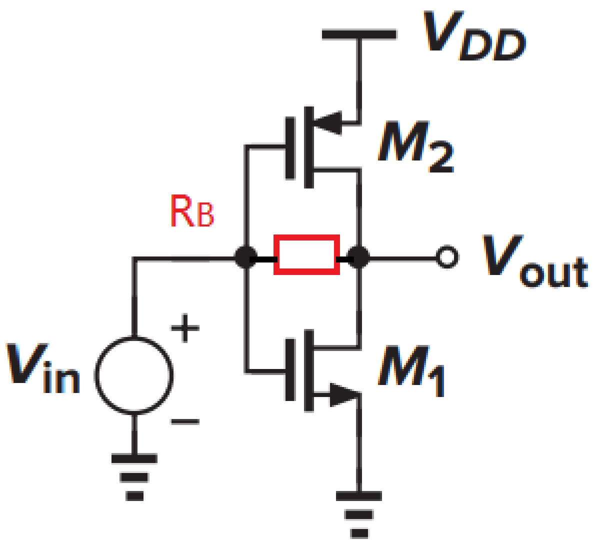
Vin增大，ID1增大ID2减小，CL放电，Vout下降加快
Vin降低，ID1减小ID2增大，CL充电，Vout上升加快

$$A_v = -(g_{m1} + g_{m2})(r_{o1}||r_{o2})$$

有源负载CS与反相器的区别

反相器属于数字电路，输入大幅度变化(VDD和GND) 2值逻辑高低电平信号。

有源负载CS放大器属于模拟电路，输入是直流电平（工作点）为1/2电源的小幅度变化模拟信号，合理的输出工作点电平也应是1/2电源。



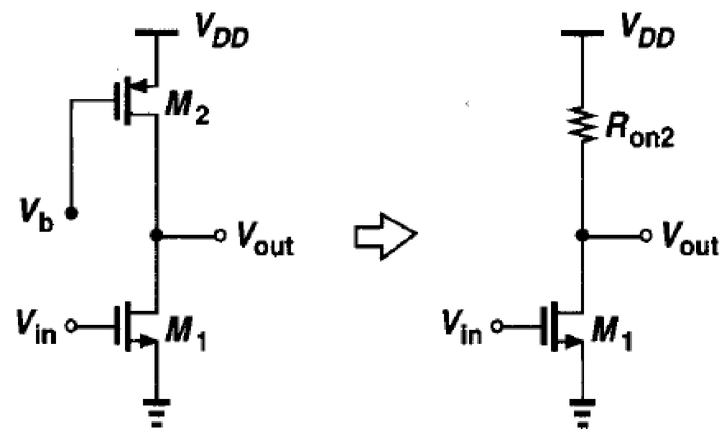
Vin变化幅度很小，可近似视为交流接地。

$$A_v = -(g_{m1} + g_{m2})(r_{o1}||r_{o2}||R_B)$$

3.3.5 CS with Triode MOSFET Load

深线性区工作的MOS器件，伏安特性为电阻。

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



缺点是$|V_{GS}|=V_{DD}-V_b$中噪声较大。

3.3.6 CS with source degeneration

源极负反馈(源简并)

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

$$I_D = \frac{1}{2}\mu_n C_{OX}\frac{W}{L'}(V_{GS} - V_{Tn})^2$$

VGS变化大时有明显的非线性。
设计思想：采用负反馈（假设需要利用负反馈的其它特性）

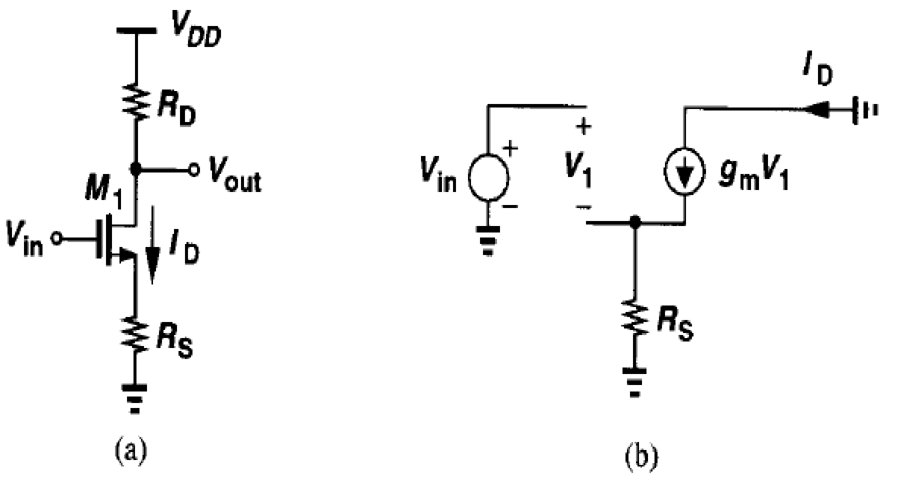


图3.23 带源极负反馈的共源极

使输入电压Vin一部分损失在Rs上，

VGS减小，使得 ID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

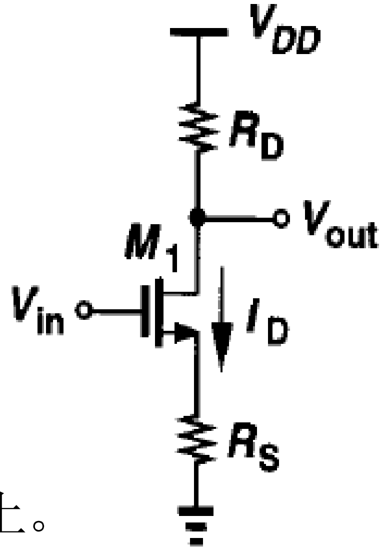
$R_S \neq 0$时：$V_{GS} = V_{in} - I_D R_S$

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。

带源极反馈电阻CS增益的简便计算方法

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

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**

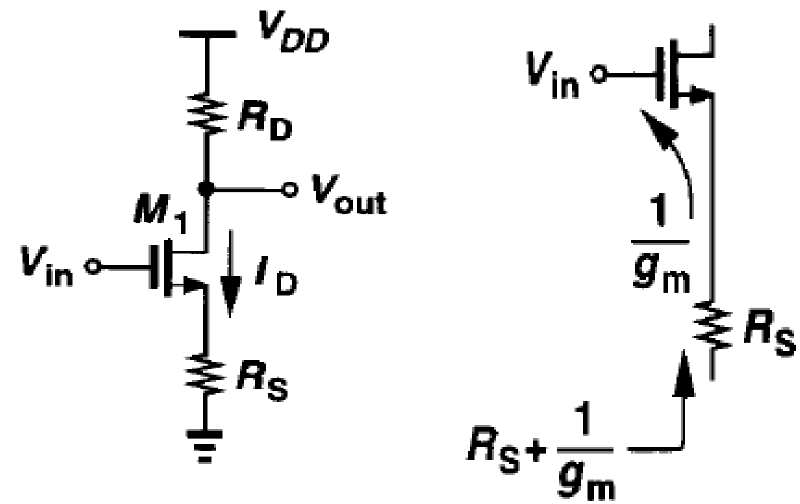
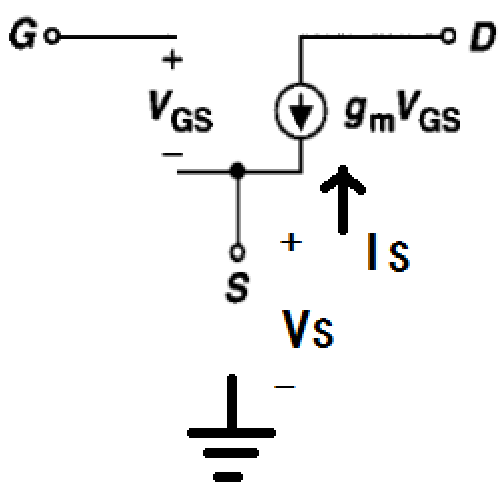


Figure 3.27 resistance seen in the source path



假设可近似：

$$\lambda = \gamma = 0$$

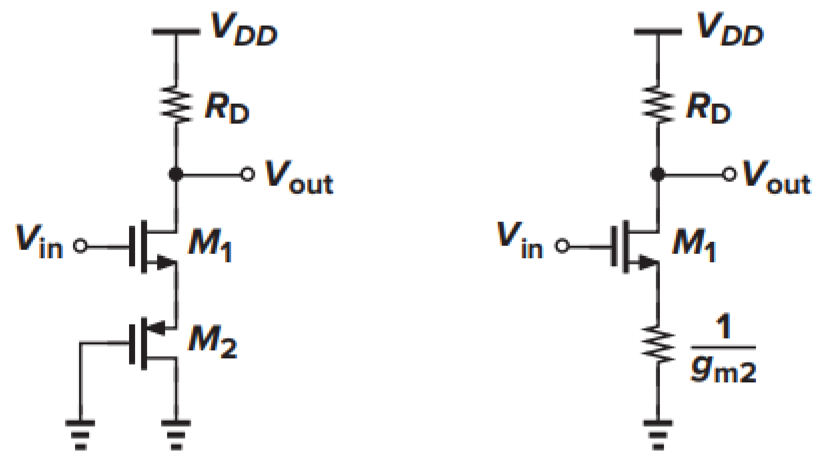
$$MOS\text{的}S\text{极小信号电阻} = \frac{V_S}{I_S} = \frac{-V_{GS}}{-g_m V_{GS}} = \frac{1}{g_m}$$

（∵ 求输出阻抗时输入电压 = 0，即G交流接地）

Example

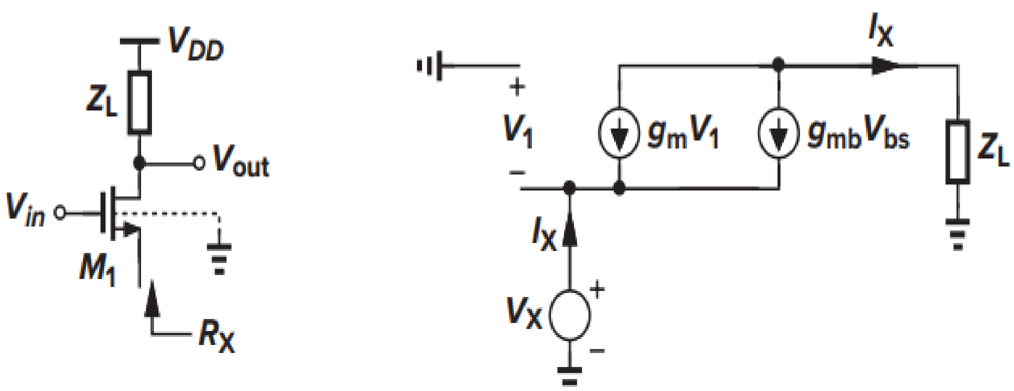
(1)设 $\lambda = 0, \gamma = 0$

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



(a) (b)

(2)求Rx? 设 $\lambda = 0, \gamma \neq 0$



(a) (b)

V1= -Vx,Vbs= -Vx

(gm+gmb)Vx=Ix

Rx=Vx/Ix=1/(gm+gmb)

因gmb实质是从源-衬底加入信号，增益将减小，Av计算不能直接将gm1改为gm1+gmb1

辅助定律（电路基本理论，适用交流信号）

在线性电路中，开路电压增益等于$-G_mR_{out}$，其中G_m表示输出与地短接（恒压）时的电路跨导（$G_m=I_{out}/V_{in}$）。R_{out}表示当输入电压为零时（$I_{out}= G_mV_{in}=0$ ）电路的输出电阻。

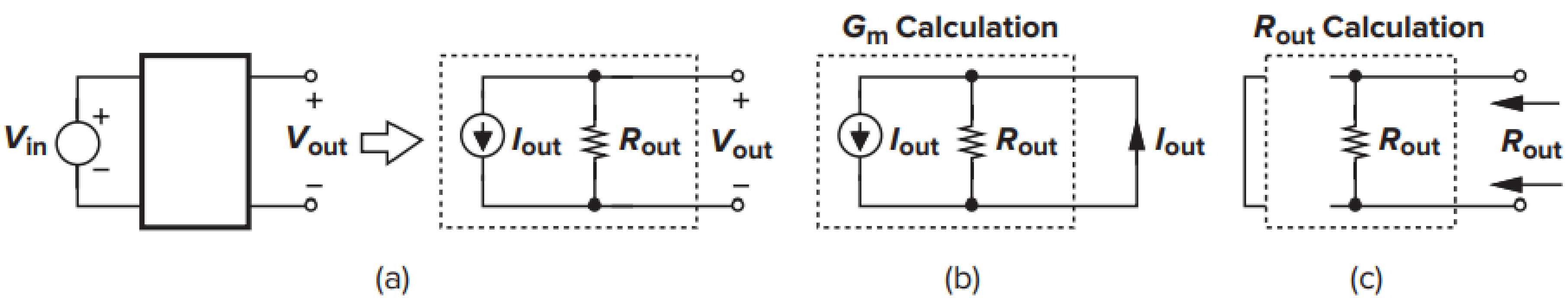


Figure 3.32 (a) Norton equivalent of a linear circuit; (b) G_m calculation; and (c) R_{out} calculation.

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

带负载R_D时小信号输出电压：$V_{out}=-I_{out}(R_{out}\ ||\ R_D)$

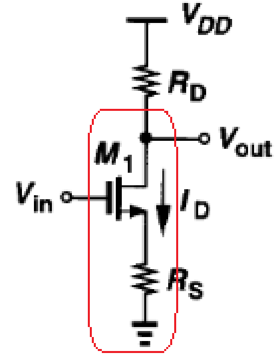
$=-V_{in}G_m(R_{out}\ ||\ R_D)$

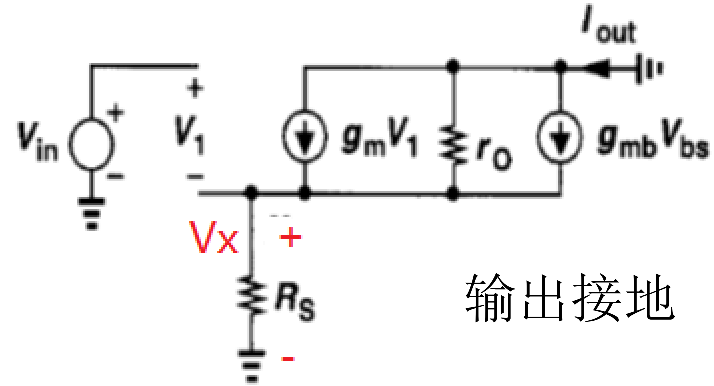
负载RD可后加计算

Gm with body effect and channel-length modulation

考虑沟道长度调制和体效应。

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



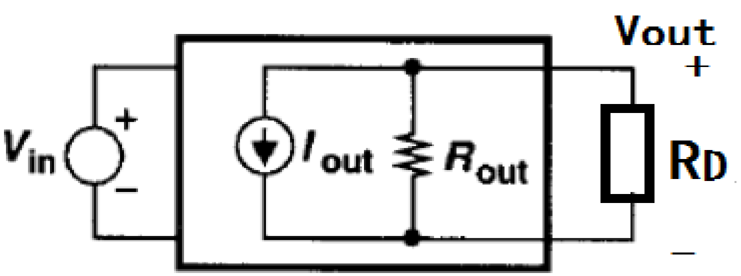


输出接地

$$\text{输出接地交变电流} I_{out} = g_mV_1 + g_{mb}V_{bs} - \frac{V_X}{r_o} = g_mV_1 - g_{mb}V_X - \frac{V_X}{r_o}$$

$$= g_mV_1 - g_{mb}I_{Rs}R_S - \frac{I_{Rs}R_S}{r_o} = g_m(V_{in} - I_{out}R_S) - g_{mb}I_{out}R_S - \frac{I_{out}R_S}{r_o}$$

$$\longrightarrow \quad G_m = \frac{I_{out}}{V_{in}} = \frac{g_mr_o}{R_S + [1 + (g_m + g_{mb})R_S]r_o}\text{，式（3.61）}$$



Iout=GmVin

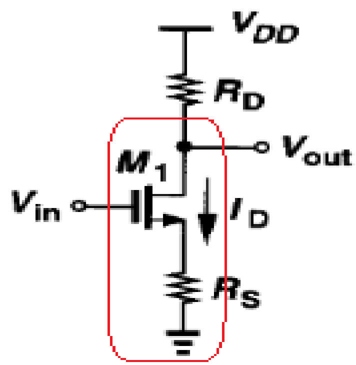
$$A_v = -G_m(R_{out} \parallel R_D) \approx -G_mR_D$$

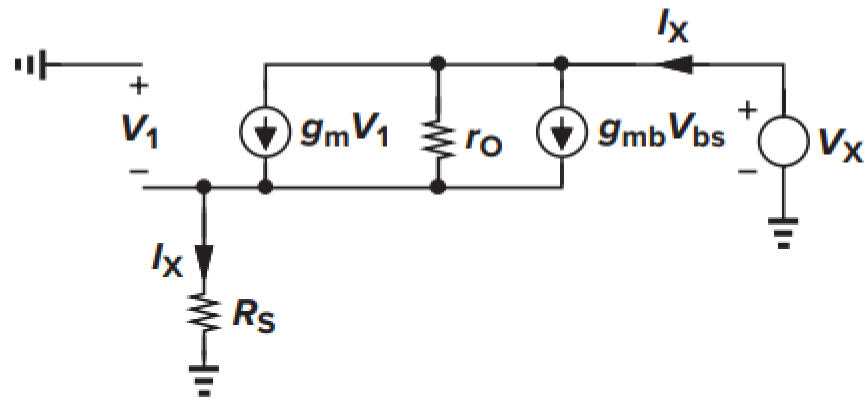
Rout见式（3.66）

Output resistance of degenerated CS

$$R_{out}=\frac{V_x}{I_X}$$

令$V_{in}=0$，先不考虑R_D





$$V_{bs}=V_1=-I_XR_S$$

图3.29

$$V_X=r_0[I_X-g_mV_1-g_{mb}V_{bs}]+I_XR_S$$
$$=r_0[I_X+(g_m+g_{mb})I_XR_S]+I_XR_S$$

$$R_{out}=\frac{V_x}{I_X}=[1+(g_m+g_{mb})R_S]r_0+R_S=[1+(g_m+g_{mb})r_0]R_S+r_0,\quad 式(3.66)$$

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

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

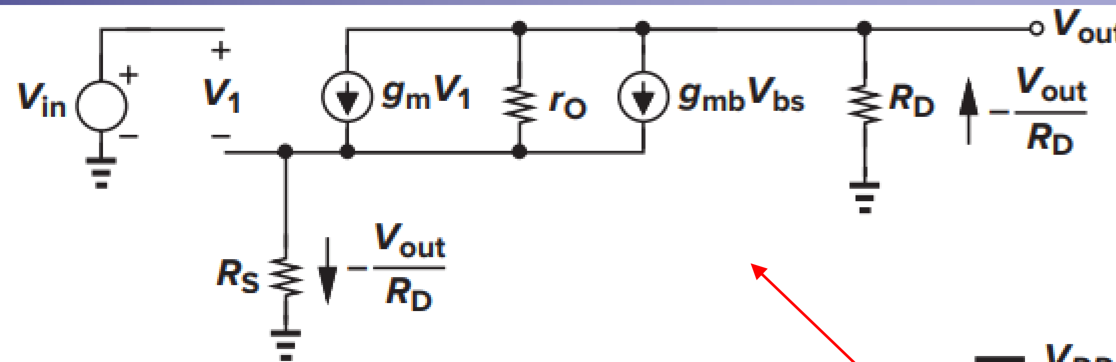
源极负反馈:

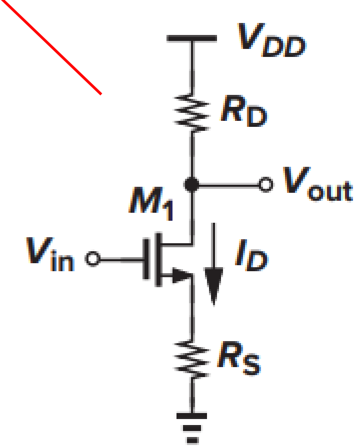
增大共源级的输出电阻。

验证：包含RD的公式推导

$$I_{R_S} = -I_{R_D} = -\frac{V_{out}}{R_D}$$

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





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

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

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

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

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

（3.77）

验证：包含RD的公式推导（续）

分析Av分母大括号项（=带负反馈电阻的共源极的输出电阻Rout）。

$$\because \quad R_{out} = [1+(g_m+g_{mb})r_0]R_S + r_0\text{，式}(3.66)$$

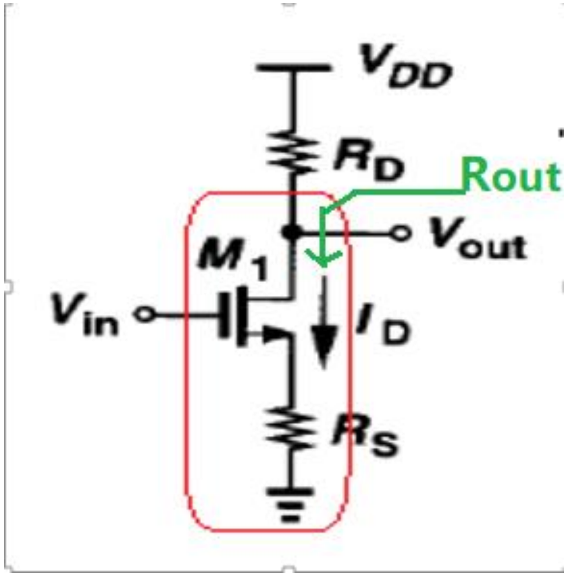
$$A_V = -\frac{g_m r_o R_D}{R_D + \{[1+(g_m+g_{mb})r_o]R_S + r_o\}} \bullet \frac{[1+(g_m+g_{mb})r_o]R_S + r_o}{[1+(g_m+g_{mb})r_o]R_S + r_o}$$

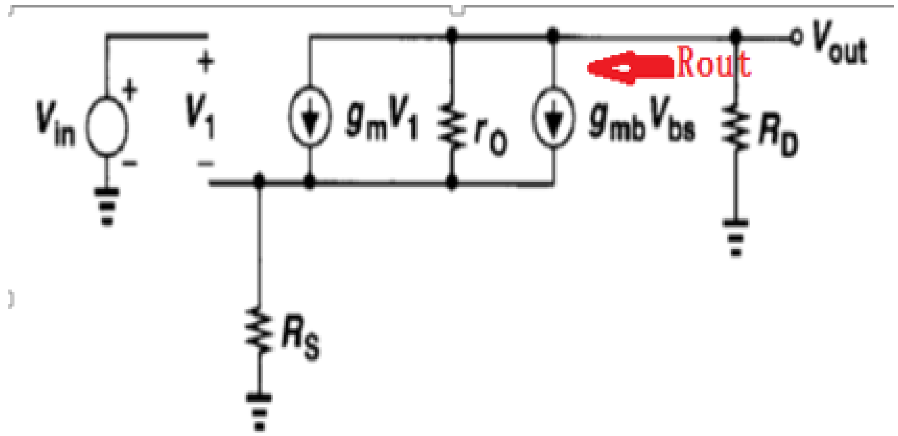
$$= -\frac{g_m r_o}{[1+(g_m+g_{mb})r_o]R_S + r_o} \bullet \frac{R_D\{[1+(g_m+g_{mb})r_o]R_S + r_o\}}{R_D + [1+(g_m+g_{mb})r_o]R_S + r_o}$$

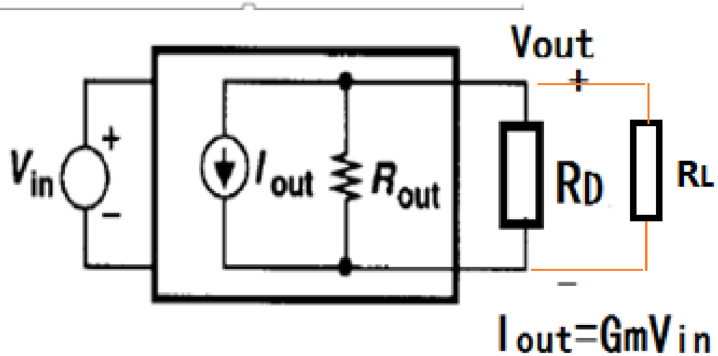
$$= -G_m(R_D \,||\, Rout)$$

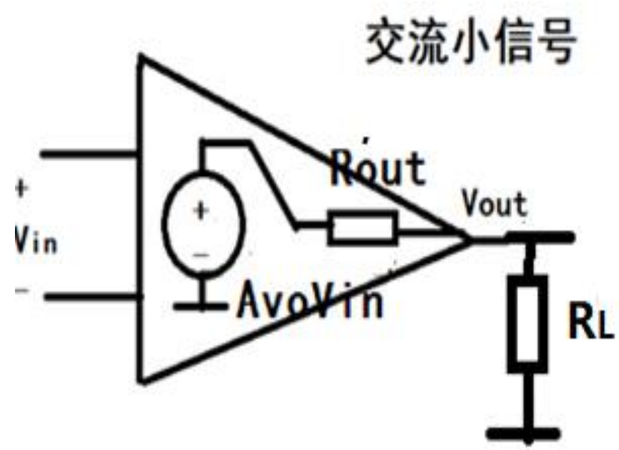
与先不计算**RD**，再将**RD**加入电路的计算方法结果一致

Gain of a degenerated CS





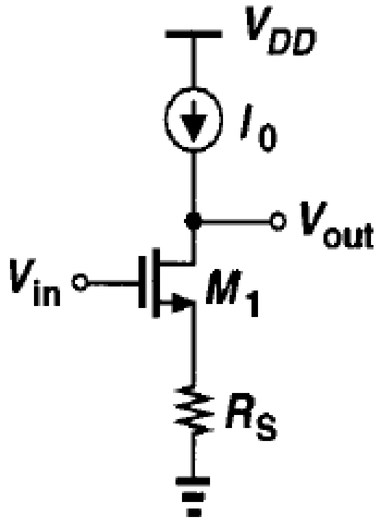




Vout= -GmVin*(Rout||RD||RL)
Gm输出接地求出

Av0= -Gm(Rout||RD)
Av=Av0*RL/(Rout||RD+RL)

例 3.11 calculate the voltage gain



Solution

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

$$= -g_m r_o, \quad \text{式(3. 77)中} R_D = \infty$$

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

3.4 source follower 源跟随器

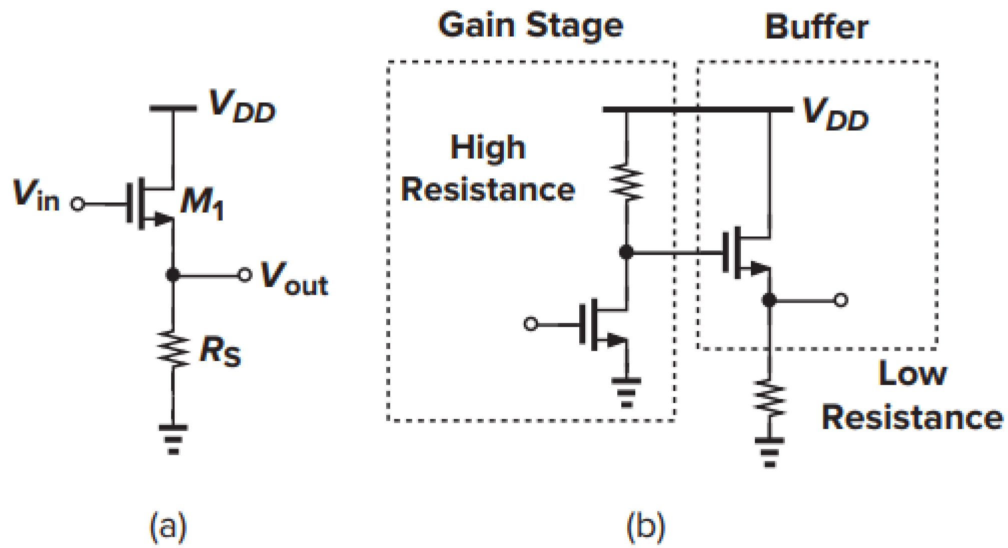
common-drain stage

特点:

*小信号输出电压近似等于输入电压（小信号Rs大）；

*低频输入阻抗大,

*输出阻抗小（1/gm）。

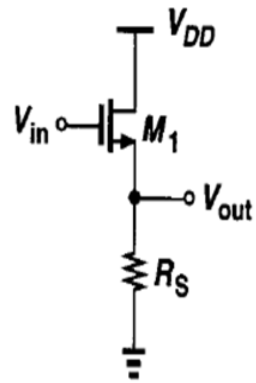


作用： 当两级放大器级联时，若前级输出电阻大，或后级输入电阻RL小，则在两级放大器之间插入源跟随器，可以提高多级放大器的电压增益，即用作缓冲（驱动）电路，要求：输出阻抗（1/gm)远小于后级电路输入阻抗RL。

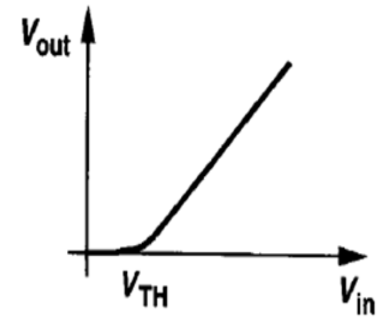
Large-signal & Small-signal behavior

忽略沟道长度调制效应

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



(a)



(b)

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

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

由第2章： $\dfrac{\partial V_{TH}}{\partial V_{in}} = \dfrac{\partial V_{TH}}{\partial V_{SB}}\dfrac{\partial V_{SB}}{\partial V_{in}} = \eta\dfrac{\partial V_{out}}{\partial V_{in}} = \eta A_v$

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

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

Gain of Source Follower(SF)

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

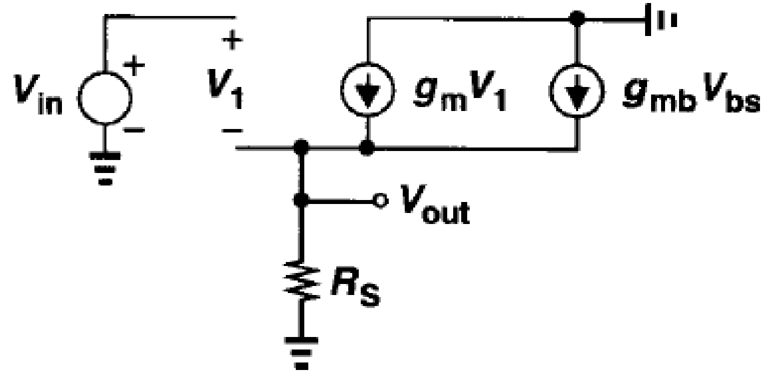
$$g_m = \mu_n C_{ox} \frac{W}{L}(V_{in} - V_{out} - V_{TH})$$

$$g_{mb} = \eta g_m = \frac{\gamma}{2\sqrt{2\Phi_F + V_{SB}}} g_m$$

SF小信号等效电路得到相同结果

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

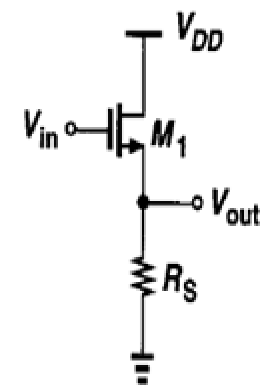
$$V_{bs} = -V_{sb} = -V_{out}$$ 为何ro可忽略?



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

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

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

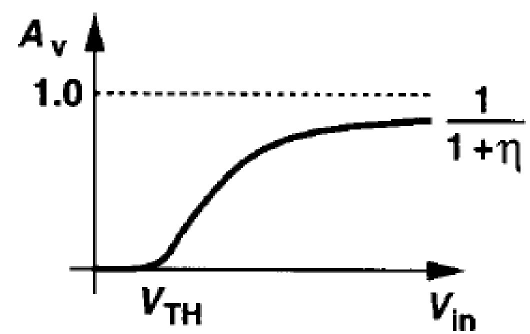


$$\eta = \frac{g_{mb}}{g_m}$$ 可取0. 2估算

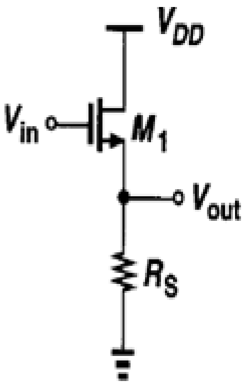
与大信号推导结果相同

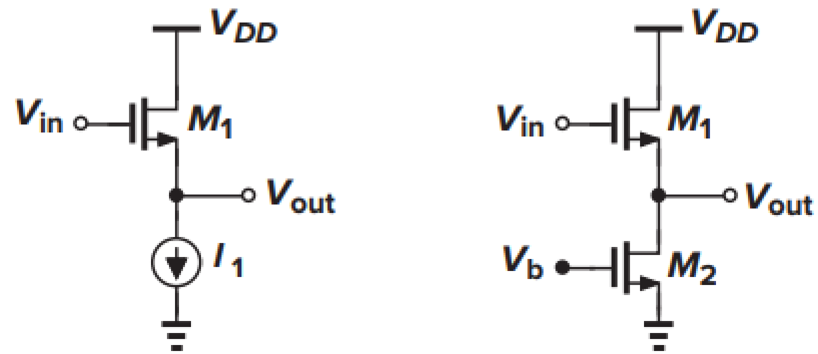
当$g_m R_s >> 1$ 时

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



采用电流源的源跟随器





(a) (b)

$R_s = \propto$ 时

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

最大增益=MOS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）；

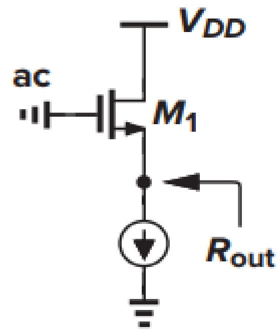
2。直流电平移动导致电压余度(headroom)减少。

源跟随器的输出电阻

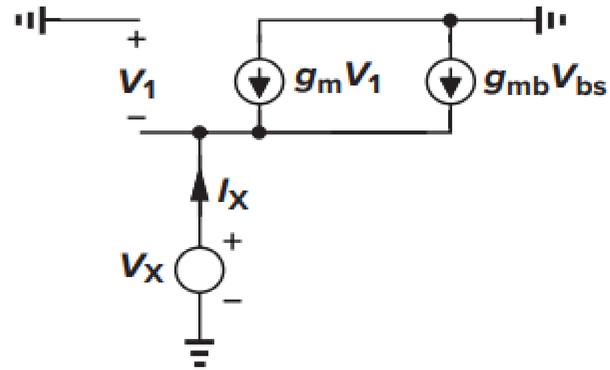
$$\sup pose:\lambda \approx 0$$

$$V_1 = -V_X,$$

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



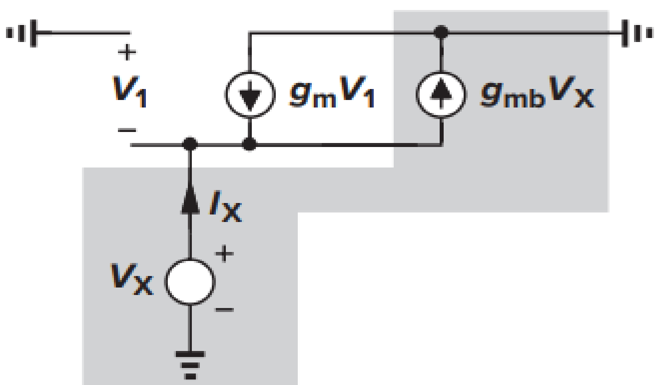
(a)



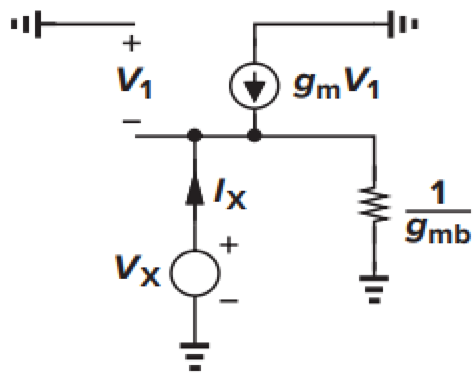
(b)

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

与二极管相同!
几十欧~几千欧，很小。



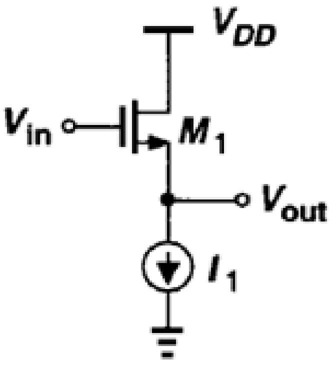
(a)

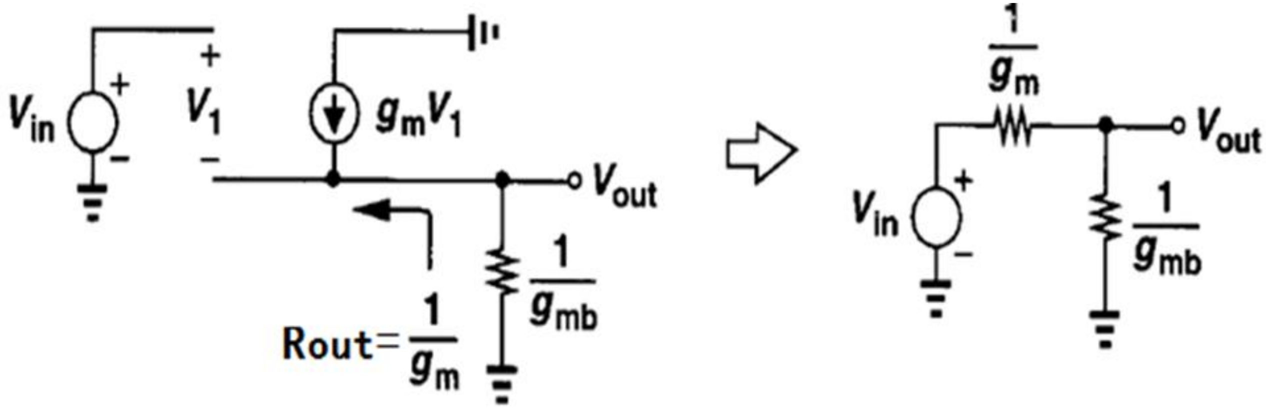


(b)

体效应等效于
输出电阻1/gmb.

另一种方法：用辅助定理求Intrinsic gain of SF





求R_{out}：令$V_{in}=0$，则$V_{out}=-V_1$

负载开路$R_{out}=\dfrac{V_{out}}{-g_mV_1}=\dfrac{1}{g_m}$

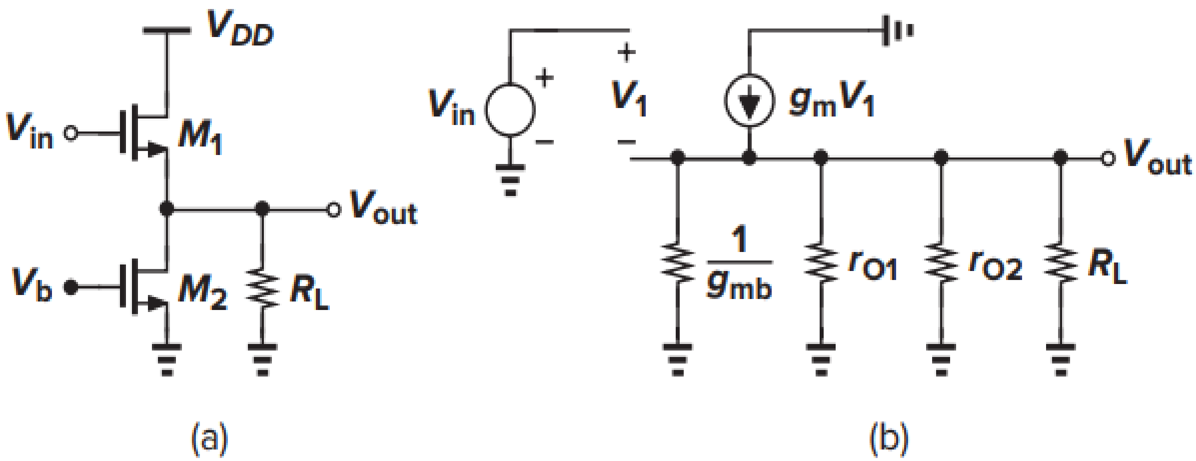
交流输出短路（$V_1=V_{in}$）电流：$\quad I_{SC}=g_mV_1=g_mV_{in}$

开路输出电压（戴维宁电路）：$\quad V_{OC}=R_{out}I_{SC}=\dfrac{1}{g_m}g_mV_{in}=V_{in}$

无实际R_L负载时$A_v=\dfrac{V_{out}}{V_{in}}=\dfrac{1/g_{mb}}{1/g_m+1/g_{mb}}\Rightarrow A_v=\dfrac{g_m}{g_m+g_{mb}}=\dfrac{1}{1+\eta}$

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。

考虑体效应和沟道长度调制效应的SF



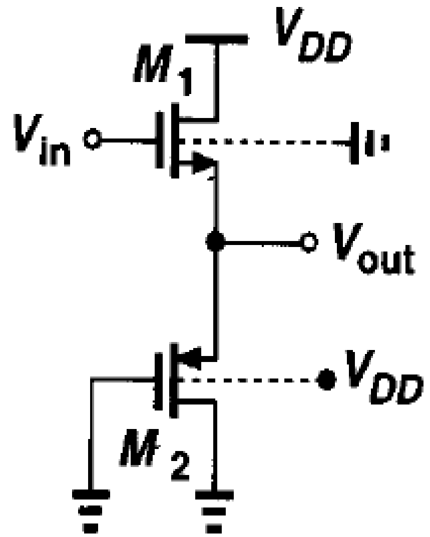
Vin>V_{GS1}+ Vb-V_{TH2}

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

公式中的**gm**和**gmb**是**M1**参数.

例3.14

$M2$为二极管，输出阻抗 $= \dfrac{1}{g_{m2}+g_{mb2}} || r_{o2}$



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

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

NMOS SF输出电平Vout低于输入Vin至少V_{TH}

PMOS的SF:消除体效应的非线性

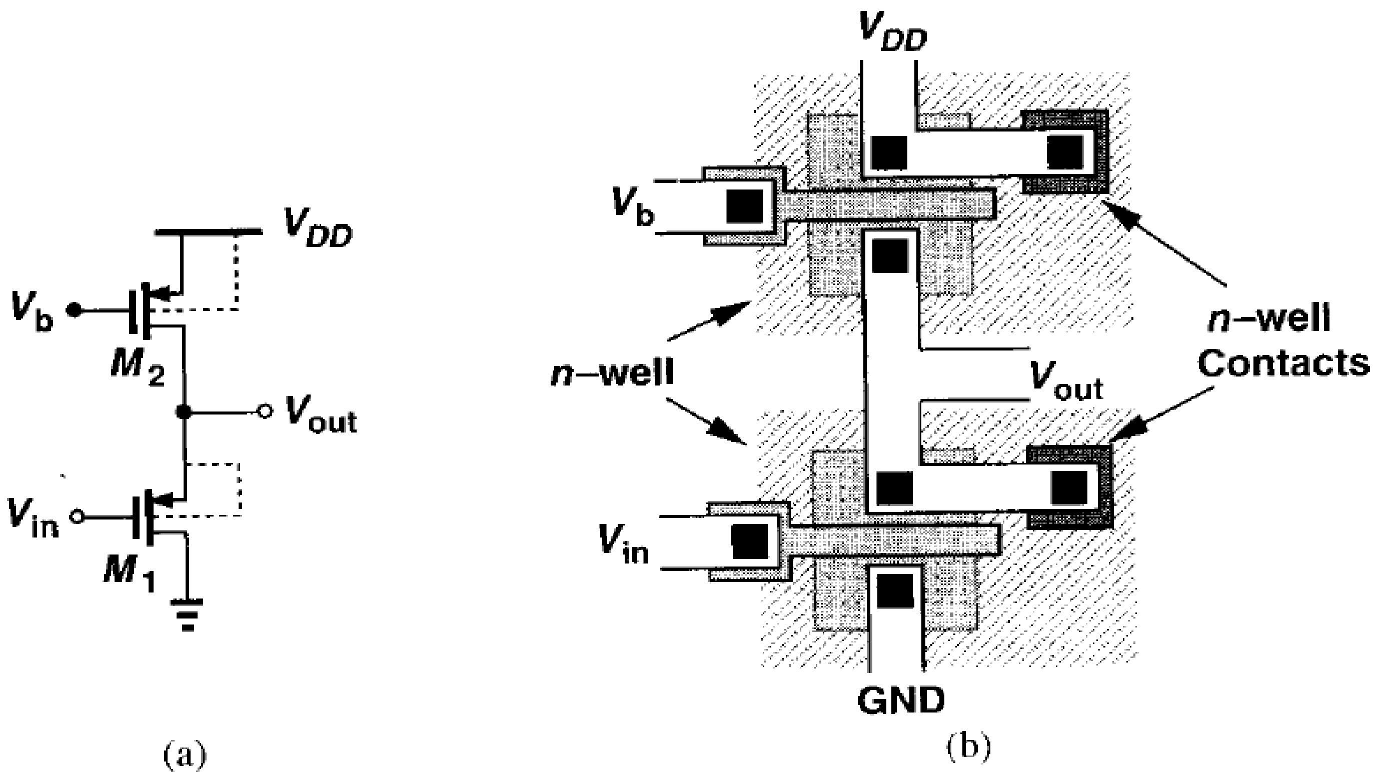
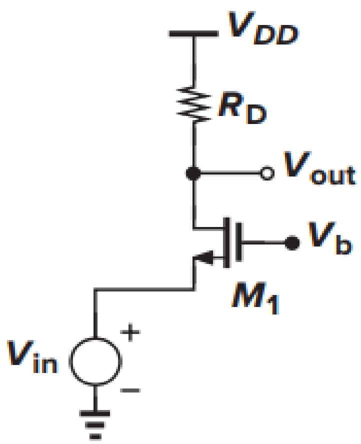


Figure 3.36 PMOS source follower with no body effect.

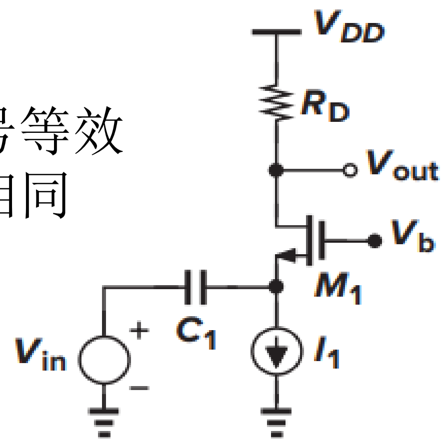
PMOS SF输出电平Vout高于输入Vin至少V_{TH}

3.5 Common-gate stage



(a)

小信号等效电路相同



(b)

信号从源端输入，从漏极输出。小信号输入输出共栅

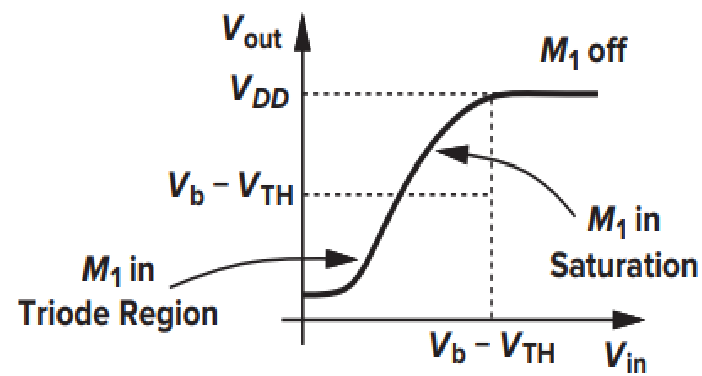
图(b)中，设C1足够大。Vin不影响M1直流工作点

低频直接连输入信号源，工作点与前级有关

高频信号通过C1耦合，I1确定工作点

设图(a)Vin从大的正值减小：

图(a): Vin>Vb-VTH时M1截止, 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

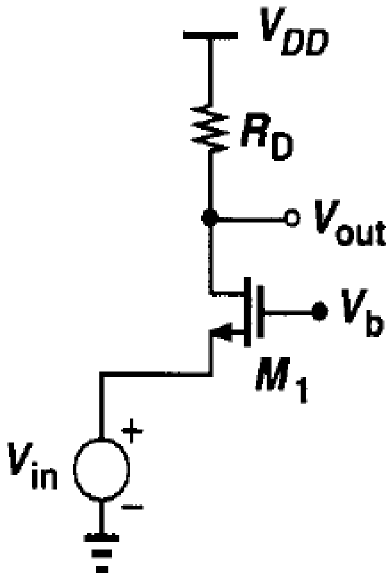
Gain of CG

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

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

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

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



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。

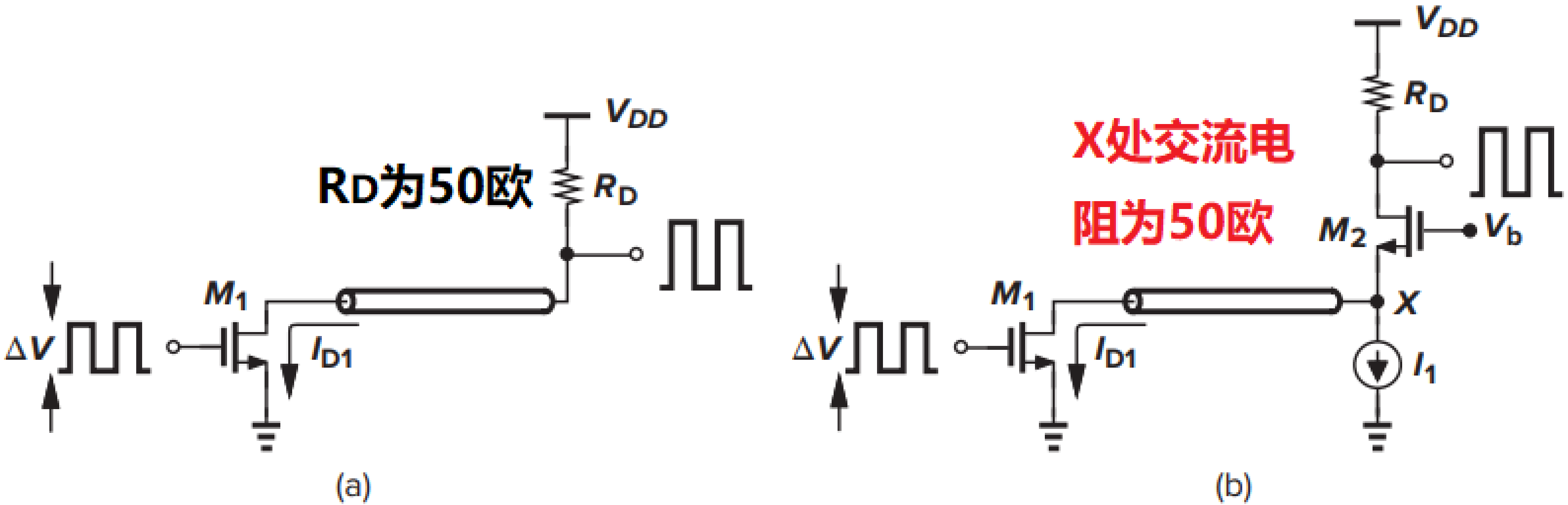
设$\lambda \approx 0$即ΔV_{DS}对I_D交变无影响，

或R_D不太大(I_D不恒定)时：

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

衬偏效应使输入阻抗降低。
共栅级电路应用在 低输入阻抗的场合

例3.17：传输线电路CS和CG的比较

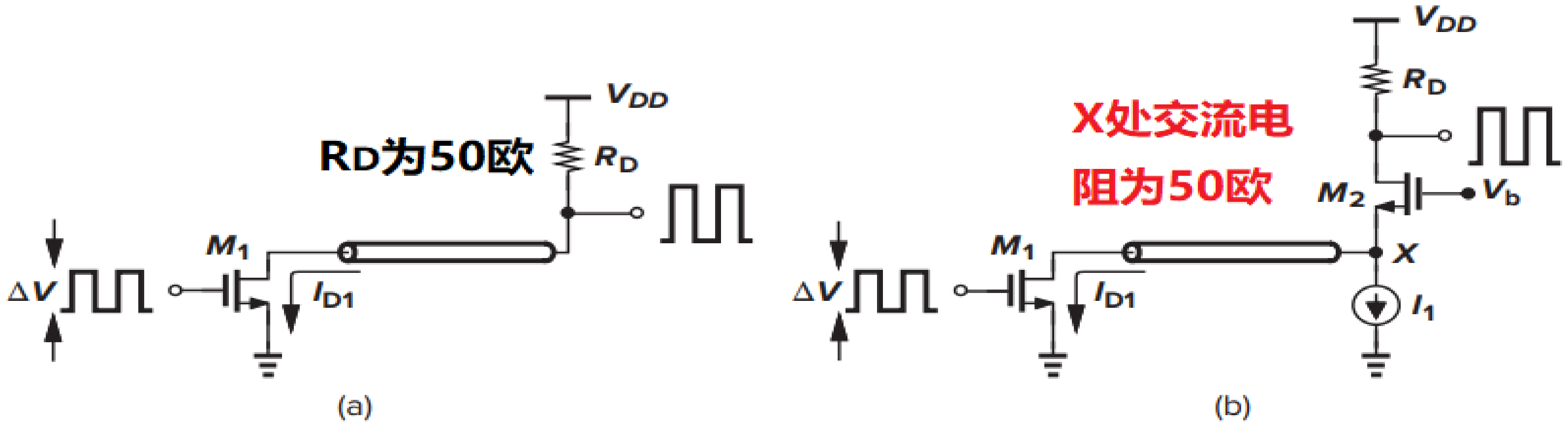


传输线特征电阻50欧，RD=50欧，忽略沟道长度调制 和体效应 。

(a) 两种接法的增益（忽略MOS寄生电容）。

(b) X点反射最小的条件下 (即M2源极看到的交流电阻为50欧)，M2放大器如何设计？

例3.17（续）



解：设两电路输入Vin变化 ΔV，$\Delta I_{D1} = g_m \Delta V$ 相同， $g_m = \sqrt{2\mu_n C_{ox} \frac{W}{L} I_D}$

$$A_v = \frac{\partial V_{out}}{\partial V_{in}} = -g_m R_D$$ 公式形式相同

注意：图(b)共栅级M2的输入信号（X处）是电流，而不是电压。

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

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均表示交流小信号

$$V_{out} =$$

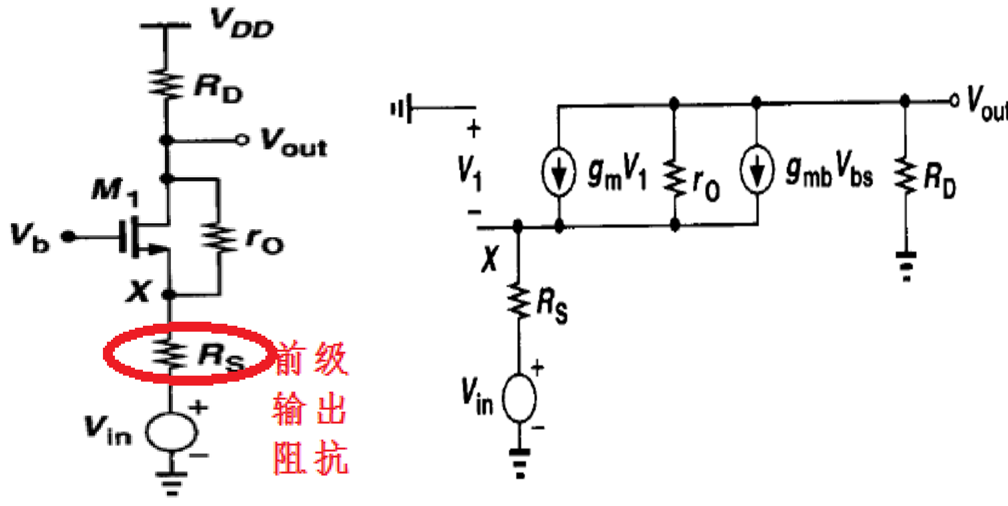
$$r_o\left(\frac{-V_{out}}{R_D} - g_m V_1 - g_{mb} V_{bs}\right) - \frac{V_{out}}{R_D} R_S + V_{in}$$

$$V_{bs} = V_1$$

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

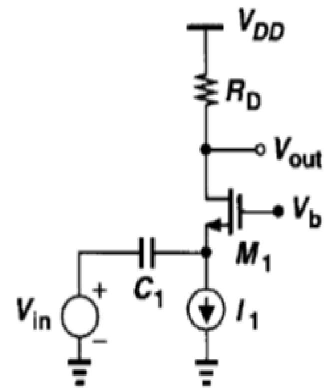
$$A_V = \frac{V_{out}}{V_{in}} = \frac{r_o(g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb}) r_o R_S + R_S + R_D} R_D \qquad (3.111)$$

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



(a)

图3.51 输出电阻有限的共栅级



CG 比较CS

- CS $$A_V = \frac{V_{out}}{V_{in}} = \frac{-r_o g_m}{r_o + (g_m + g_{mb})r_o R_S + R_S + R_D} R_D \qquad (3.77)$$

- CG $$A_V = \frac{V_{out}}{V_{in}} = \frac{r_o(g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb})r_o R_S + R_S + R_D} R_D \qquad (3.111)$$

CG增益略高一些，CG噪声比较大。

例3.18 求电压增益

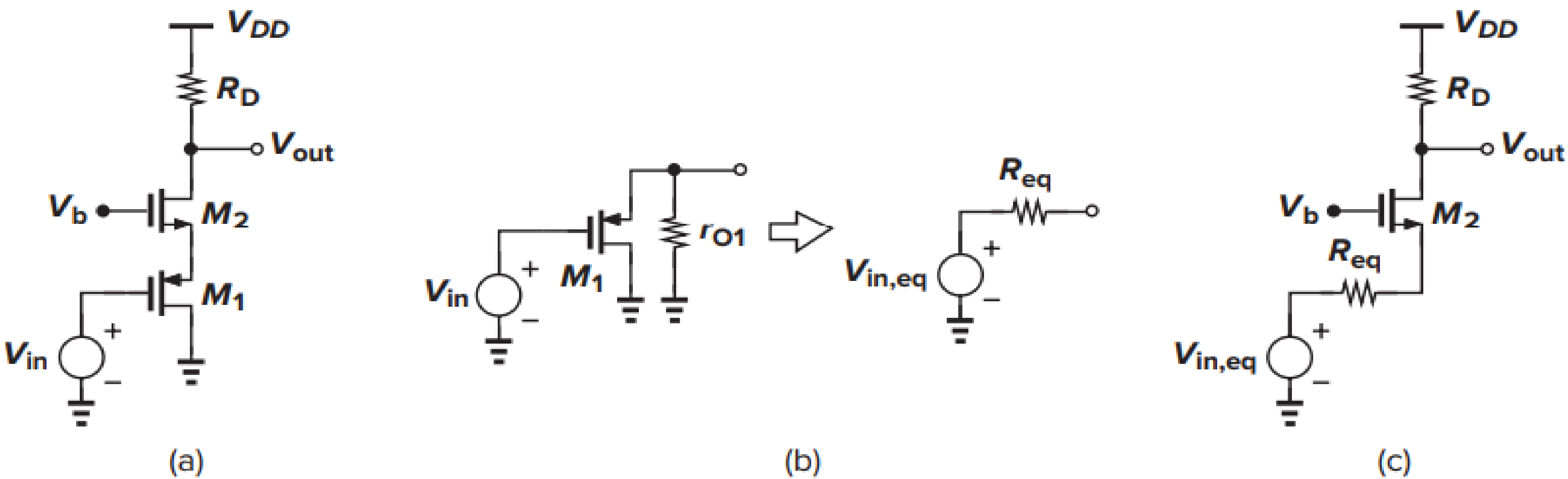


Figure 3.52

M1为PMOS源跟随器。将M1等效成戴维南电路，**化为标准CG电路形式**。**求戴维南电路等效电阻时，Vin=0，则M1成二极管**

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

M1输出短路电流为 $g_{m1}V_{in}$，PMOS管g_{mb1}为0

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

例3.18 求电压增益（续）

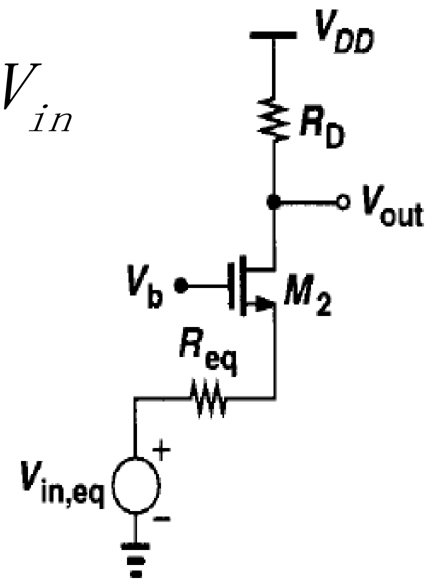
$$R_{eq} = \frac{1}{g_{m1}} \, || \, r_{o1} = \text{公式中}R_S$$

由前CG公式：$$A_V = \frac{V_{out}}{V_{in}} = \frac{r_o(g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb})r_o R_S + R_S + R_D} R_D \qquad (3.111)$$

$$V_{in,eq} \approx V_{in}$$

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

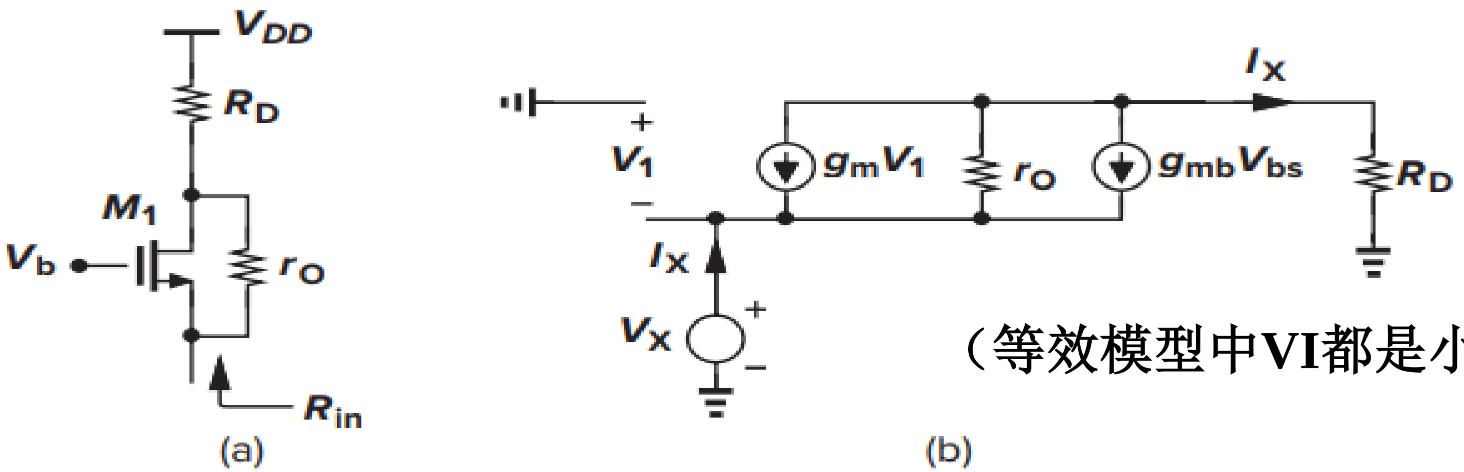
$$= \frac{r_{o2}(g_{m2} + g_{mb2}) + 1}{r_{o2} + [1 + (g_{m2} + g_{mb2})r_{o2}]\left(\frac{1}{g_{m1}} \, || \, r_{o1}\right) + R_D} R_D$$



(c)

设PMOS管gmb1为0，否则为教材式3.114

共栅级输入电阻



（等效模型中VI都是小信号）

Figure 3.53 (a) Input resistance of a CG stage; (b) small-signal equivalent circuit.

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

从下往上流过ro的电流： $$I_X + g_m V_1 + g_{mb} V_{bs} = I_X - (g_m + g_{mb}) V_X$$

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

共栅级输入电阻（从源极端往MOS看）

$$R_{in} = \frac{V_X}{I_X} = \frac{R_D + r_o}{1 + (g_m + g_{mb}) r_o} \qquad \textbf{(3.116)}$$

CG极端情况:RD=0和无穷大

$$R_{in} = \frac{V_X}{I_X} = \frac{R_D + r_o}{1+(g_m + g_{mb})r_o}$$ （比较先前假设ro无穷大）

（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（3.91）式相同

（2） 图3. 54， $R_D = \infty$时$R_{in} = \infty$

原因：交变小信号 $I_X = 0$

结论：只有RD不很大情况下，共栅级输入电阻才会较小。

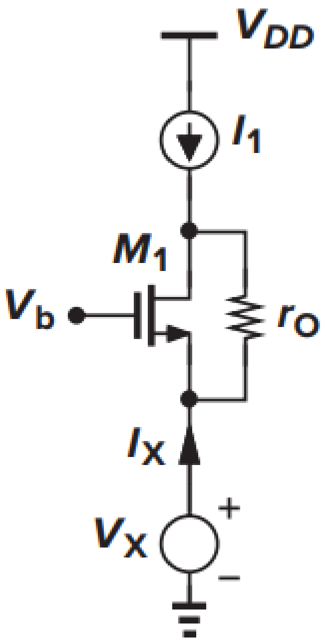


Figure 3.54 Input resistance of a CG stage with ideal current-source load.

例3.19 电流源负载CG的电压增益

直接耦合的共栅极。

由式（3.111 ）

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

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

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

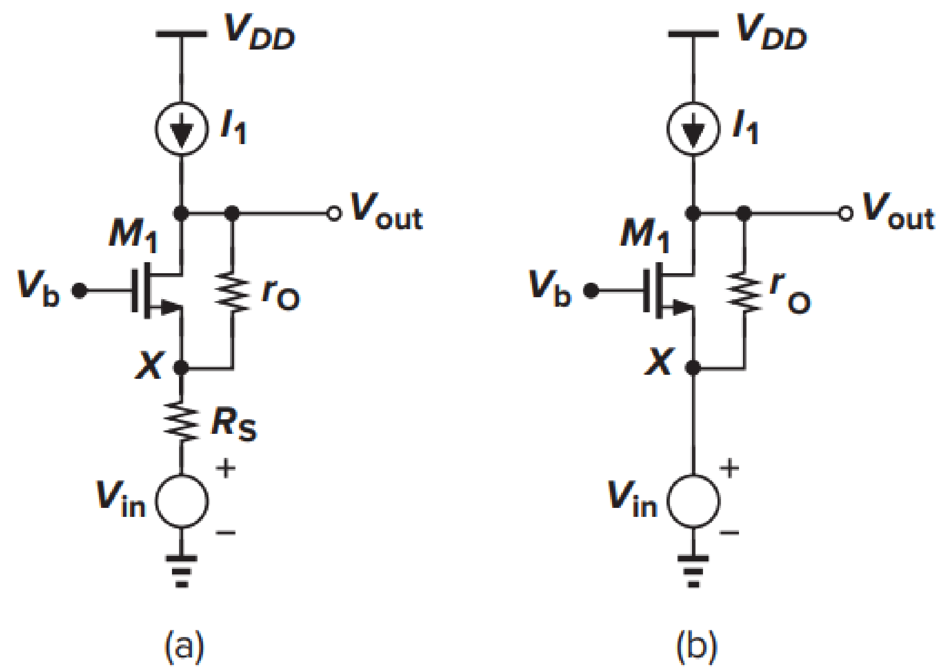


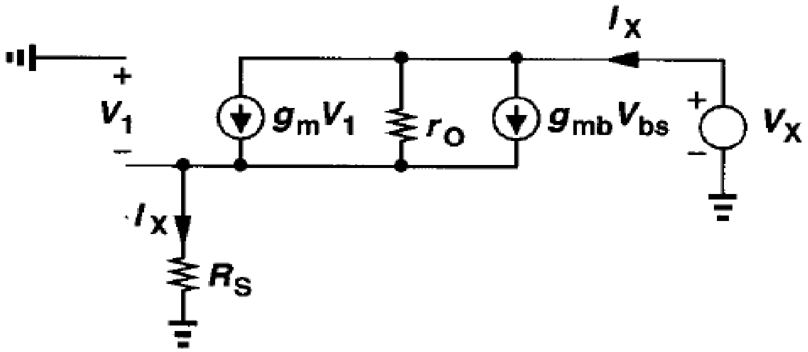
Figure 3.55

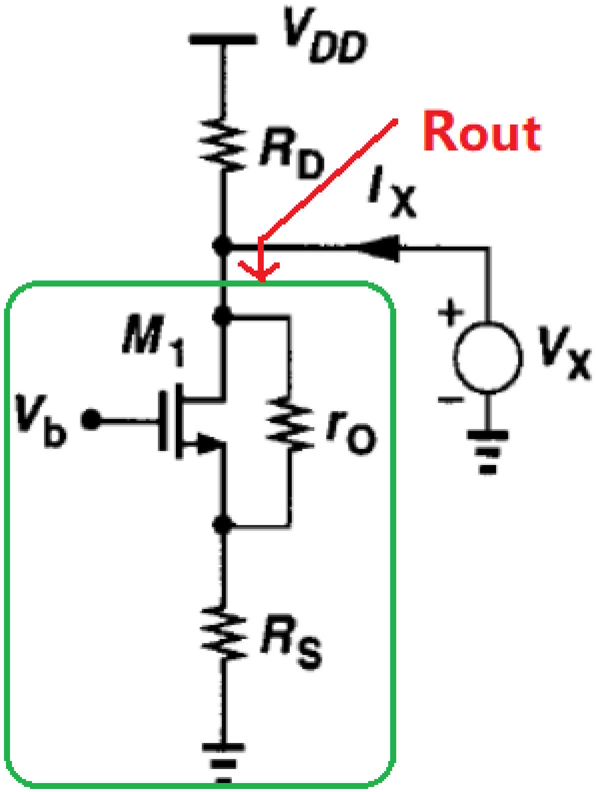
$R_D = \infty$ VRS恒定，交流信号与RS无关。

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

Output impedance of common-gate stage

与图3.29相同。



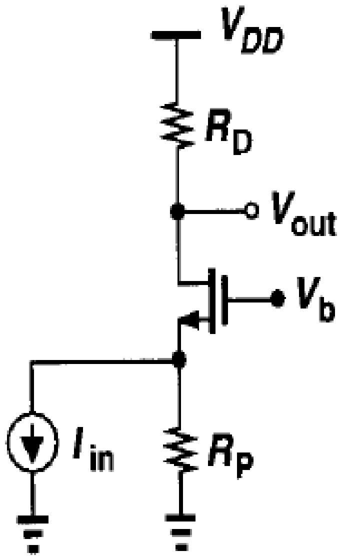


式(3. 66)： $R_{out} = [1 + (g_m + g_{mb})r_o]R_S + r_o$

Same as Degenerated CS

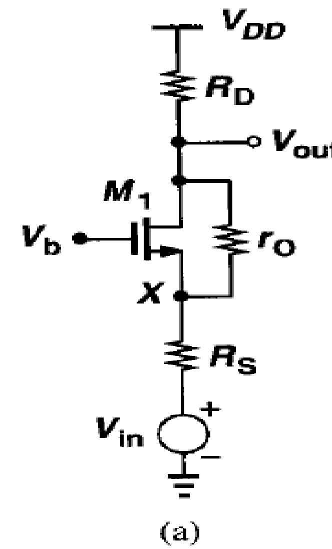
总的共栅CG ： $R'_{out} = \frac{V_X}{I_X} = \{[1 + (g_m + g_{mb})r_o]R_S + r_o\} \,||\, R_D$

例3.20 计算Vout/Iin和输出阻抗



已知右图，有式（3.111）：$A_v = \frac{V_{out}}{V_{in}}$

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

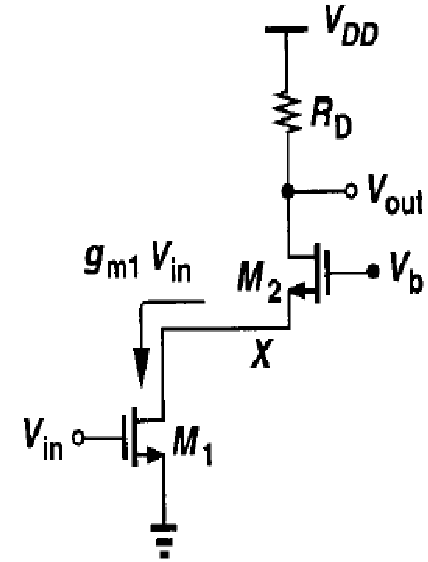


$$\therefore \frac{V_{out}}{I_{in}} = \frac{V_{out}}{V_{in}}\frac{V_{in}}{I_{in}} = \frac{r_o(g_m + g_{mb}) + 1}{r_o + (g_m + g_{mb})r_oR_P + R_P + R_D} R_D \times R_P \qquad (3.122)$$

$$总R_{out} = \{[1 + (g_m + g_{mb})r_o]R_P + r_o\} \,||\, R_D, \quad 令I_{in} = 0$$

3.6 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输入管和M2（cascode器件）工作在饱和区：

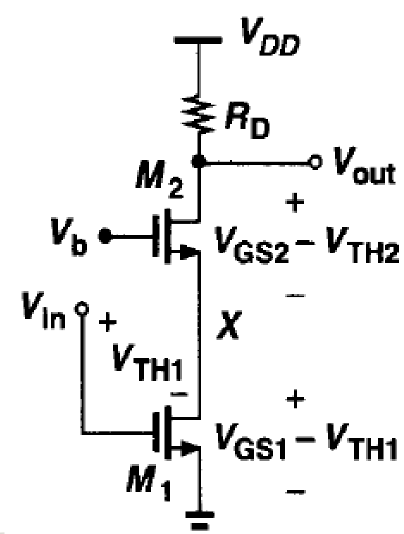
$$V_X = V_b - V_{GS2}\text{，且}V_X \geq V_{in} - V_{TH1}$$

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

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

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

即 $V_{out} \geq M1\text{过驱动电压} + \text{M2过驱动电压}$



直筒式CASCODE.
输入器件和Cascode器件是同一类型MOS

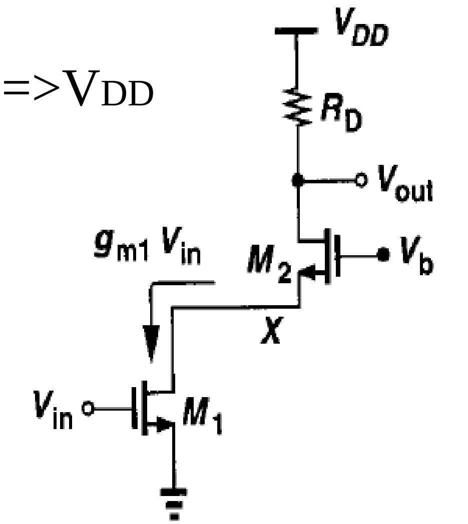
3.6.1 直筒式Cascode

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

$I_D = 0$， $V_{GS2} - V_{TH} = 0$

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

Vin从0=>VDD



（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

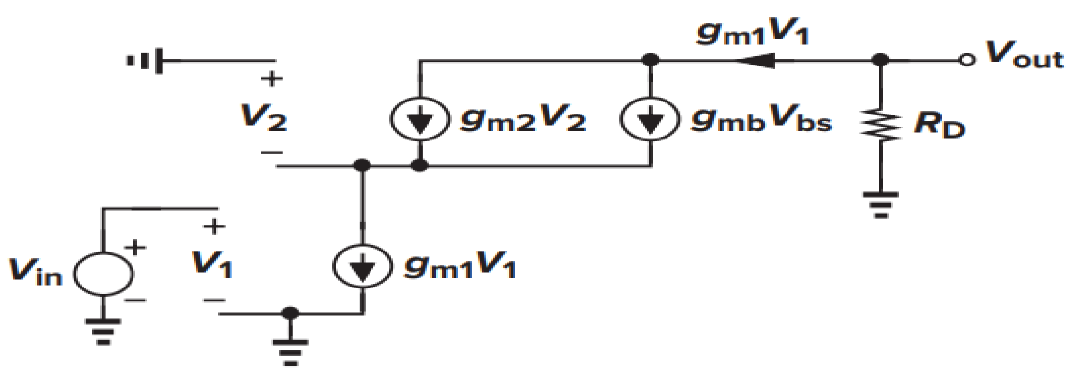
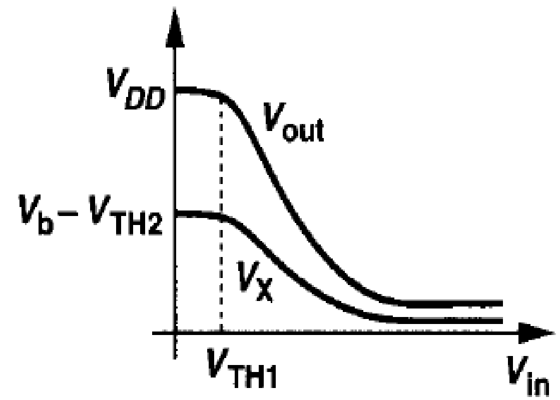


Figure 3.63 Small-signal equivalent circuit of cascode stage.



(3)Vin增加到很大时，M1或M 2 进入线性区。

若Vb比较小时，M1先进入线性区。

若M_2进入线性区，则$V_{out} \approx V_X$

例3.21 calculate the voltage gain

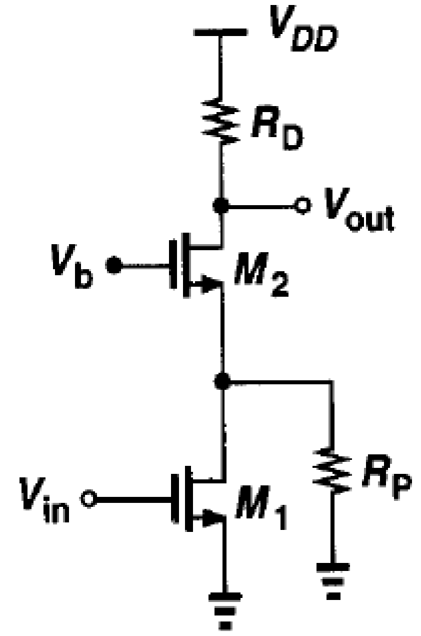
M1的小信号漏电流（受控电流源）$g_{m1}V_{in}$，被R_P（例如r_{o1}，或节点寄生电容）和向M2源极看进去的阻抗$1/(g_{m2}+g_{mb2})$分流。

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

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

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

V_{out}向下看的输出阻抗R_{out}很大



例3.22 理想电流源负载Cascode电压增益

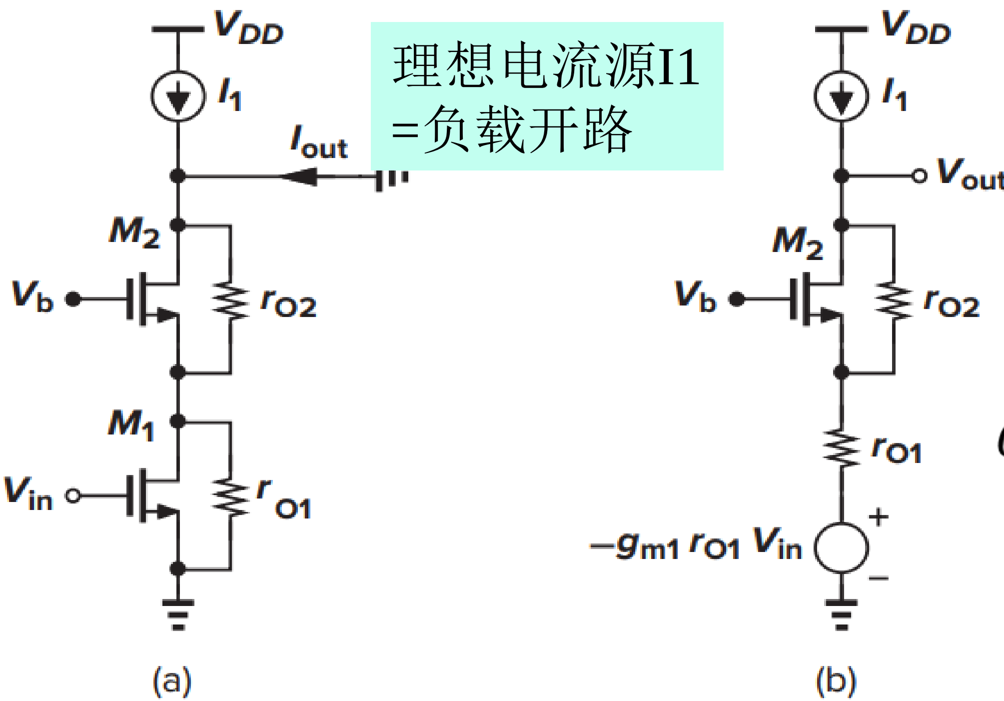


Figure 3.68

理想电流源I1 =负载开路

回顾辅助定理：

开路电压增益=-GmRout，其中Gm表示输出与地短接时的电路跨导=Iout/Vin，Rout为输出电阻

$$G_m = \frac{I_{out}}{V_{in}} = \frac{g_{m1}r_{o1}}{r_{o1}+\frac{1}{g_{m2}+g_{mb2}}||r_{o2}}$$

$$= \frac{g_{m1}r_{o1}}{r_{o1}+\frac{r_{o2}}{1+(g_{m2}+g_{mb2})r_{o2}}}$$

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

$$G_m \approx g_{m1}$$

$$I_{out} = g_{m1}V_{in}\frac{r_{o1}}{r_{o1}+\frac{1}{g_{m2}+g_{mb2}}||r_{o2}}$$

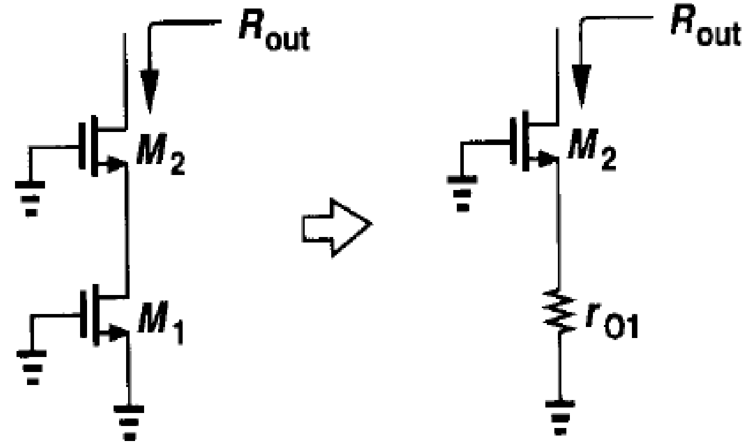
$$A_v = -G_mR_{out} \approx -g_m * r_{o1}[1+(g_{m2}+g_{mb2})r_{o2}] \approx -g_m(g_{m2}+g_{mb2})r_{o1}r_{o2}$$

cascode结构的重要特性：输出阻抗大

由式（3.66）：

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

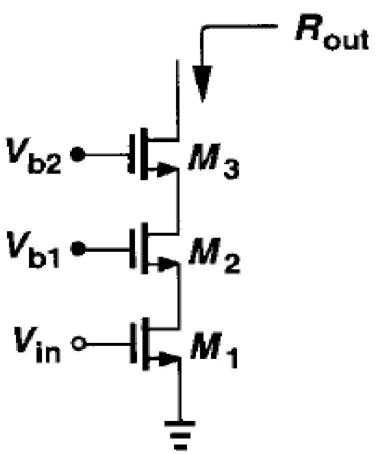
$$\approx (g_{m2} + g_{mb2})r_{o2}r_{o1}$$



将共源M_1输出电阻r_{o1}提高到约

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

前提：$M1$和$M2$均在饱和区



将共源M_1输出电阻r_{o1}提高到约

$(g_{m3} + g_{mb3})r_{o3} \times (g_{m2} + g_{mb2})r_{o2} \times r_{o1}$

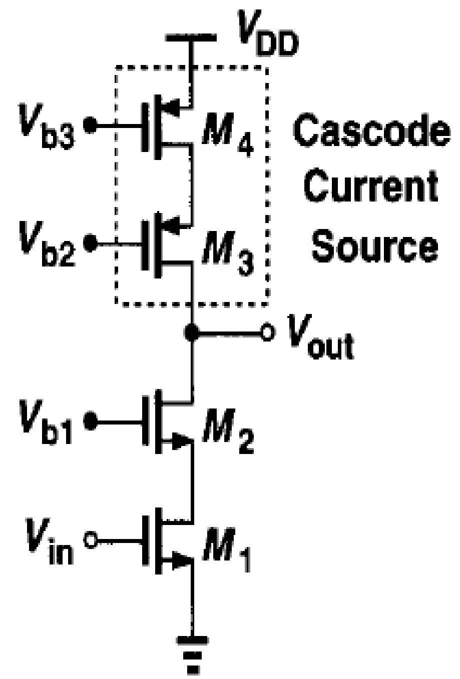
现代工艺电源电压较低，每个MOS管需要V_{DS}，故此结构难以应用

放大电路要上下阻抗对称

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

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

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



直流输出电平需另外电路确定

直流输入电压动态范围小

$$V_{in} < V_{b1} - V_{GS2} + V_{TH1}$$
$$= V_{b1} - (V_{OD2} + V_{TH2}) + V_{TH1} \approx V_{b1} - V_{OD2}$$

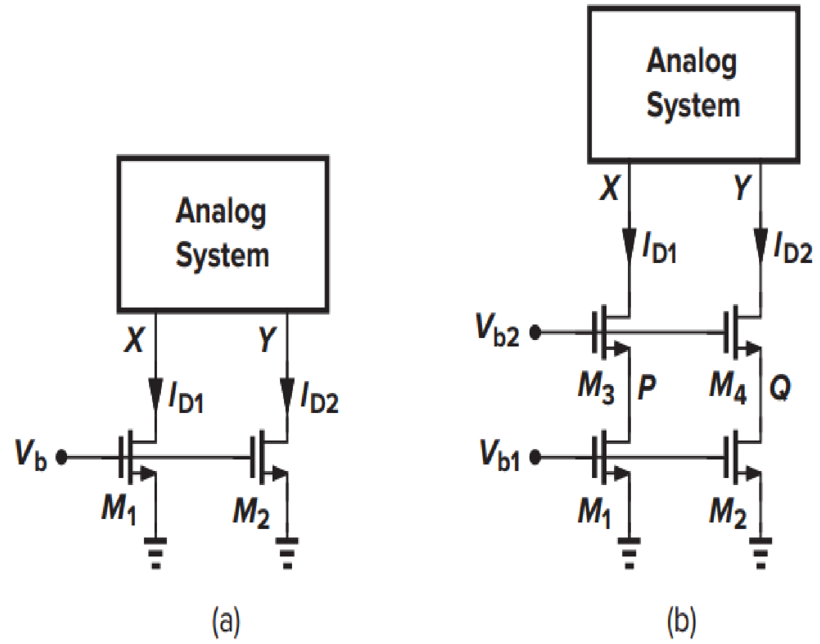
Shielding property

设M_1M_2相同，M_3M_4相同，都是饱和区。

图（a）：$V_X - V_Y = \Delta V$

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

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



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

3.6.2 折叠式共源共栅 Folded cascode

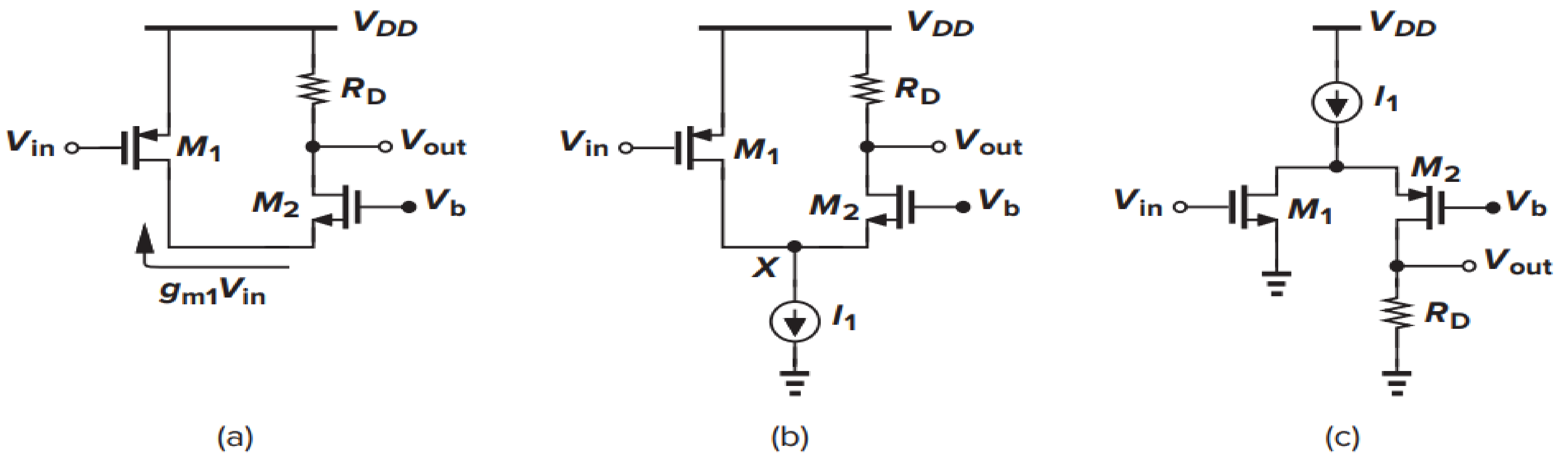
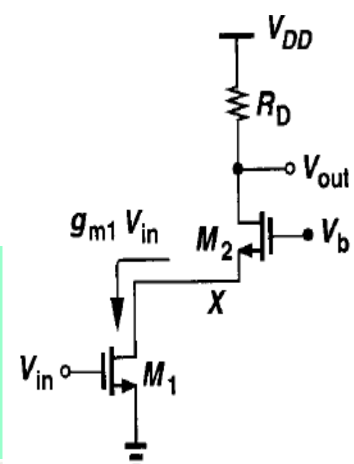


Figure 3.74 (a) Simple folded cascode; (b) folded cascode with proper biasing; (c) folded cascode with NMOS input.

图3.74 (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



折叠式输入器件和cascode器件不是同一类型。

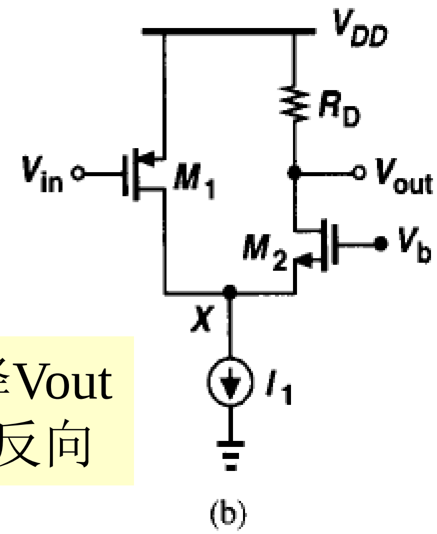
由于I1电流源，若I_{D1}增大则I_{D2}减小。

折叠式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



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

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

Vin下降Vout升高，反向

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

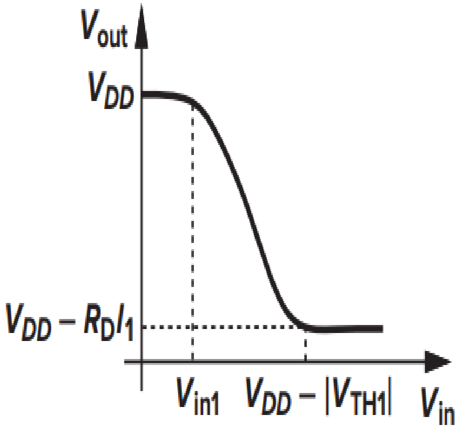
$$= V_{DD} - I_1 R_D + \frac{1}{2}\mu_p C_{ox}\left(\frac{W}{L}\right)_1 (V_{DD} - V_{in} - |V_{TH1}|)^2 R_D$$

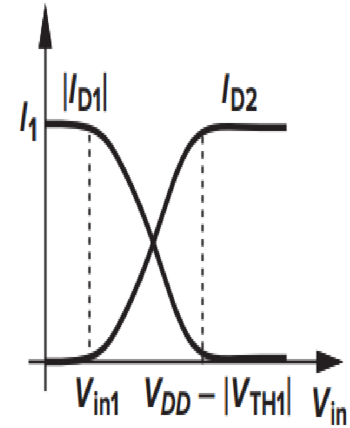
$$I_{D1} + I_{D2} = I_1$$

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





折叠管的输出阻抗

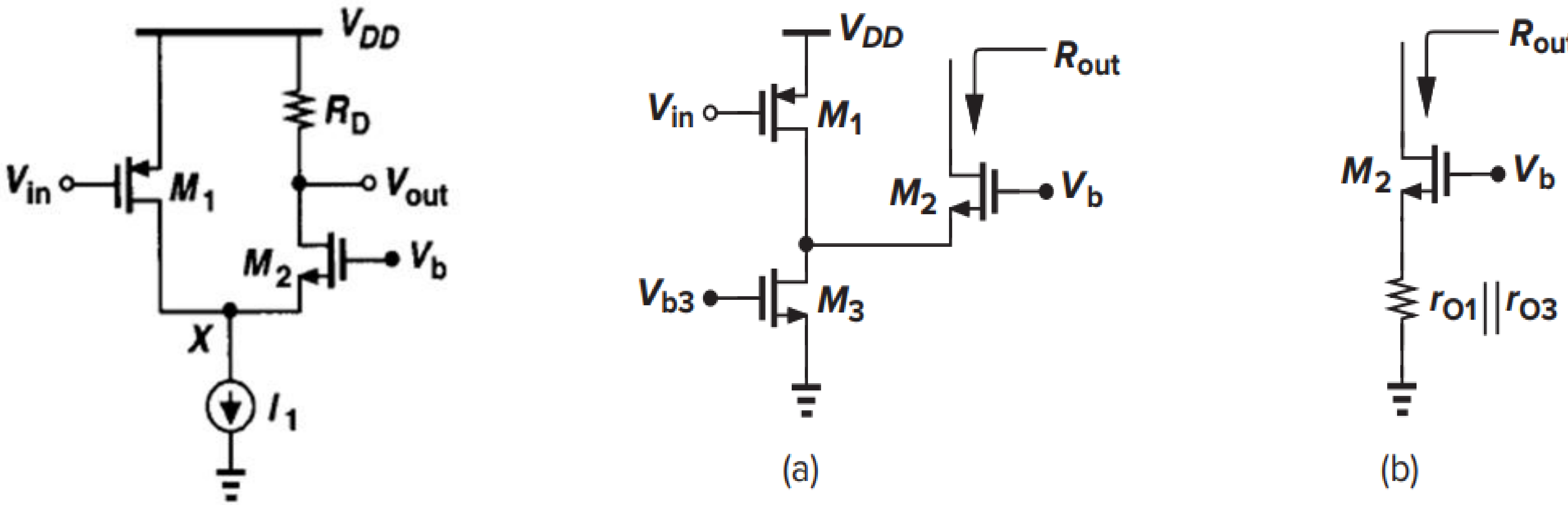
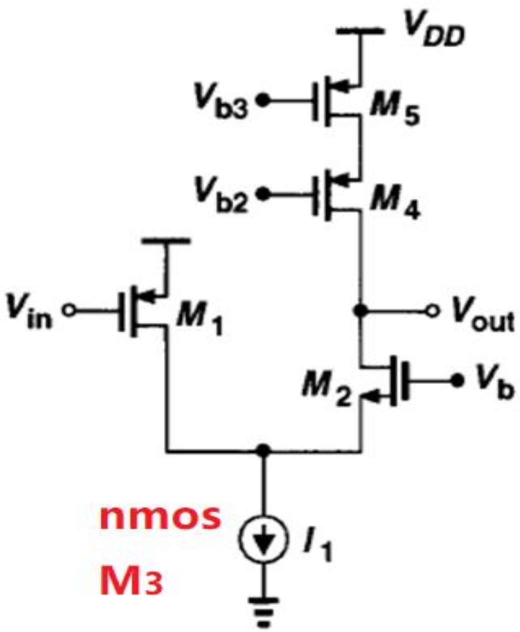


Figure 3.76

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

手工计算简化规则：求和计算可忽略相差一个数量级以上的参数

The load of a folded cascode



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

- Four types of amplifiers（MOSFET的小信号输入输出公共端）
- 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
- 掌握电路分析方法：

  * Large-signal characteristics：计算输入输出范围

  * Small-signal characteristics (Low frequency behavior)：列出节点电路方程，计算低频增益

  * 忽略小分量的手工近似计算方法