Amplifiers, Source followers & Cascodes

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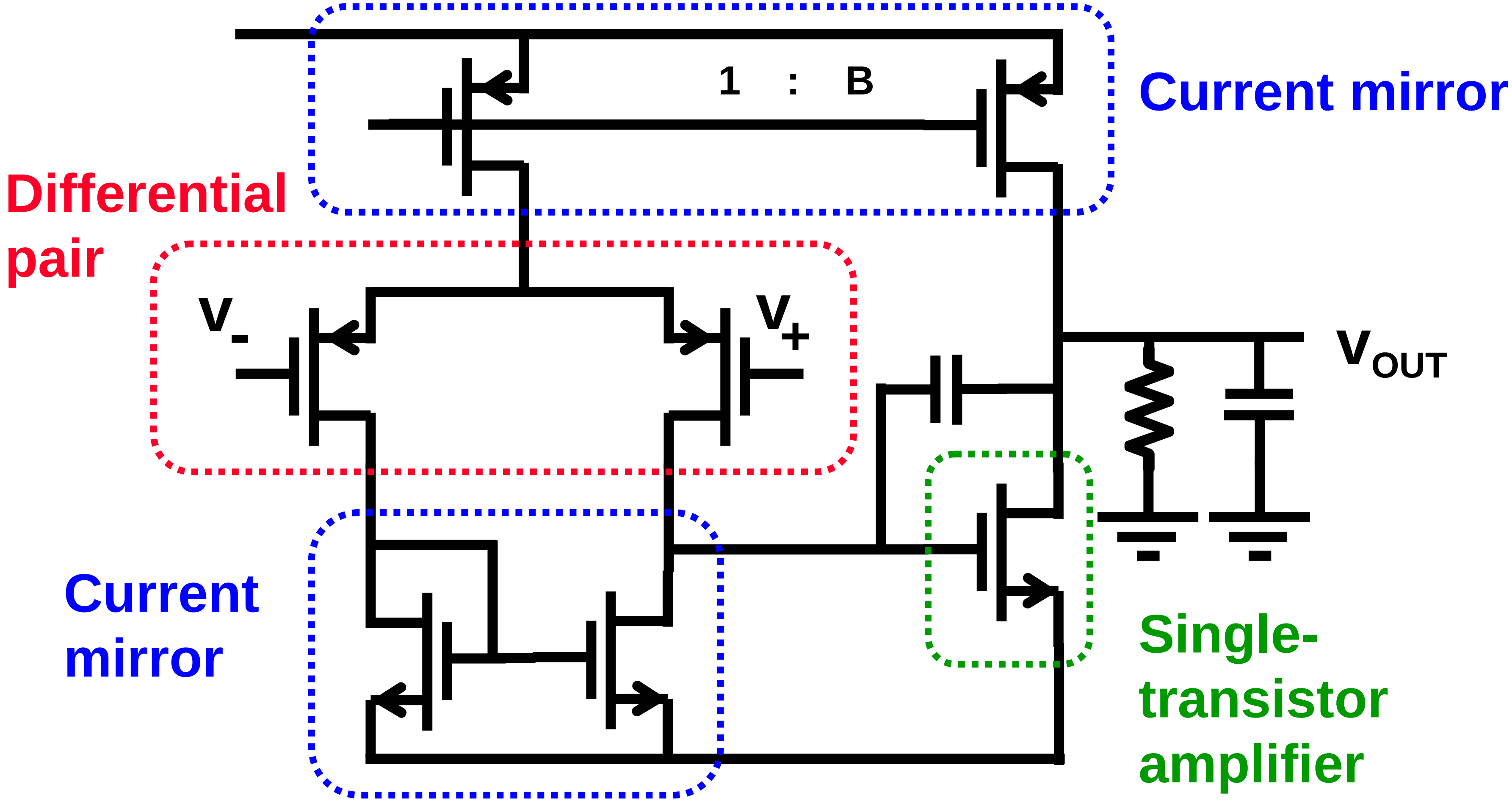
Operational amplifier



电路分析：电路结构

- 共源放大器
  - 电阻负载
  - 理想电流源
  - MOS管电流源
  - 源极负反馈
  - 二极管负载
  - 推挽结构
- 源极跟随器
- 共栅放大器
- 共源共栅放大器
- 衬底输入结构*

电路分析：电路指标

- 直流指标
  - 静态功耗
  - 输入电压范围
  - 输出电压范围
- 低频交流指标
  - 低频（直流）增益
  - 输入电阻
  - 输出电阻
- 高频交流指标
  - 带宽
  - 增益带宽积
  - 零极点
  - 输入阻抗
  - 输出阻抗

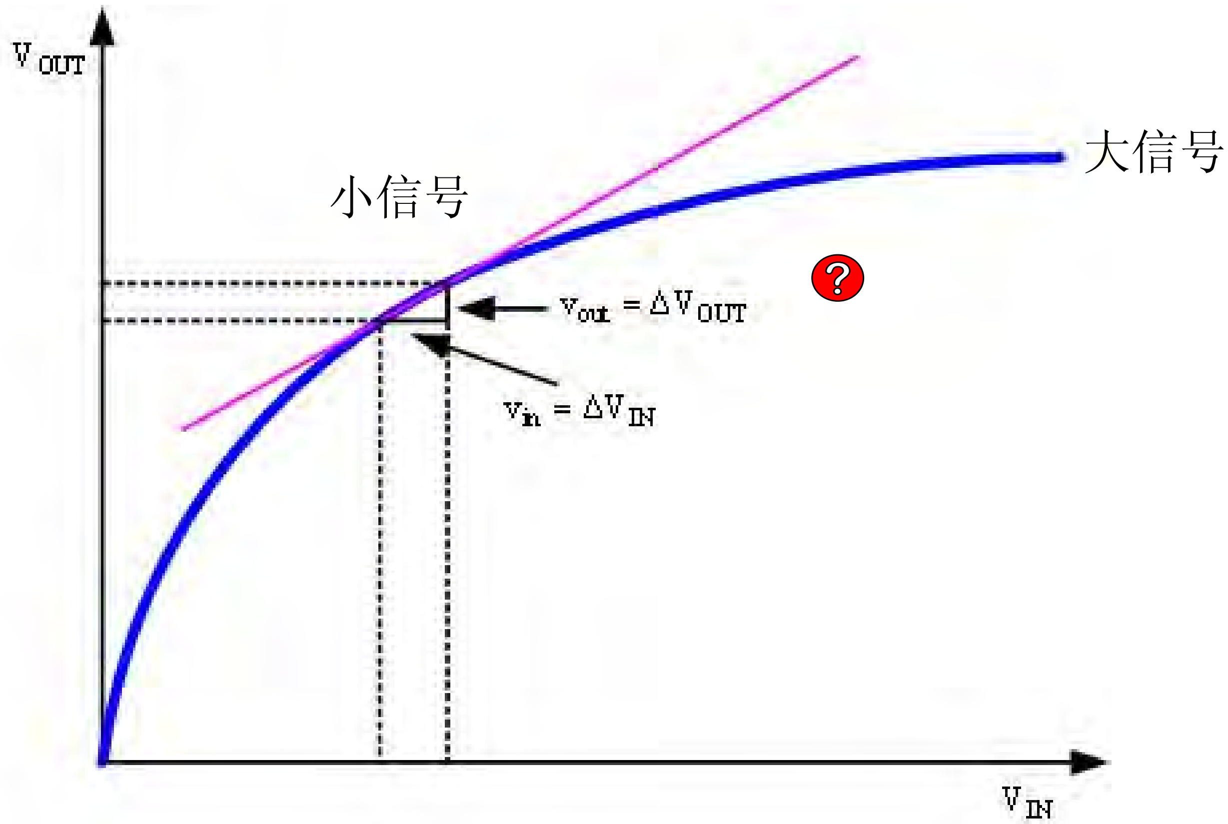
电路分析：分析方法

定量分析	理论推导	直观分析
直流分析	列直流方程组 解直流工作点	输入、输出电压范围 （MOS管工作在饱和区）
低频交流分析	低频小信号等效电路图 公式推导增益、电阻	输出电阻=上看下看 增益=跨导×输出电阻
高频交流分析	高频小信号等效电路图 公式推导BW、GBW、PZ、阻抗 辅助工具：波特图 特殊结论：miller效应	极点和节点对应 极点频率=节点RC BW~主极点频率 GBW~跨导/主电容

定性分析：了解电路的工作原理

	输入电压范围	输出电压范围	直流增益	BW	GBW	零极点分布	输入阻抗	输出阻抗	电路特点
共源电流源									
源极负反馈									
二极管负载									
推挽									
源极跟随器									
共栅									
共源共栅									

大信号分析和小信号分析



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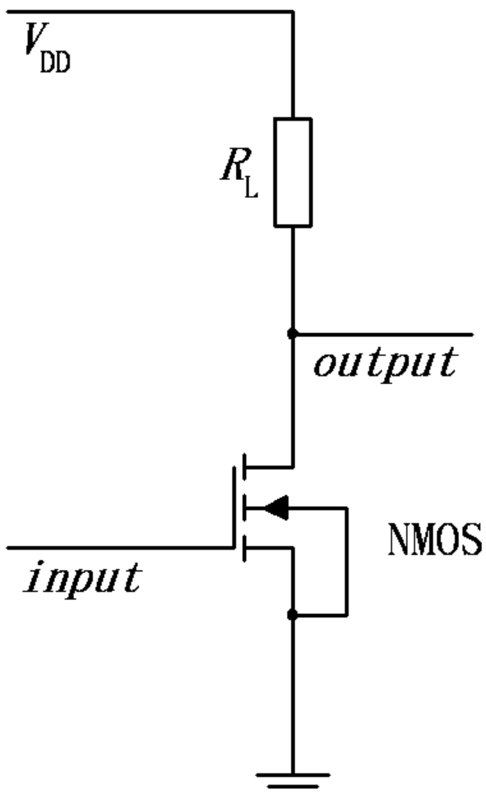
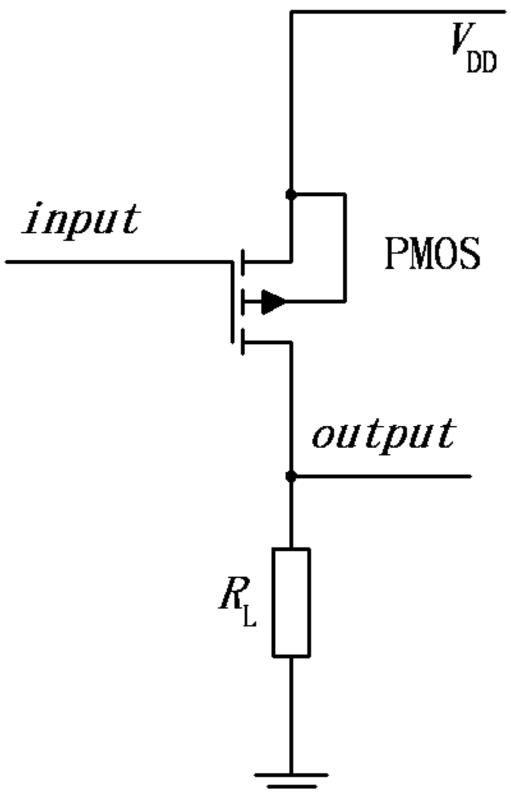
- ❑ Cascodes

共源放大器

电阻负载

电路结构





直流指标1-直流工作点-列方程组

$$\frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(V_{IN}-V_{TH}\right)^2\left(1+\lambda V_{OUT}\right)=\frac{V_{DD}-V_{OUT}}{R_L}$$

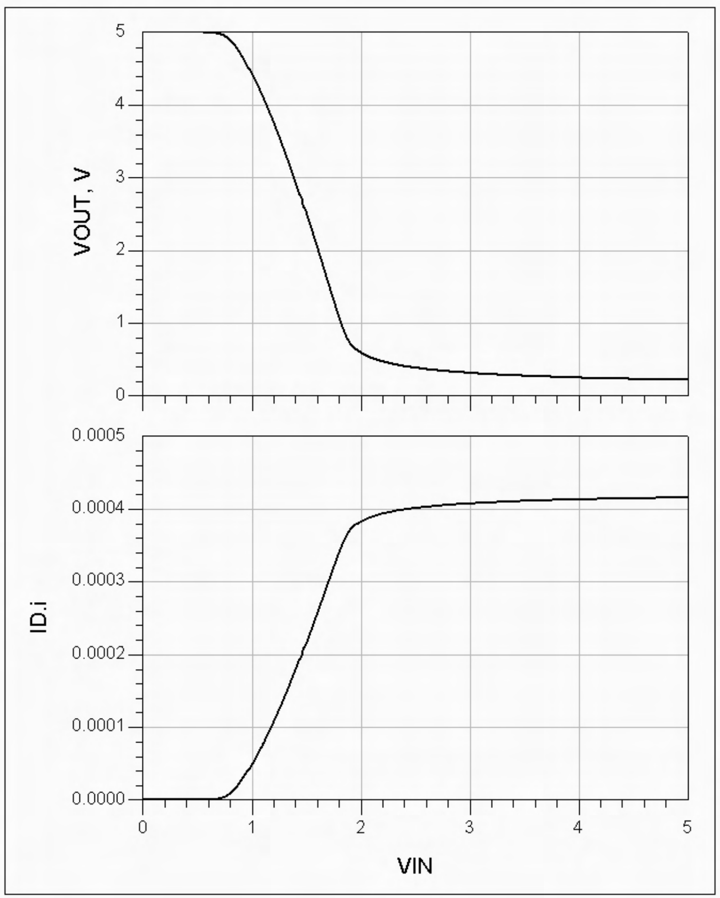
$$V_{OUT}=\frac{V_{DD}-V_R}{1+\lambda V_R}$$

$$I_D=\frac{V_R}{R}\frac{1+\lambda V_{DD}}{1+\lambda V_R}$$

$$V_R=\frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(V_{IN}-V_{TH}\right)^2 R_L$$

- 直流工作点状况
  - 各点电压
  - 各支路电流
  - 静态功耗
  - 小信号参数
    - 跨导、交流输出电阻…

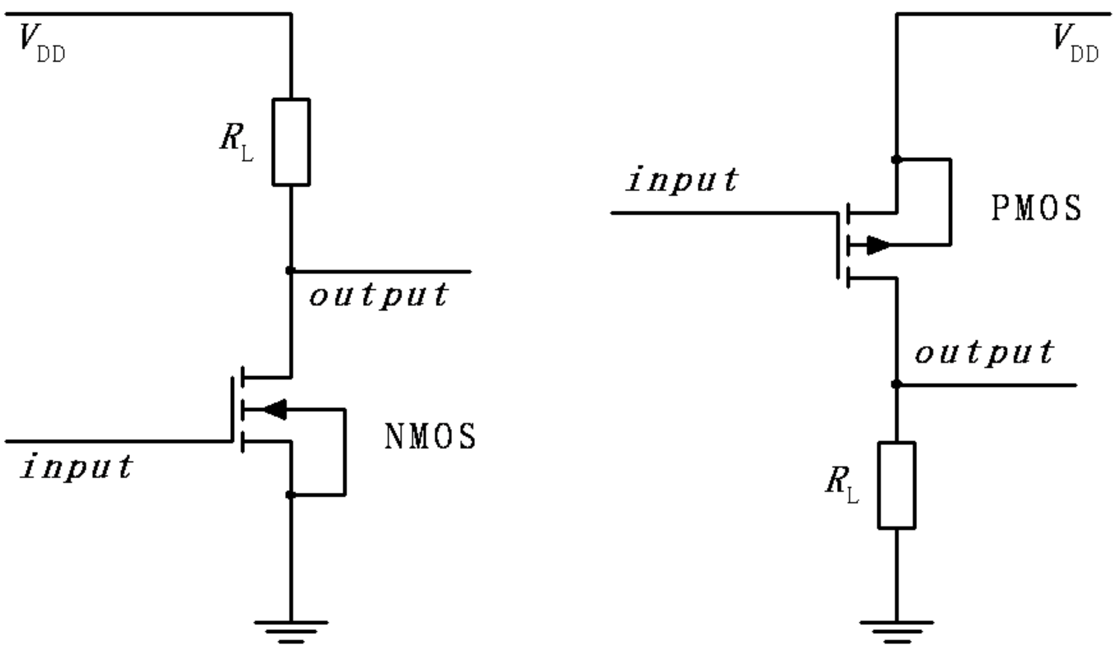
直流工作点扩展-直流扫描



	MOS管状态	输出电压	沟道电流
$V_{IN} < V_{TH}$	截止	V_{DD}	0
$V_{IN} \geq V_{TH}$	饱和	下降	上升
$V_{IN} - V_{TH} \geq V_{OUT}$	线性	→ 0	缓升，到最大

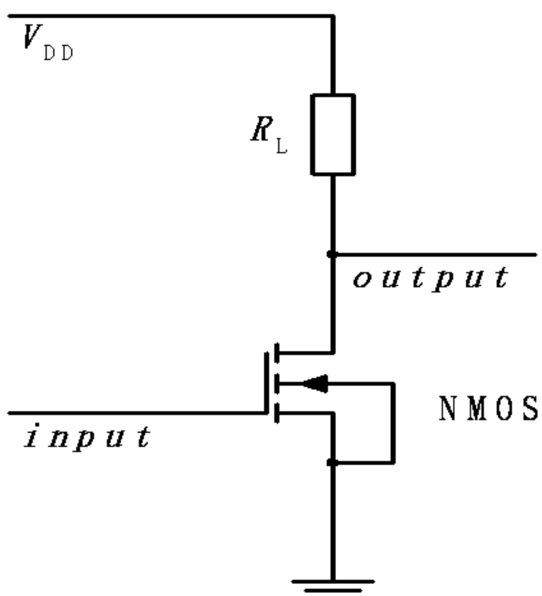
直流指标2-输入电压范围-直观分析

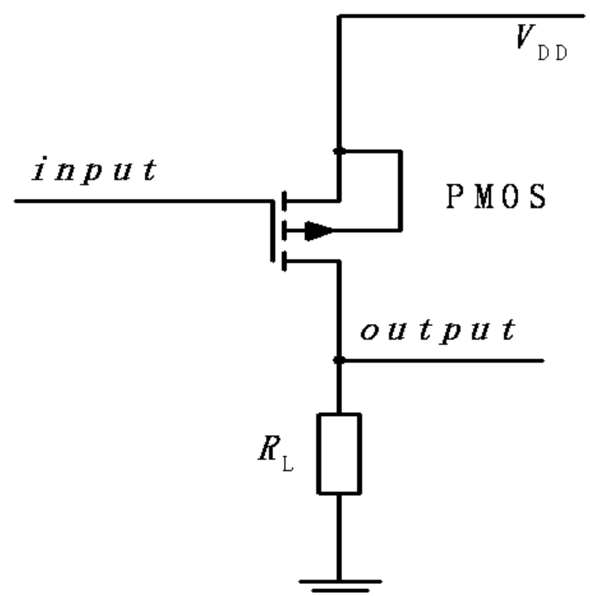
- 大于阈值电压
- 大于阈值电压200mV



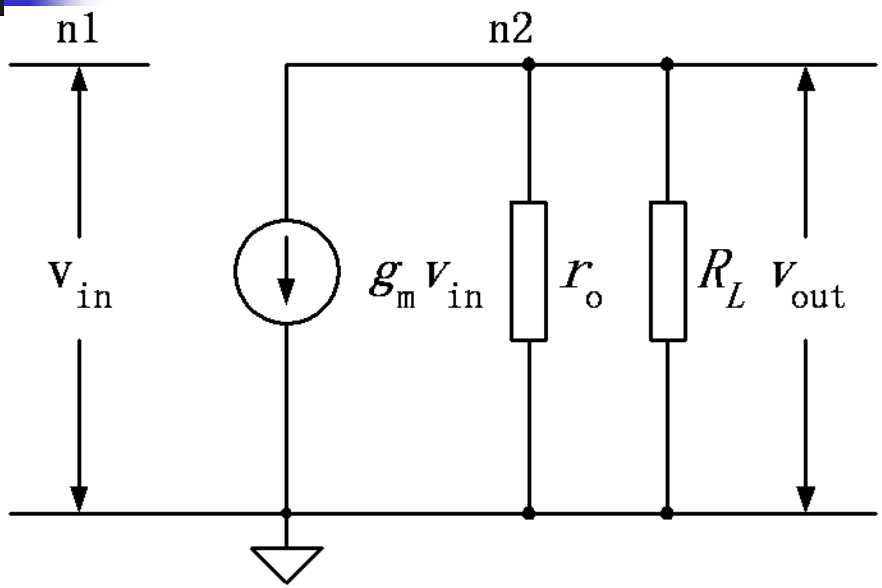
直流指标3-输出电压范围-直观分析

- 向上受到电源电压的限制
- 向下受到MOS管最小漏源电压的限制
- $V_{IN}-V_{TH} \sim V_{DD}$



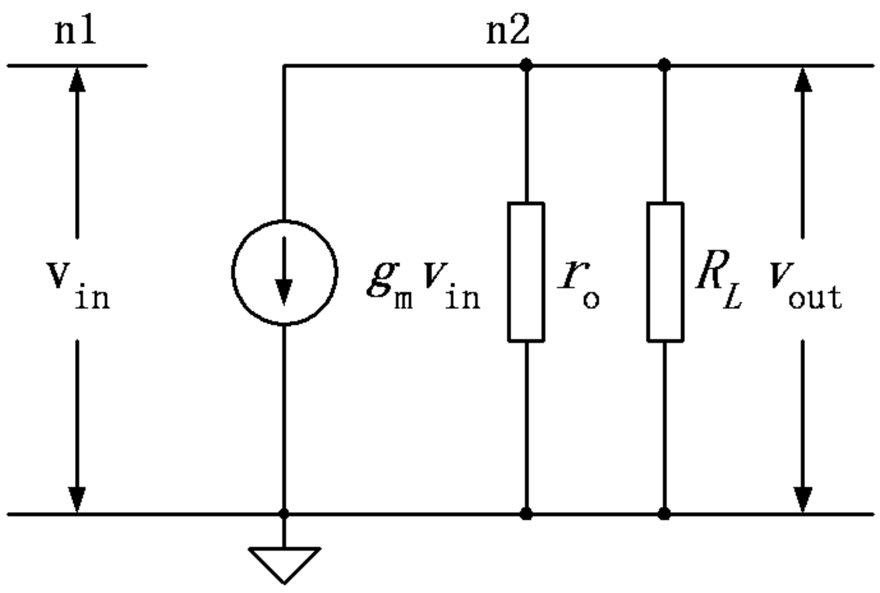


低频交流指标1-直流增益-低频小信号等效电路图



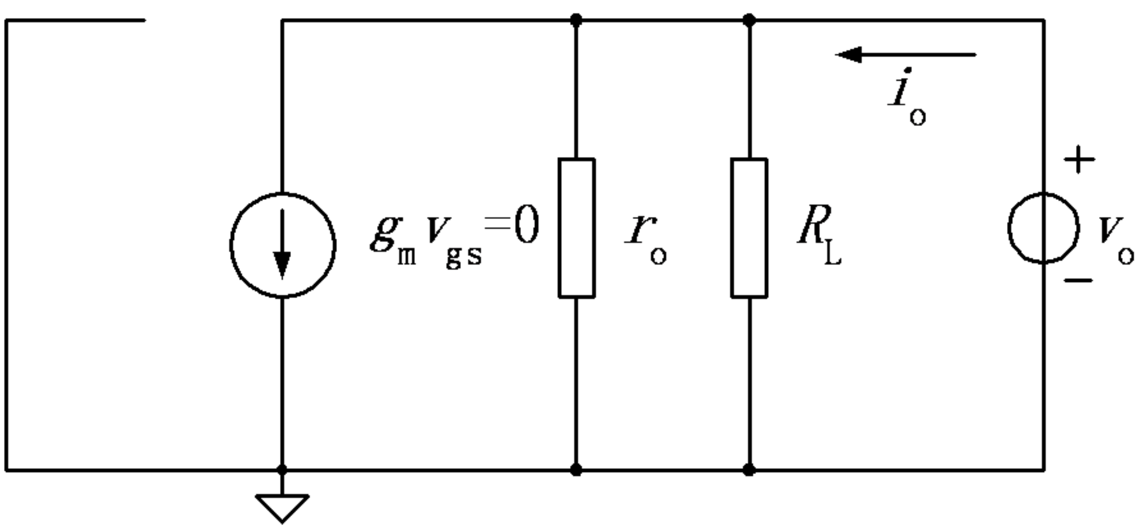
$$\boxed{\frac{v_{out}}{v_{in}} = -g_m\left(r_o // R_L\right)}$$

低频交流指标2-输入电阻-低频小信号等效电路图



- 求输入电阻时，负载开路
- 输入电阻：无穷大

低频交流指标3-输出电阻-低频小信号等效电路图



- 求输出电阻时，去掉信号源，保留内阻
- 输出电阻：$\boxed{r_o // R_L}$

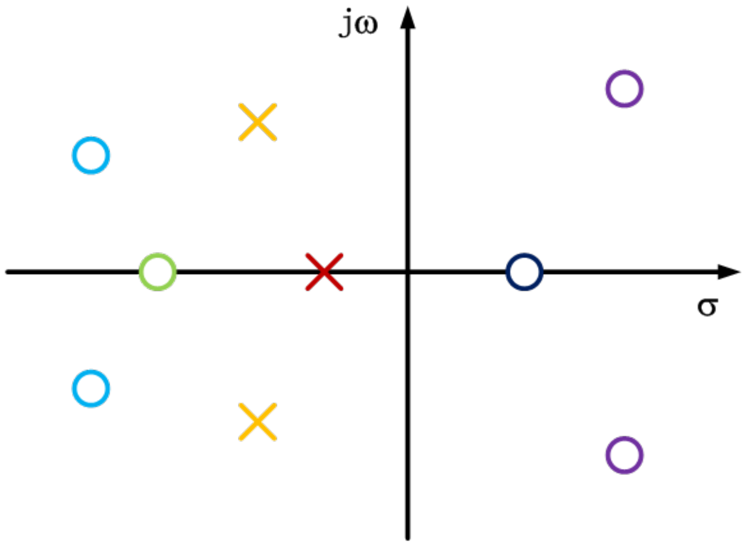
低频交流指标的直观分析

- 输出电阻
  - 向上看：R_L
  - 向下看：r_o
- 输入电阻
- 直流增益
  - 输入电压—输出电流—输出电压
  - 输入电压—输出电流：跨导g_m
  - 输出电流—输出电压：电阻$r_o//R_L$
  - $\boxed{\text{增益：跨导×输出电阻}}$

补充：传输函数

$$H(s)=\frac{N(s)}{D(s)}=\frac{a_m s^m+a_{m-1}s^{m-1}+\ldots+a_0}{b_n s^n+b_{n-1}s^{n-1}+\ldots+b_0}$$

$$H(s)=\frac{a_0}{b_0}\frac{(1-\frac{s}{z_1})(1-\frac{s}{z_2})\ldots(1-\frac{s}{z_m})}{(1-\frac{s}{p_1})(1-\frac{s}{p_2})\ldots(1-\frac{s}{p_n})}$$



参考：信号与系统（第二版），郑君里，第**4**章！

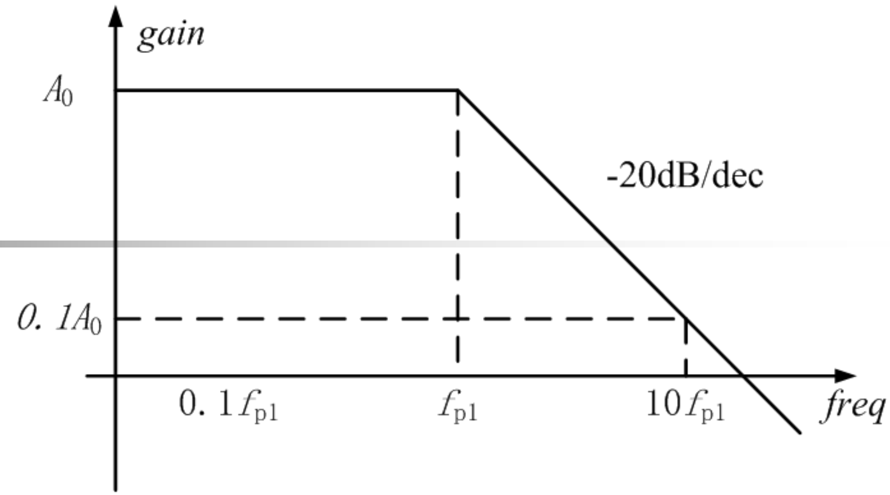
补充：极点

- 幅频特性
- 相频特性
  - 左半平面
  - 右半平面

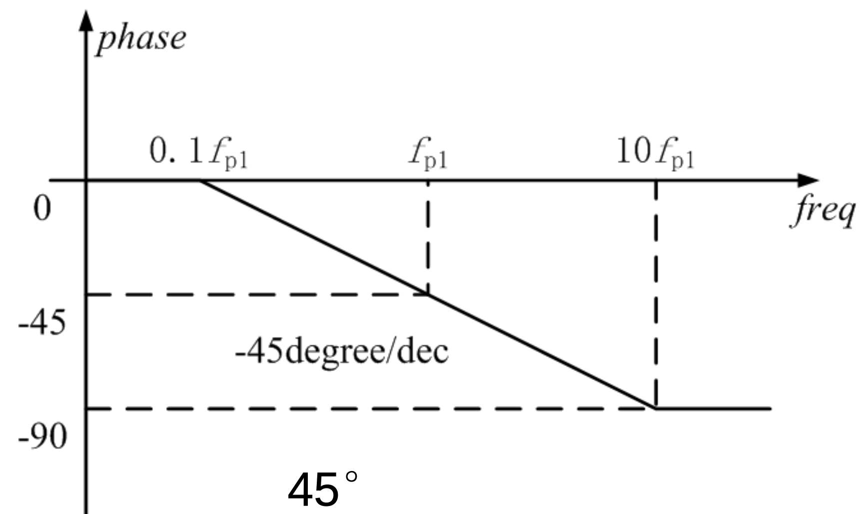
$$A(s)=\frac{A_0}{1-\frac{s}{p_1}}$$

$$|A(j\omega)|=\frac{A_0}{\sqrt{1+\left(\frac{\omega}{\omega_{p1}}\right)^2}}$$

$$\varphi(j\omega)=\varphi_0-\arctan\left(\omega/\omega_{p1}\right)$$



3dB



45°

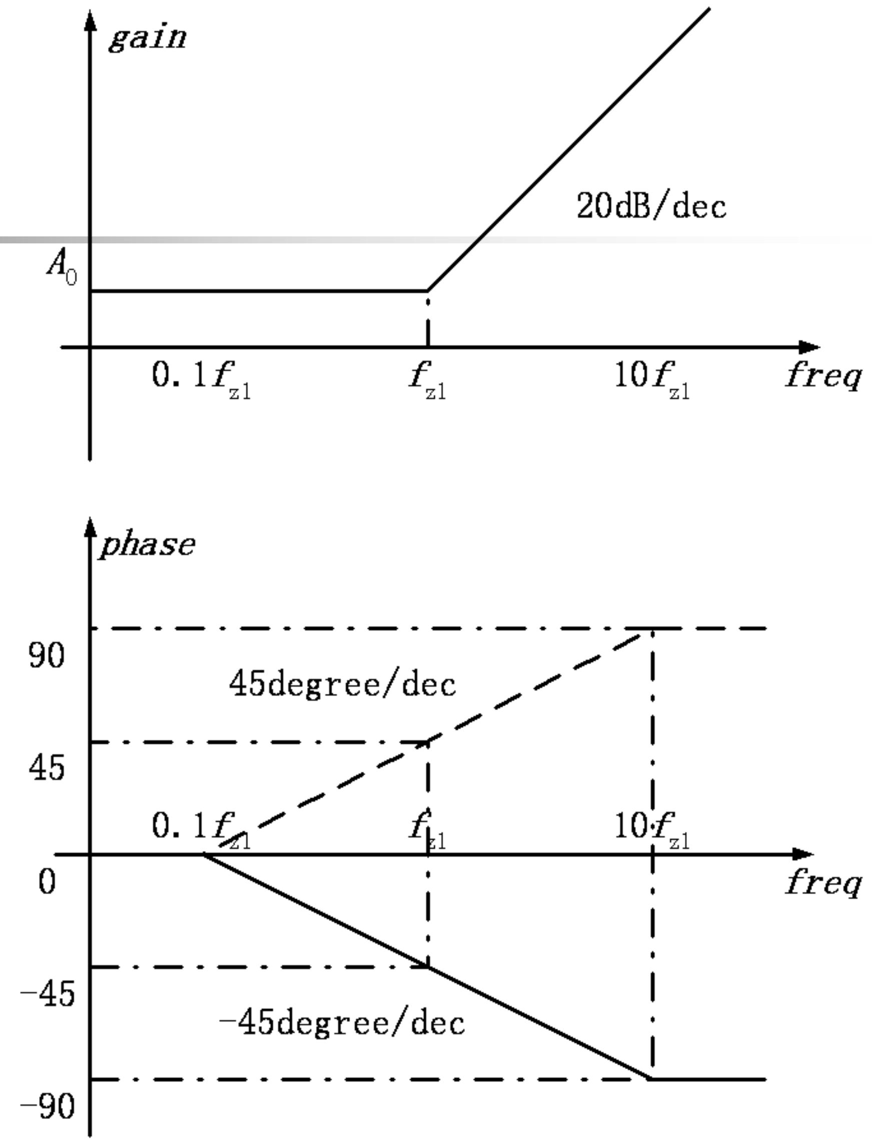
补充：零点

- 幅频特性
- 相频特性
  - 左半平面
  - 右半平面

$$A(s)=A_0\left(1-\frac{s}{z_1}\right)$$

$$|A(j\omega)|=A_0\sqrt{1+\left(\frac{\omega}{\omega_{z1}}\right)^2}$$

$$\varphi(j\omega)=\varphi_0-\arctan\left(\omega/\omega_{z1}\right)$$

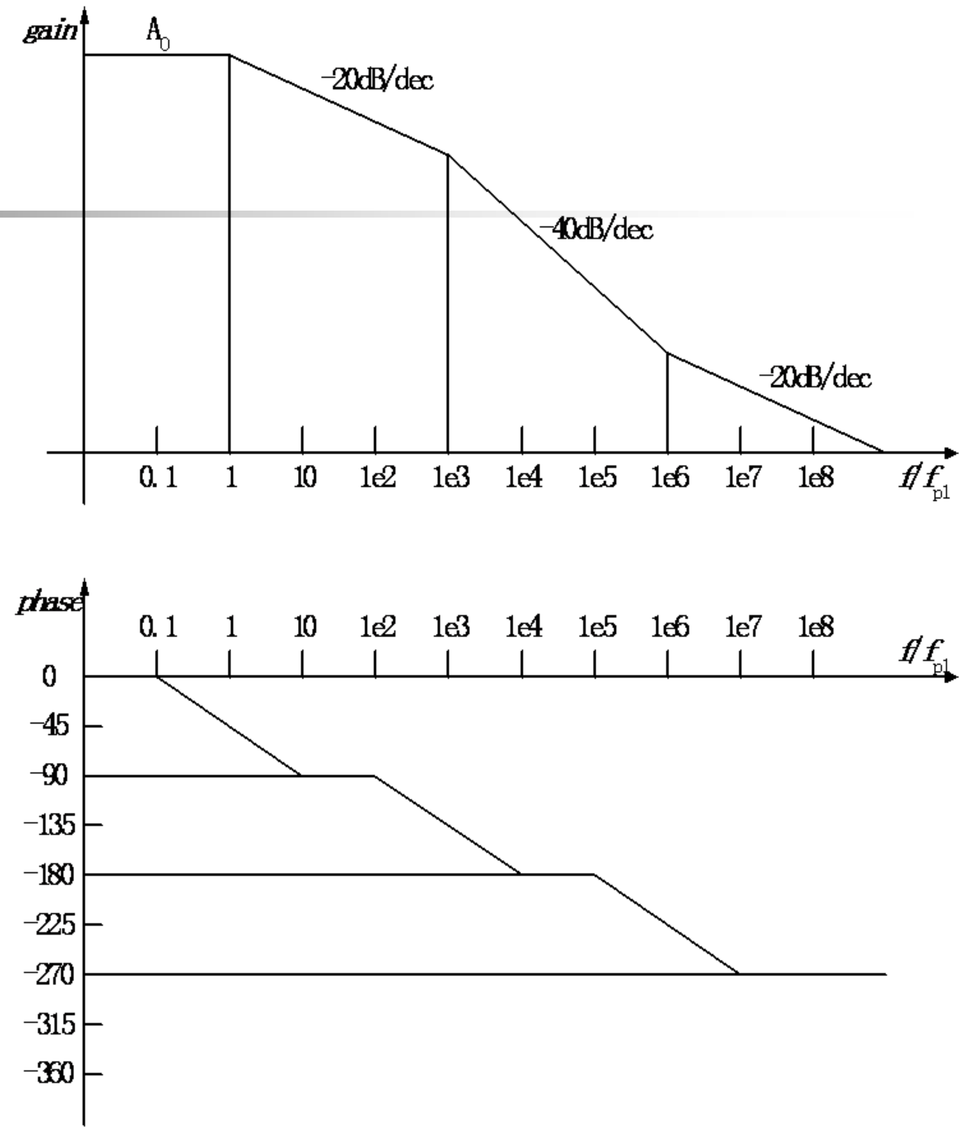


补充：波特图

- 幅频响应
  - 横坐标：频率，log表示
  - 纵坐标：应变量，dB表示
    - 信号（电压、电流等）
    - 信号的比值
    - 其它和频率相关的参数（阻抗等）
- 相频响应
  - 横坐标：频率，log表示
  - 纵坐标：角度，线性表示

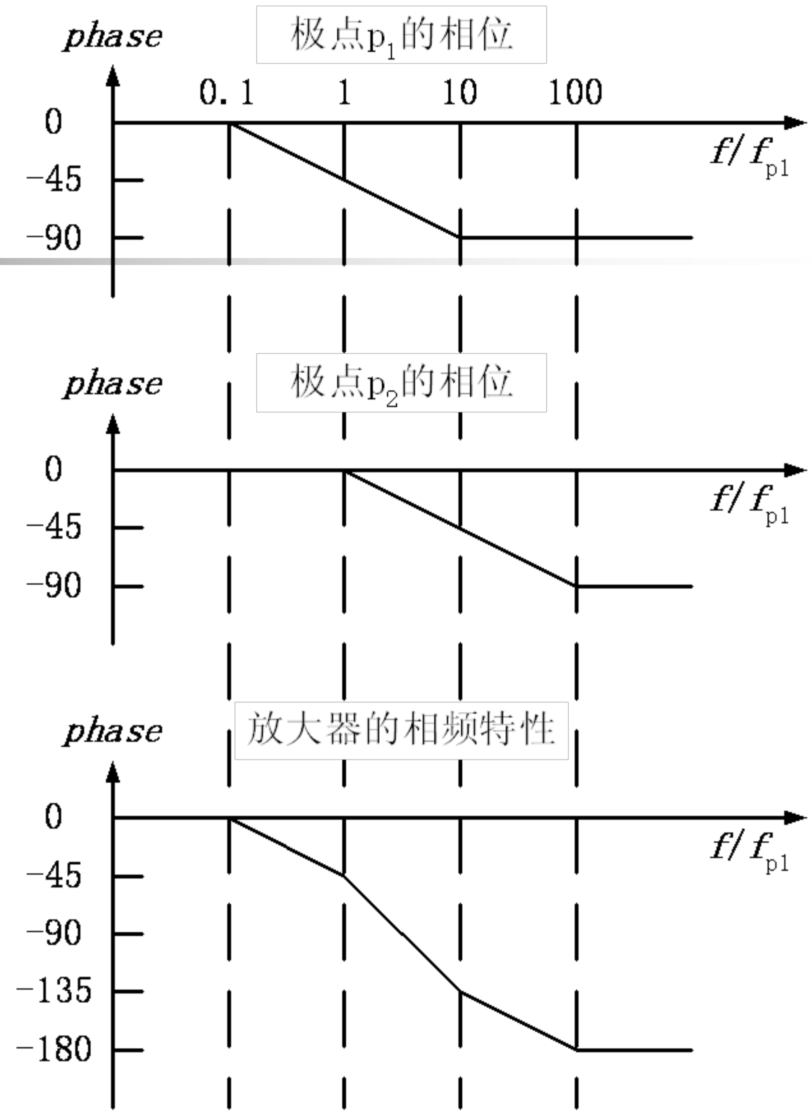
波特图-例1

- f_{p1}、 f_{p2}、 f_{z1}
- $f_{p2}=1000f_{p1}$
- $f_{z1}=1000f_{p2}$

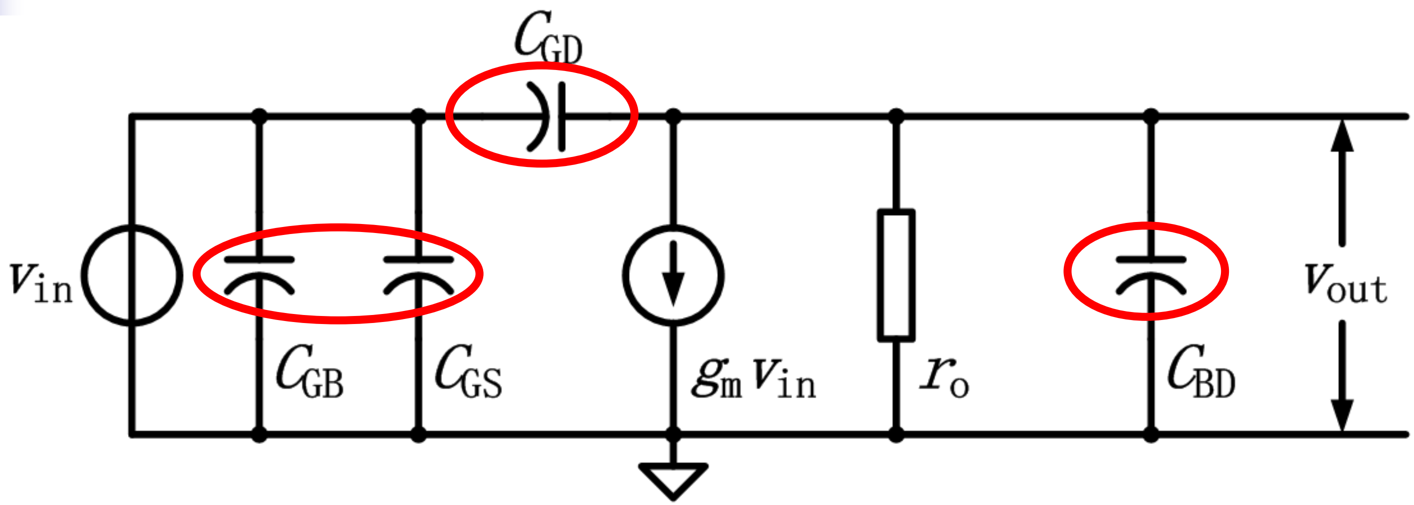


波特图-例2

- f_{p1}、 f_{p2}
- $f_{p2}=10f_{p1}$

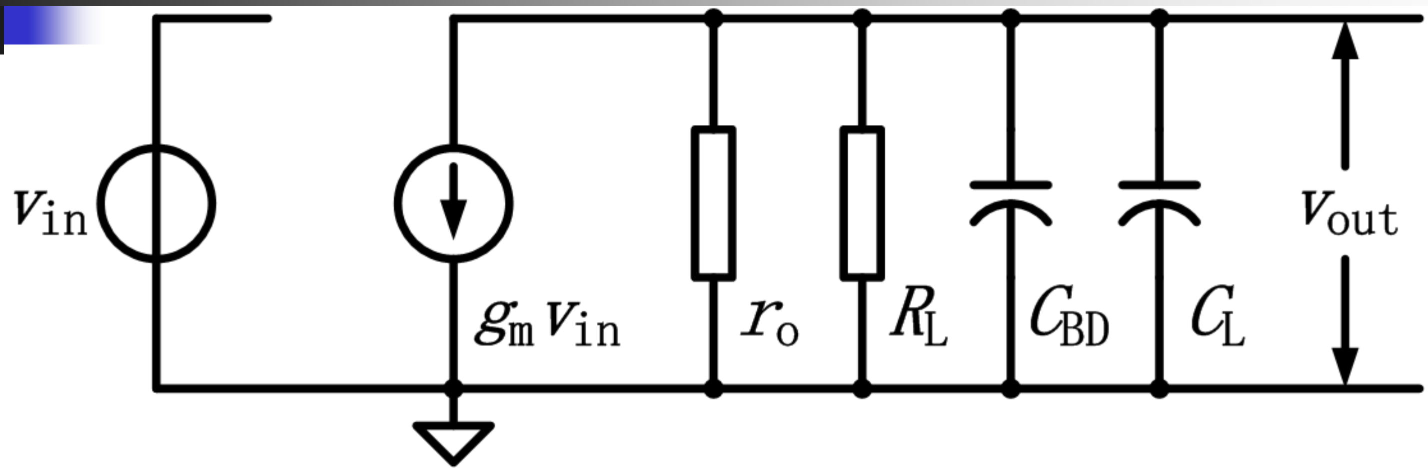


MOS管的高频小信号等效电路



- 不同位置电容的意义
  - 前级负载电容
  - 本级负载电容
  - 输入-输出之间的耦合电容
- 考虑不同的电容时会出现不同的零极点分布

高频交流指标1-3dB带宽-高频小信号等效电路图



- 3dB带宽：
  - -3dB带宽
  - 比直流增益下降3dB时对应的频率

$$A(s)=\frac{A_0}{1-\frac{s}{p_1}}=-\frac{g_m\left(R_L//r_o\right)}{1-\frac{s}{\left[-\frac{1}{\left(R_L//r_o\right)//\left(C_{BD}+C_L\right)}\right]}}$$

更多说明

负号代表输入输出反相

直流增益

$$A(s)=\frac{A_0}{1-\frac{s}{p_1}}=-\frac{g_m\left(R_L//r_o\right)}{1-\frac{s}{\left[-\frac{1}{\left(R_L//r_o\right)//\left(C_{BD}+C_L\right)}\right]}}$$

负号代表极点在s平面负实轴上

$$A(s)=\frac{A_0}{1+\frac{s}{p_1}}=-\frac{g_m\left(R_L//r_o\right)}{1+\frac{s}{\left[\frac{1}{\left(R_L//r_o\right)//\left(C_{BD}+C_L\right)}\right]}}$$

极点的大小正好等于输出节点“到交流地”的电阻和电容的乘积

单极点系统中：极点频率也是3dB带宽的频率

高频交流指标2-增益带宽积-公式计算

- 增益

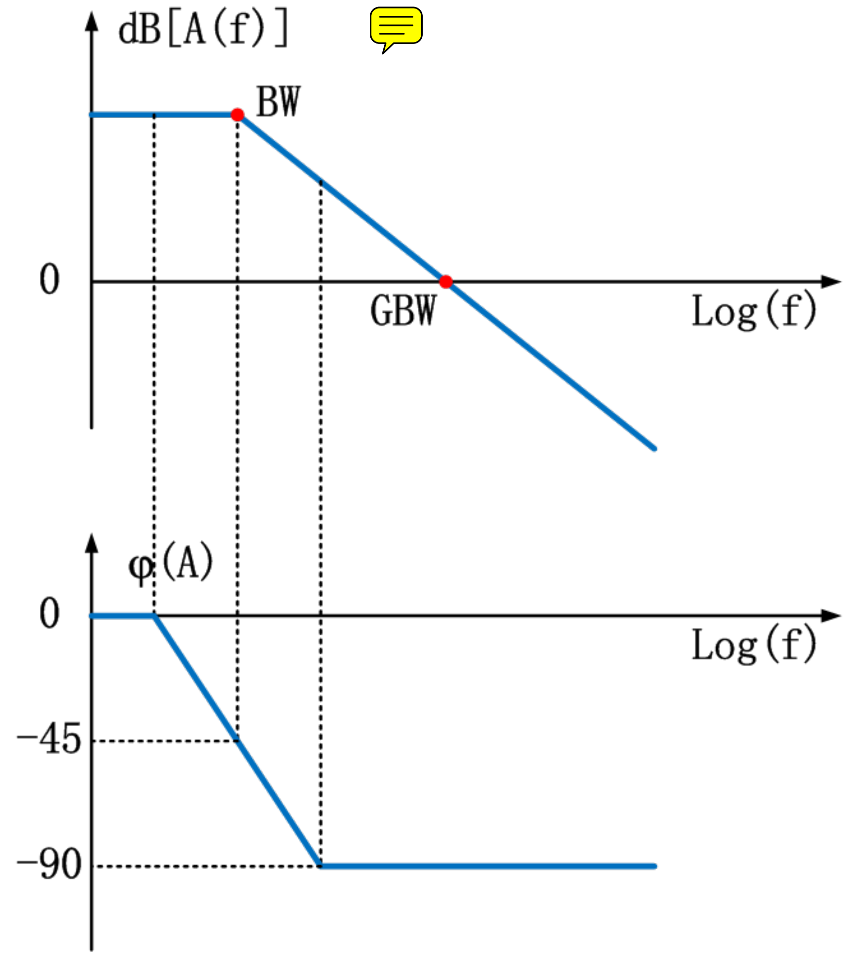
$$A_0 = g_m\left(R_L//r_o\right)$$

- 带宽

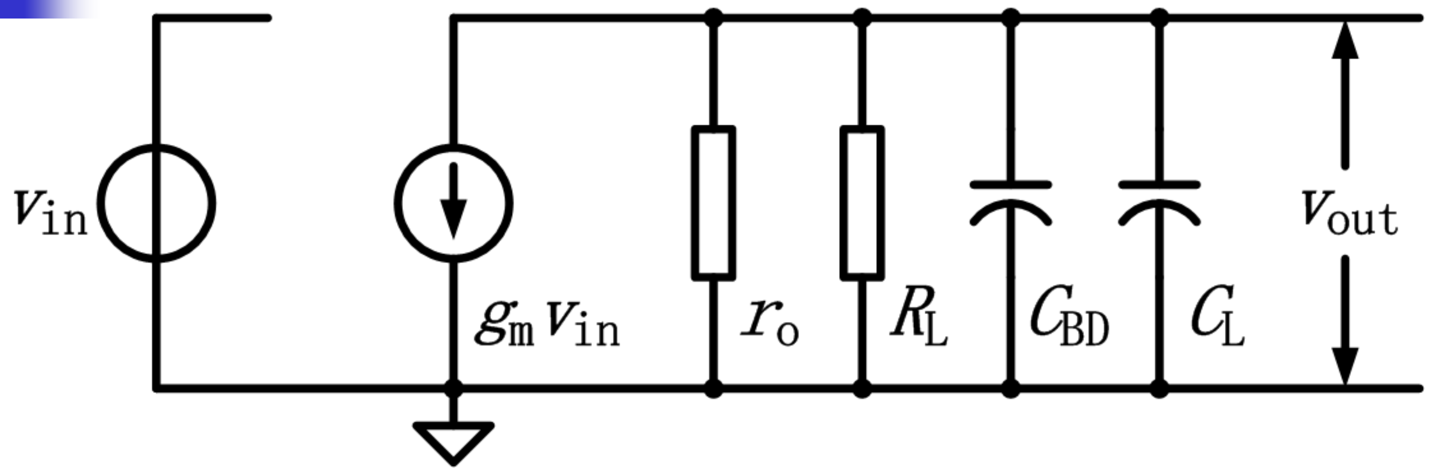
$$BW = \frac{1}{2\pi\left(R_L//r_o\right)\left(C_{BD}+C_L\right)}$$

- 增益带宽积

$$GBW = A_0 BW = \frac{g_m}{2\pi\left(C_{BD}+C_L\right)}$$

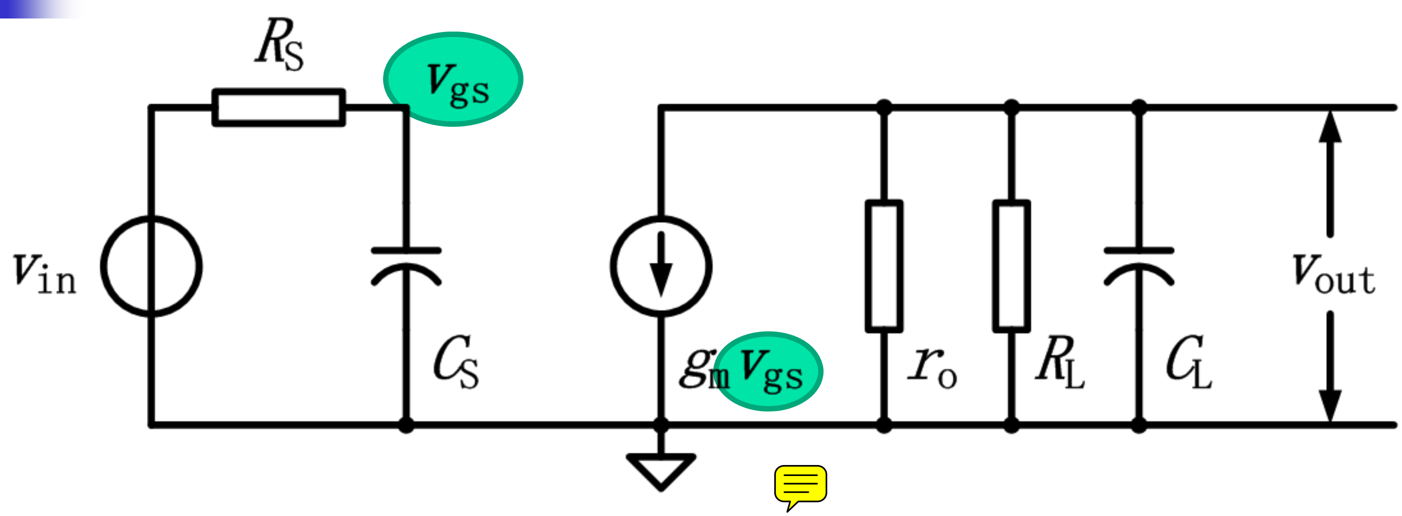


高频交流指标3-零极点分布1（1极点）-高频小信号等效电路图



- 只有一个左半平面的极点
- 极点频率就等于3dB带宽

高频交流指标3-零极点分布2（2极点）-高频小信号等效电路图

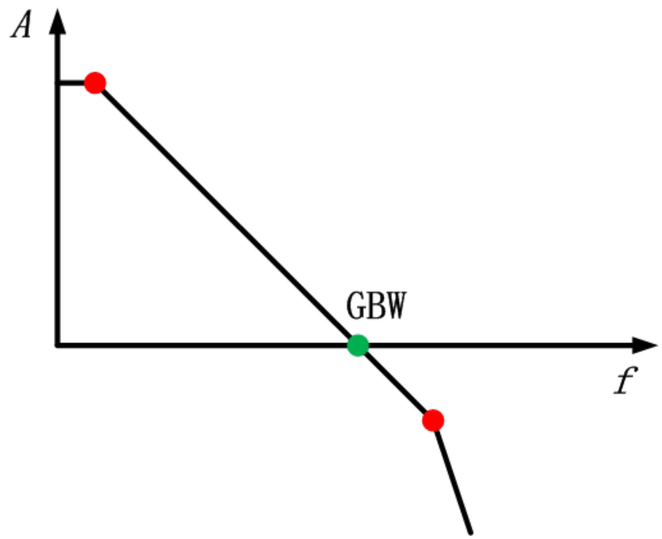


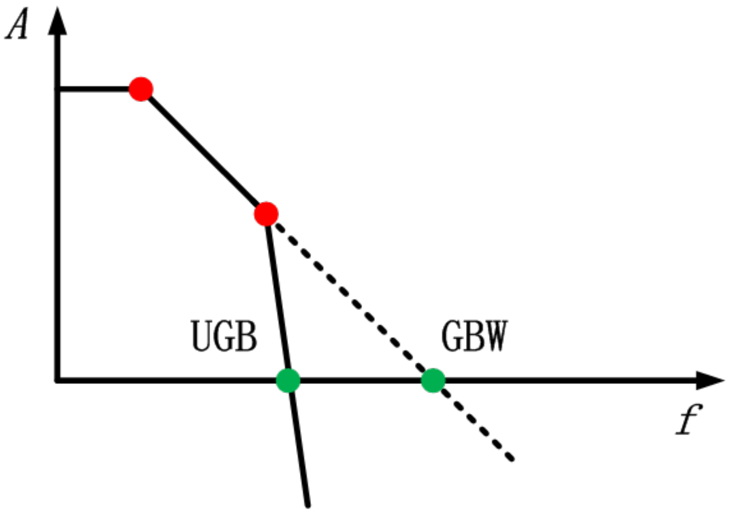
- 信号源内阻为R_S

$$|A(s)|=\frac{\frac{1}{sC_S}}{R_S+\frac{1}{sC_S}}\cdot\frac{g_m(R_L//r_o)}{1+\frac{s}{(R_L//r_o)C_L}}=\frac{1}{1+\frac{1}{\boxed{\frac{s}{(1/R_SC_S)}}}}\cdot\frac{g_m(R_L//r_o)}{1+\boxed{\frac{s}{[1/(R_L//r_o)C_L]}}}$$

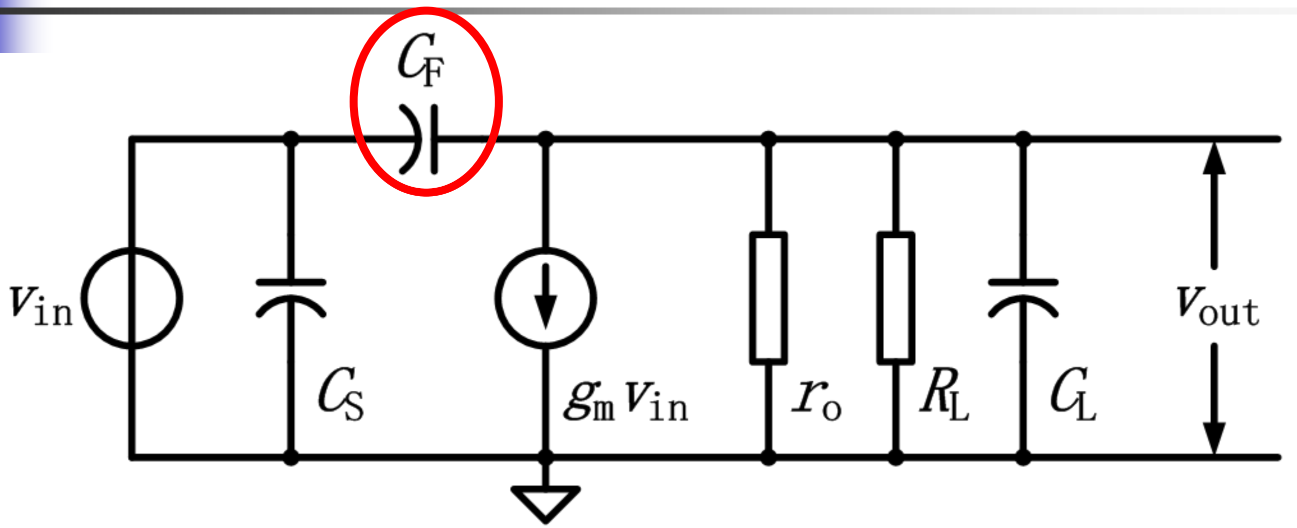
更多说明

- 主极点和非主极点
  - 1个极点在GBW之内
  - 1个极点在GBW之外
  - BW基本由主极点决定
  - 主极点在输入还是输出？
- 双主极点
  - 2个极点对应增益都大于0dB
  - 2个极点都在UGB之内
  - BW和两个极点都有关
  - UGB：Unit Gain Bandwidth





高频交流指标3-零极点分布3（1极点，1零点）-高频小信号等效电路图



$$\frac{v_{out}}{v_{in}}=\frac{-g_m+sC_F}{\frac{1}{r_{out}}+s\left(C_F+C_L\right)}=\frac{-g_m r_{out}}{1+s\left(C_F+C_L\right)r_{out}}+\frac{sC_F r_{out}}{1+s\left(C_F+C_L\right)r_{out}}$$

$$\frac{v_{out}}{v_{in}}=-g_m r_{out}\frac{1+s/\omega_{zo}}{1+s/\omega_{po}}\quad \boxed{\omega_{po}=\frac{1}{r_{out}\left(C_L+C_F\right)}}\quad \boxed{\omega_{zo}=-\frac{g_m}{C_F}}\quad r_{out}=R_L//r_o$$

见32页

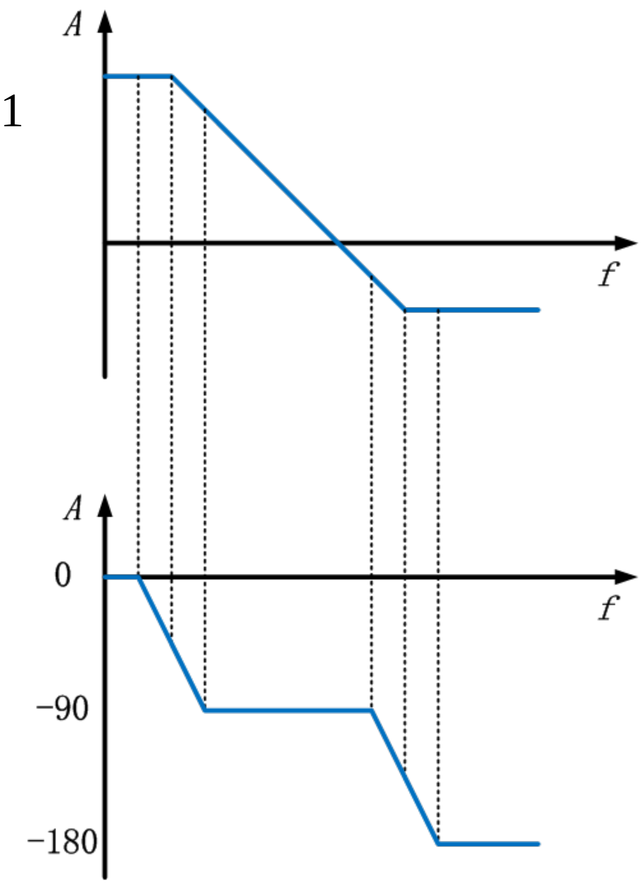
见25页

更多说明

- 零极点分布
  - 左半平面极点
  - 右半平面零点
  - 负载电容越大
    - 极点频率越低
    - 零极点越远

$$\left|\frac{f_{p0}}{GBW}\right| = \frac{\dfrac{1}{2\pi r_{out} C_L'}}{\dfrac{g_m}{2\pi C_L'}} = \frac{1}{g_m r_{out}} < 1$$

$$\left|\frac{f_{z0}}{GBW}\right| = \frac{\dfrac{g_m}{2\pi C_F}}{\dfrac{g_m}{2\pi C_L'}} = \frac{C_L'}{C_F} > 1$$

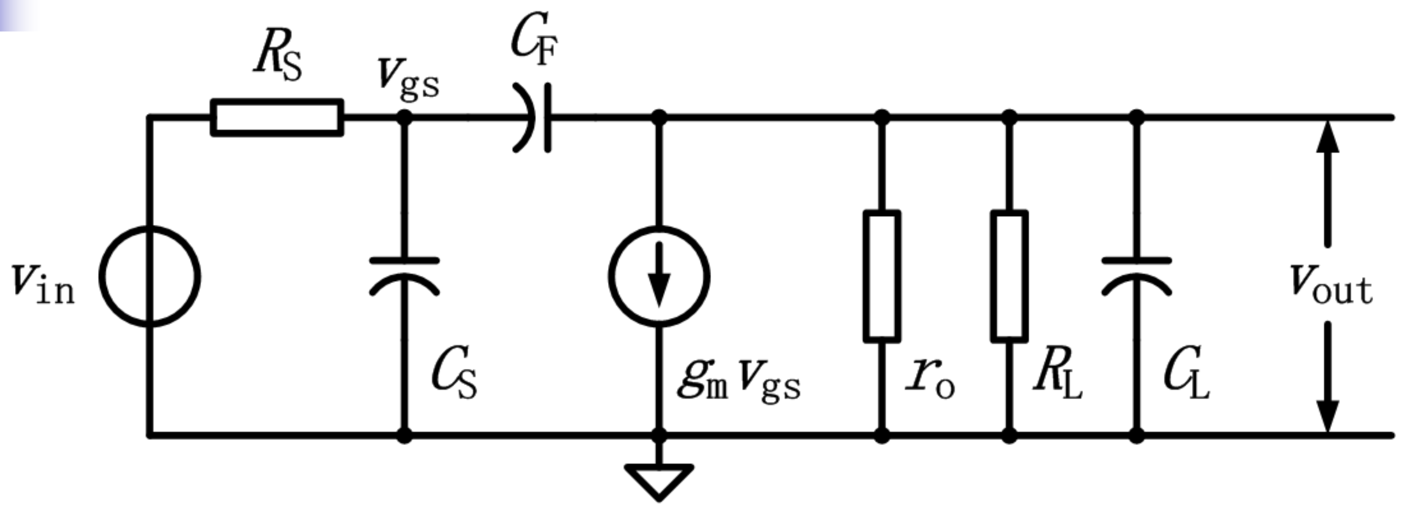


- C_S的作用
  - 零频率极点

$$C_L' = C_L + C_F$$

- 零点产生原因
  - 输入输出之间的多条通路

高频交流指标3-零极点分布4（2极点，1零点）-高频小信号等效电路图



$$\frac{v_{out}}{v_{in}} = \frac{-g_m r_{out} + sC_F r_{out}}{R_S r_{out} K s^2 + \left[\left(1+g_m r_{out}\right)C_F R_S + C_S R_S + \left(C_F + C_L\right) r_{out}\right] s + 1}$$

$$K = C_S C_F + C_S C_L + C_F C_L$$

更多说明

$$\left(\frac{s}{\omega_{p1}}+1\right)\left(\frac{s}{\omega_{p2}}+1\right)=\frac{s^2}{\omega_{p1}\omega_{p2}}+\left(\frac{1}{\omega_{p1}}+\frac{1}{\omega_{p2}}\right)s+1$$

$$\frac{1}{\omega_{p1}}+\frac{1}{\omega_{p2}}\approx\frac{1}{\omega_{p1}}=\left(1+g_m r_{out}\right)C_F R_S+C_S R_S+\left(C_F+C_L\right)r_{out}$$

$$\omega_{p1}=\frac{1}{\left[\left(1+g_m r_{out}\right)C_F+C_S\right]R_S+\left(C_F+C_L\right)r_{out}}$$

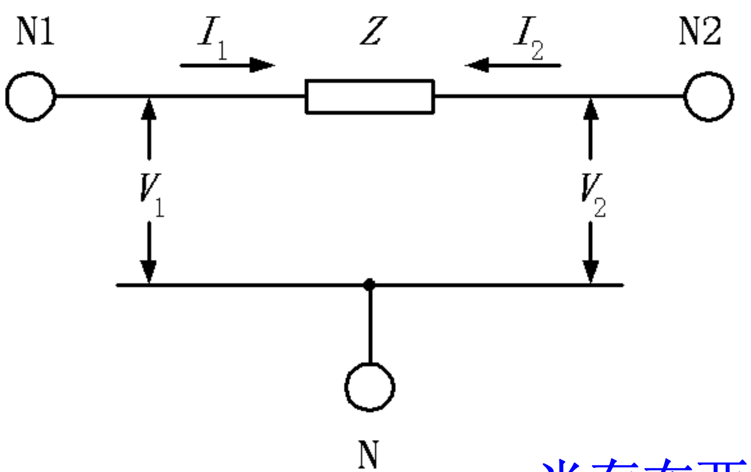
$$\omega_{p2}=\frac{1}{\omega_{p1}R_S r_{out}K}=\frac{\left[\left(1+g_m r_{out}\right)C_F+C_S\right]R_S+\left(C_F+C_L\right)r_{out}}{R_S r_{out}\left(C_S C_F+C_S C_L+C_F C_L\right)}$$

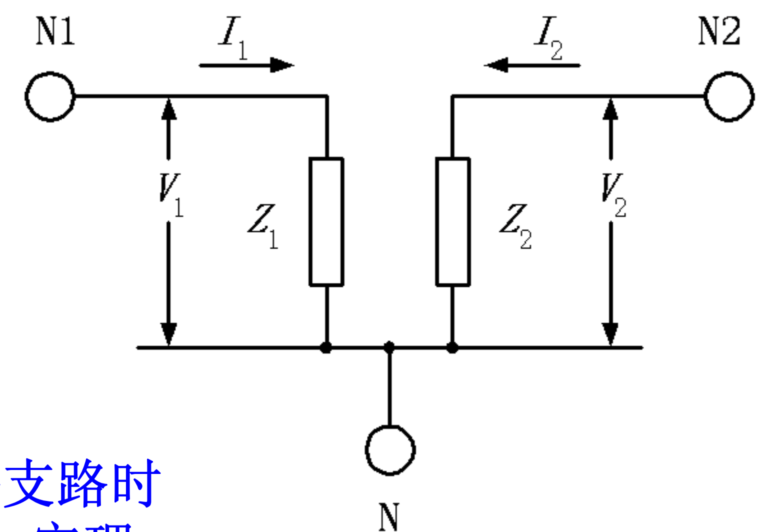
- 注意求根方法的局限性：
  - 方程的两个根一定是两个实数根
  - 两个根一定距离非常远

1、miller电容较小的情况下，输入和输出节点之间相互几乎没有影响，可以使用各自到地电阻、电容的乘积表示极点大小
2、假设此时主极点在输出节点，随着miller电容的增大，非主极点频率快速（线性）下降；主极点频率略有下降。
3、miller电容较大的情况下，主极点从输出节点转移为输入节点。即miller补偿的结果。输出极点则基本保持不变。
4、当原来的两个极点频率接近的时候，miller补偿才能看到极点分离的现象

补充：密勒定理（Miller）

- 主要作用：
  - 解决计算极点时存在的问题





当存在两条支路时
常使用这一定理

密勒定理公式推导

$$A = \frac{V_2}{V_1}$$

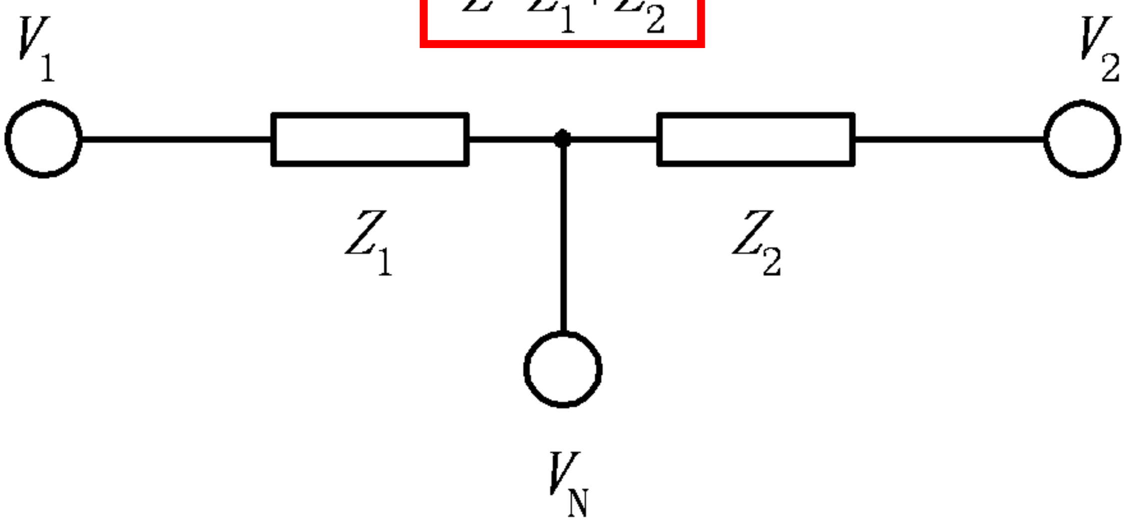
$$I_1 = \frac{V_1 - V_2}{Z} = \frac{V_1 - AV_1}{Z} = \frac{(1-A)V_1}{Z} = \frac{V_1}{Z/(1-A)}$$

$$I_2 = \frac{V_2 - V_1}{Z} = \frac{V_2 - V_2/A}{Z} = \frac{(1-1/A)V_2}{Z} = \frac{V_2}{Z/(1-1/A)}$$

$$\boxed{Z_1 = \frac{1}{1-A}Z \qquad Z_2 = \frac{-A}{1-A}Z}$$

密勒定理中阻抗之间的关系

$\boxed{Z = Z_1 + Z_2}$

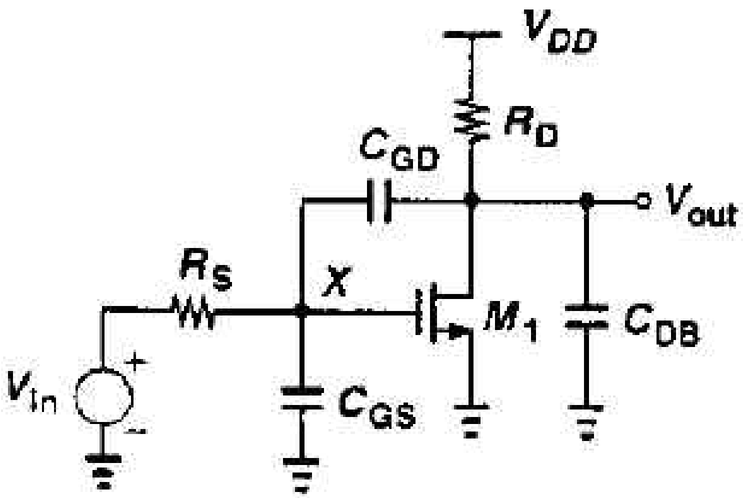


- 以V_N为交流地，从分压角度看阻抗的划分
- Z，Z_1，Z_2（实部）是正值，那么要求增益A<0

密勒定理的应用条件

- 1、两个节点之间存在2条以上（包含2条）的通路
- 2、密勒定理适用于低频情况，高频下会产生很大误差
- 3、常用情况：
  - 低频固定增益的主通路
  - 电容构成的第二条通路
- 4、虽然定理不完善，但我们照用

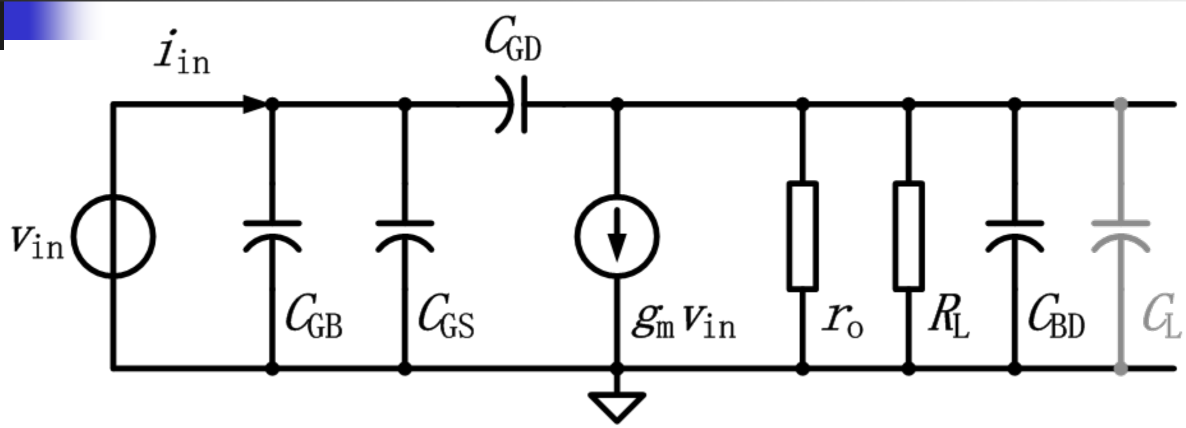
应用密勒定理的极点节点对应关系



$$C_{in} = C_{GS} + \left(1 + g_m R_D\right) C_{GD}$$

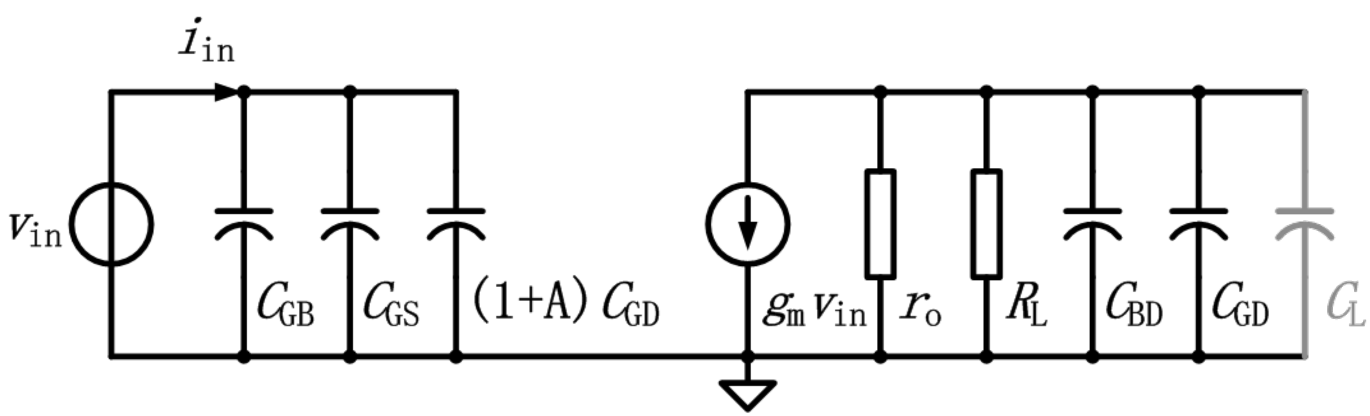
$$C_{out} = C_{BD} + \left(1 + \frac{1}{g_m R_D}\right) C_{GD} \approx C_{BD} + C_{GD}$$

高频交流指标4-输入阻抗-
高频小信号等效电路图



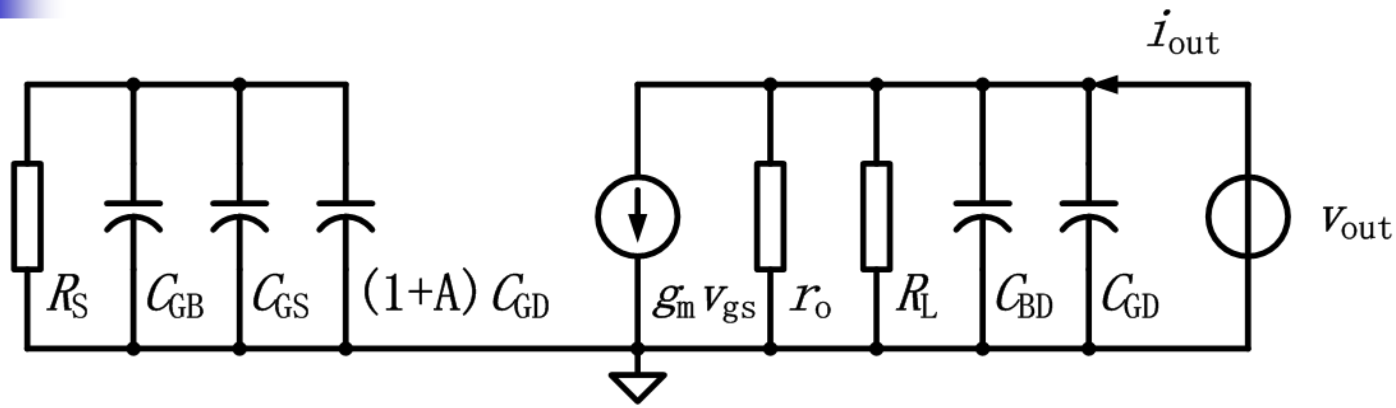
使用miller定理简化计算

否则输入-输出阻抗之间通过C_{GD}相互影响



$$Z_{in}(s) = s\left[C_{GS} + C_{GB} + \left(1 + g_m r_{out}\right) C_{GD}\right]$$

高频交流指标5-输出阻抗-高频小信号等效电路图

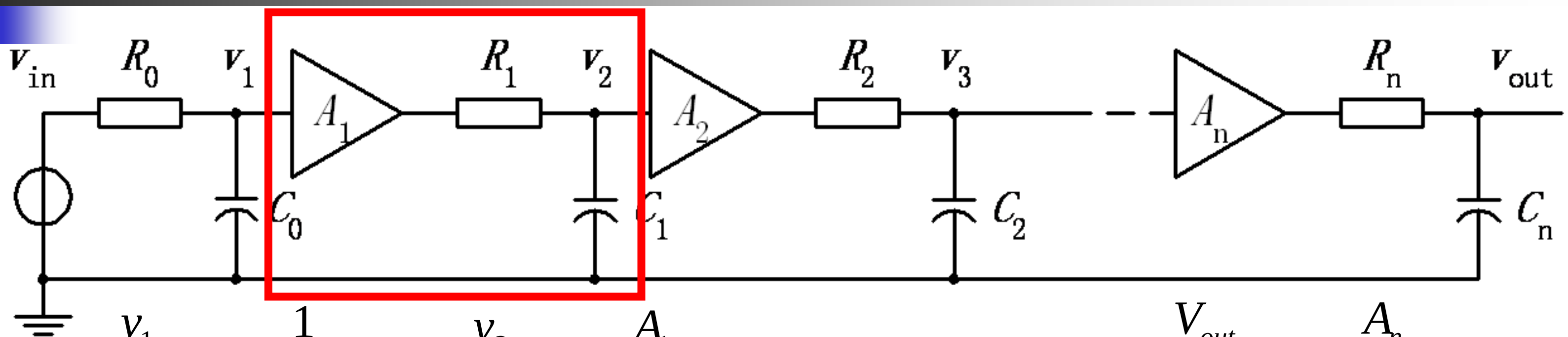


$$Z_{out}(s) = (R_L // r_o) // (C_{BD} + C_{GD})$$

高频交流指标的直观分析

- 1极点
- 对应节点的RC乘积
- 2极点
- 各自对应节点的RC乘积
- 1极点，1零点
- 对应节点的RC乘积，C考虑了miller效应
- 2极点，1零点
- 各自对应节点的RC乘积， C考虑了miller效应

补充：极点和电路节点的对应关系



$$\frac{v_1}{v_{in}} = \frac{1}{1+sR_0C_0} \qquad \frac{v_2}{v_1} = \frac{A_1}{1+sR_1C_1} \qquad \frac{V_{out}}{V_n} = \frac{A_n}{1+sR_nC_n}$$

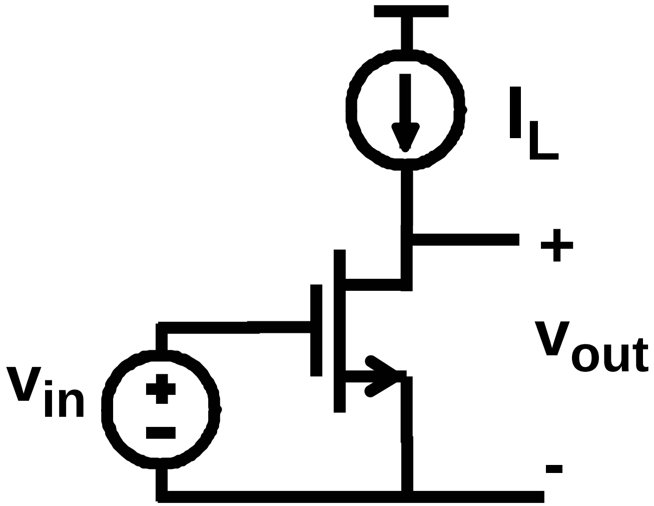
$$\frac{v_{out}}{v_{in}} = \frac{1}{1+sR_0C_0}\frac{A_1}{1+sR_1C_1}\cdots\frac{A_n}{1+sR_nC_n} = \frac{1}{1+\dfrac{s}{1/R_0C_0}}\frac{A_1}{1+\dfrac{s}{1/R_1C_1}}\cdots\frac{A_n}{1+\dfrac{s}{1/R_nC_n}}$$

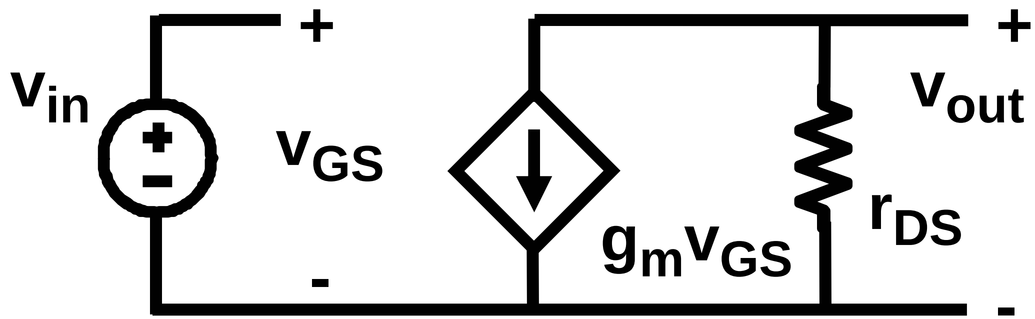
$$\frac{v_{out}}{v_{in}} = \frac{1}{(1+\dfrac{s}{\omega_{p0}})}\frac{A_1}{(1+\dfrac{s}{\omega_{p1}})}\frac{A_2}{(1+\dfrac{s}{\omega_{p2}})}\cdots\frac{A_n}{(1+\dfrac{s}{\omega_{pn}})} \qquad \omega_{pi} = \frac{1}{R_iC_i}(i=0,1,2\ldots,n)$$

开路时间常数法，短路时间常数法：P.R.Gray，4th，7.3.1，7.3.2，7.3.6

1

Single-transistor amplifier - 1





$$A_v = g_m r_{DS} = \frac{2\, I_{DS}}{V_{GS}-V_T}\,\frac{V_E\, L}{I_{DS}} = \frac{2\, V_E\, L}{V_{GS}-V_T}$$

$A_v \approx 100$ if $V_E L \approx 10$ V and $V_{GS}-V_T \approx 0.2$ V

MOST or bipolar amplifier ?

MOST $$A_v = \frac{V_E L}{(V_{GS}-V_T)/2}$$

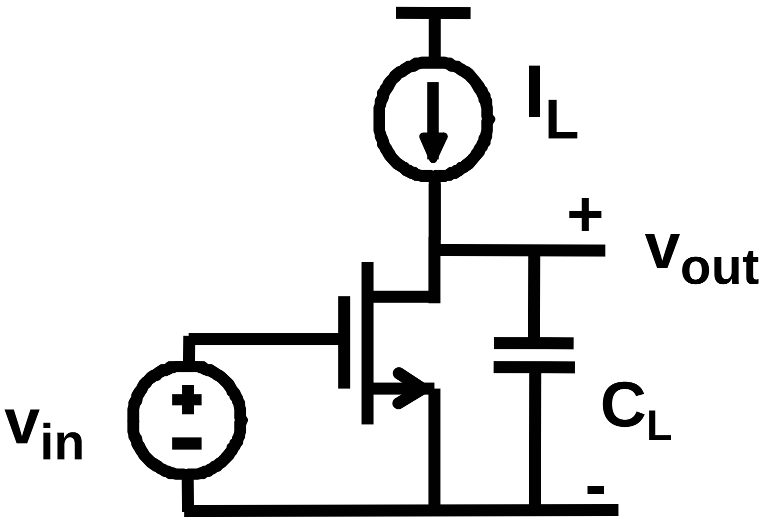
$A_v \approx 100$ if $V_E L \approx 10$ V and $V_{GS}-V_T \approx 0.2$ V

Bipolar $$A_v = \frac{V_E}{kT/q}$$

3 vs 2 stages for 10^6

$A_v \approx 1000$ if $V_E \approx 26$ V since $kT/q = 26$ mV

Gain, Bandwidth and Gain-bandwidth



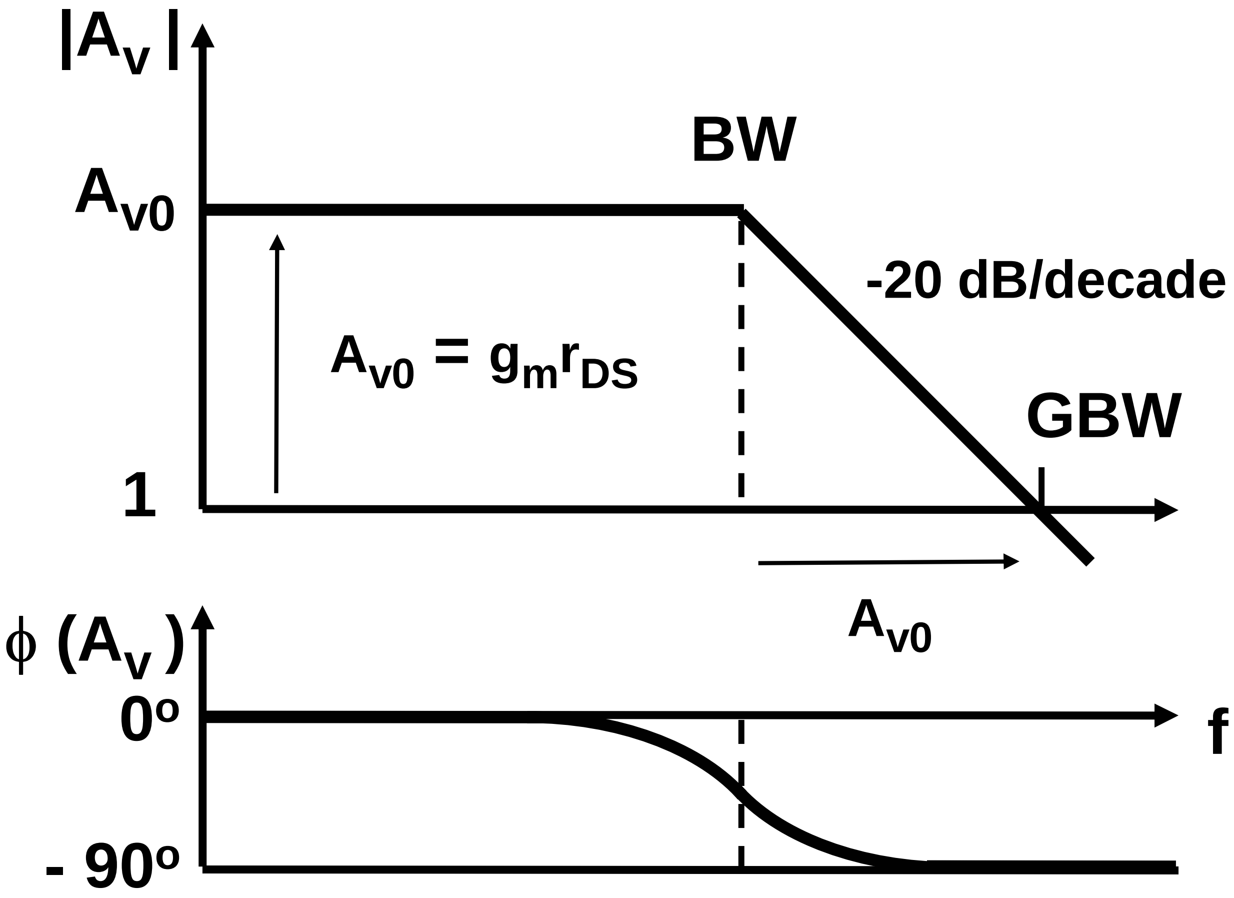
For all single-stage Operational amplifiers

$$A_{v0} = g_m r_{DS}$$

$$BW = \frac{1}{2\pi\, r_{DS} C_L}$$

$$\boxed{GBW = \frac{g_m}{2\pi\, C_L}}$$

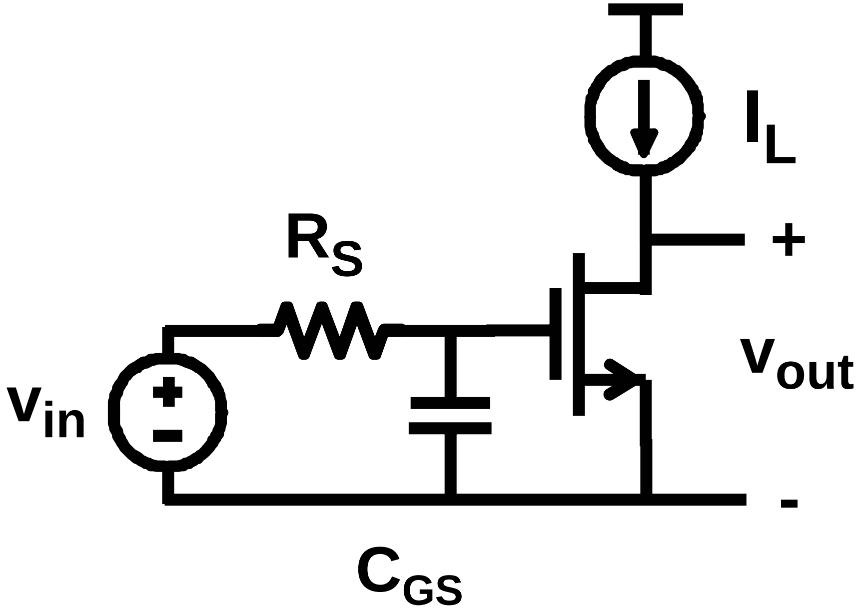
Gain A_v , BW and GBW



$$GBW = \frac{g_m}{2\pi\, C_L}$$

$\phi\,(A_v) = -\,45^o$
at BW

Gain, Bandwidth and Gain-bandwidth



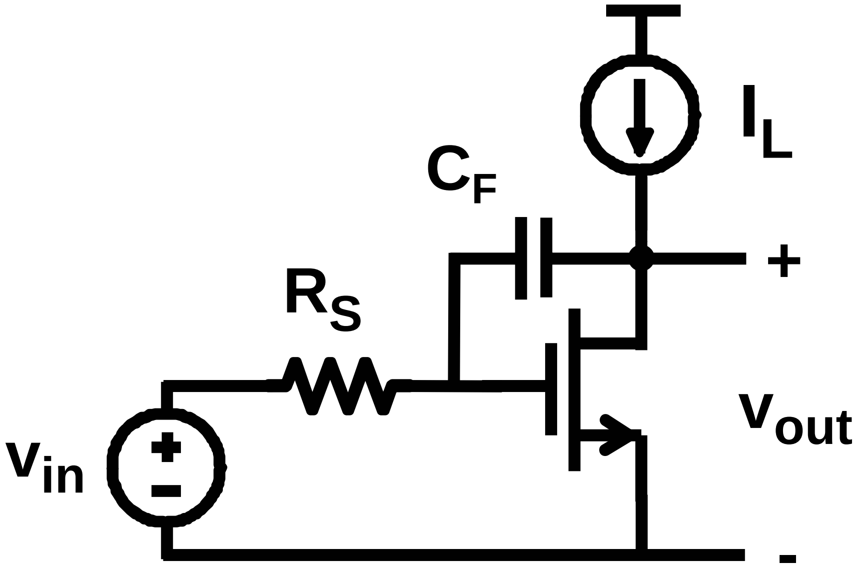
$$A_{v0} = g_m r_{DS}$$

$$\boxed{BW = \frac{1}{2\pi\, R_S C_{GS}}}$$

$$\boxed{GBW = \frac{g_m}{2\pi\, C_{GS}} \frac{r_{DS}}{R_S} = f_T \frac{r_{DS}}{R_S} \sim \frac{1}{WC_{ox}} \frac{1}{V_{GS}-V_T}}$$

W ? L ? V_{GS}-V_T?

Gain, Bandwidth and Gain-bandwidth

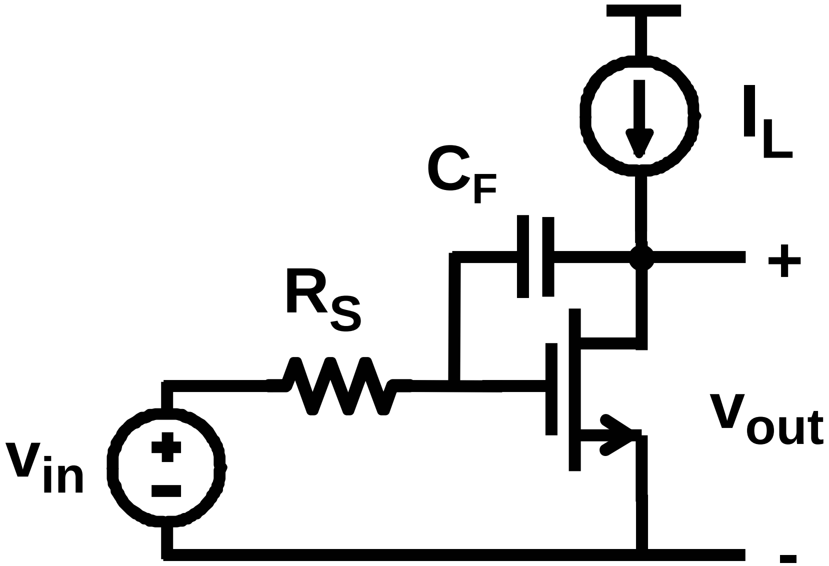


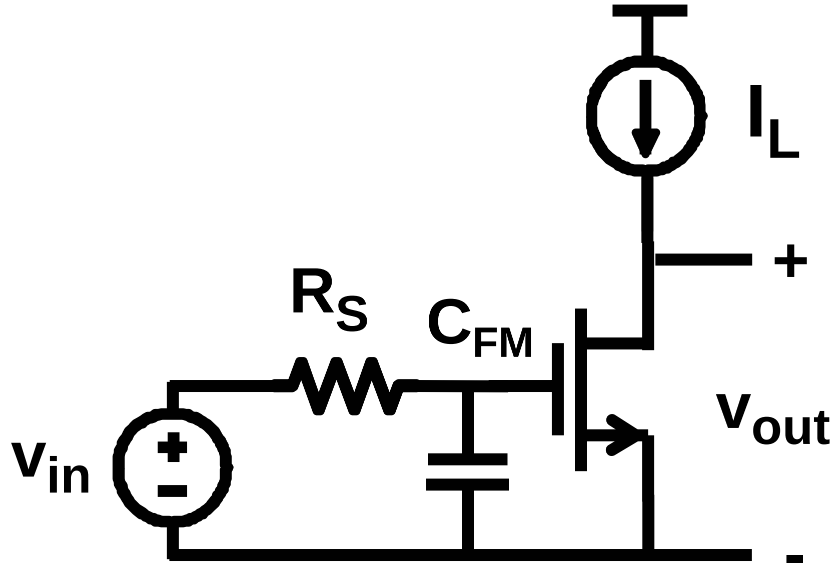
$$A_{v0} = g_m r_{DS}$$

$$BW = \frac{1}{2\pi\, R_S\, A_{v0}\, C_F}$$

$$GBW = \frac{1}{2\pi\, R_S\, C_F}$$

Miller effect



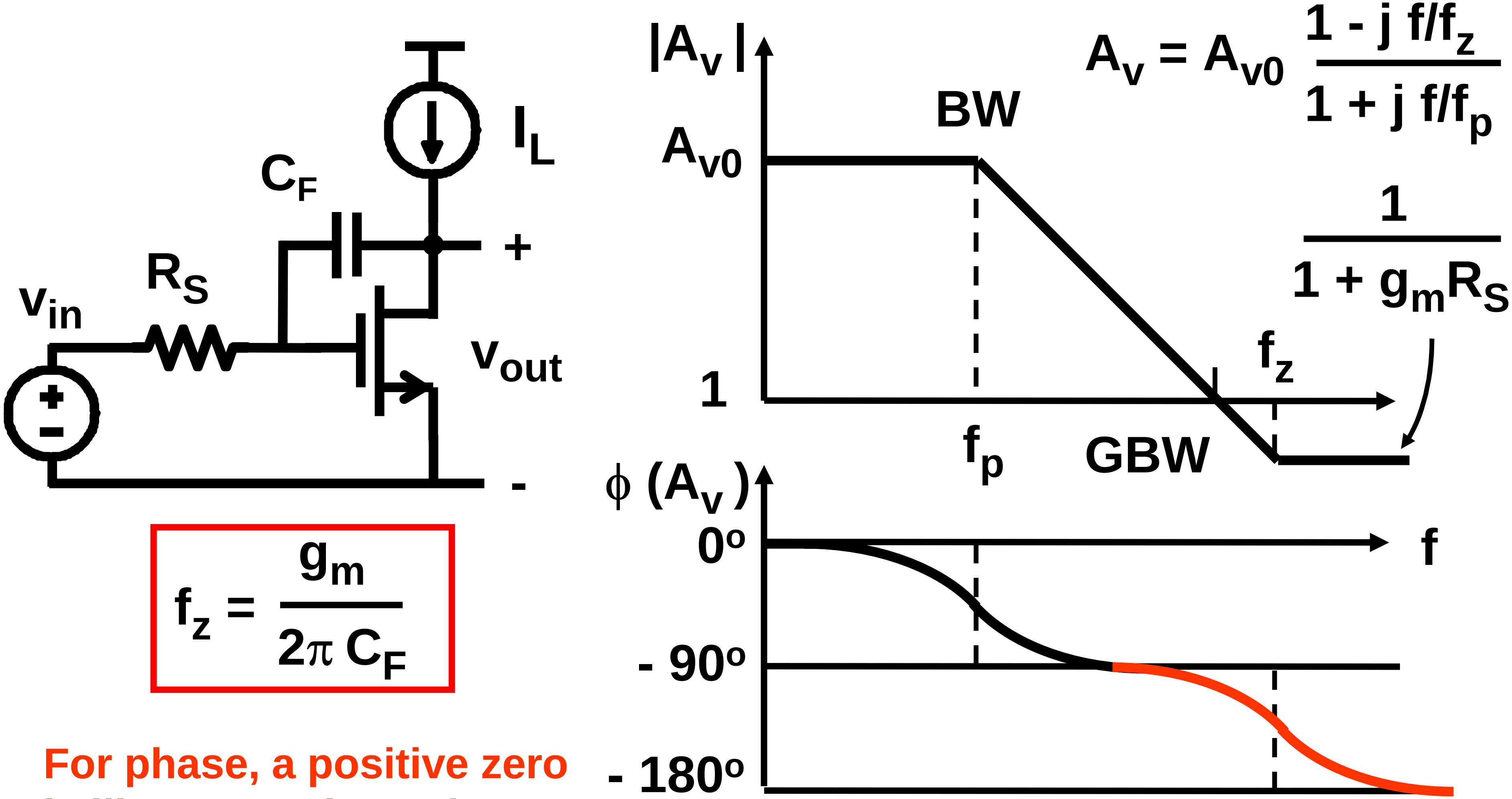


$$A_{v0} = g_m r_{DS}$$

$$\boxed{C_{FM} = (1+A_{v0})\, C_F}$$

Miller, Dependence of the input impedance of a three-electrode vacuum tube upon the load in the plate circuit, Scient. Papers Bur. Standards, 1920, 367-385.

Miller capacitance feedback effects

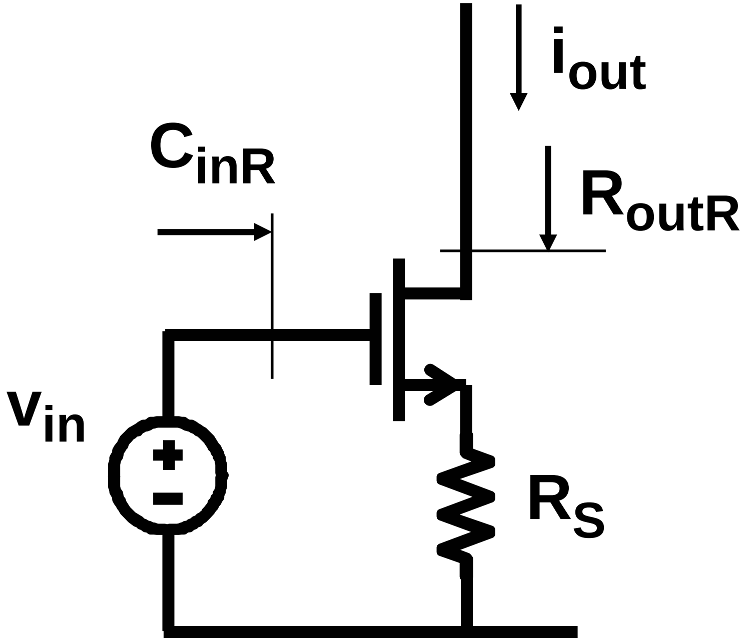


$$f_z = \frac{g_m}{2\pi\, C_F}$$

For phase, a positive zero is like a negative pole !!!

2

Amplifier with local R- (series) feedback



$$g_{mR} = \frac{g_m}{1 + g_m R_S} \sim \frac{1}{R_S}$$

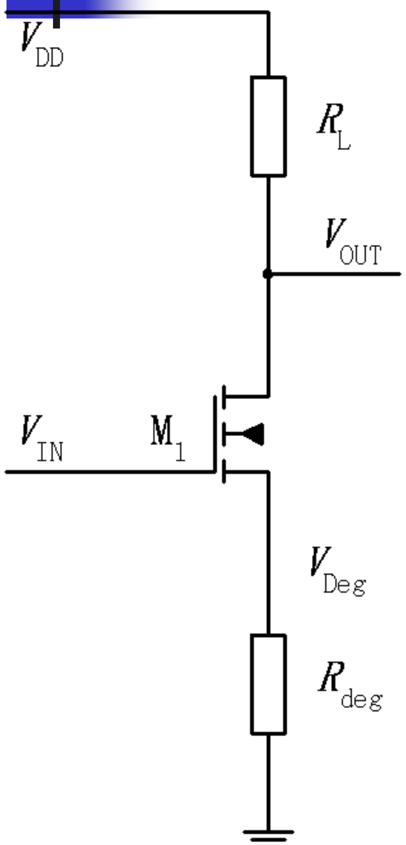
$$R_{outR} = r_{DS}\,(1 + g_m R_S)$$
$$\approx (g_m r_{DS})\,R_S$$

$$C_{inR} = \frac{C_{GS}}{1 + g_m R_S}$$

But R_S gives extra noise !

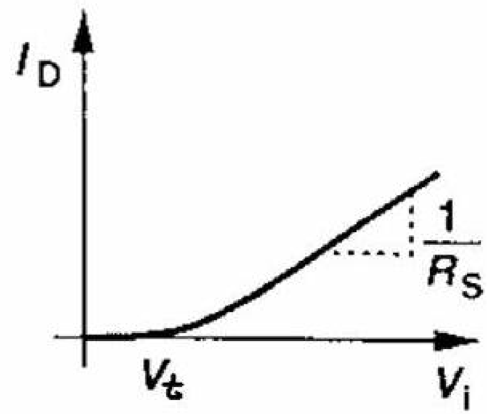
见第4章 噪声

直流工作点-列方程



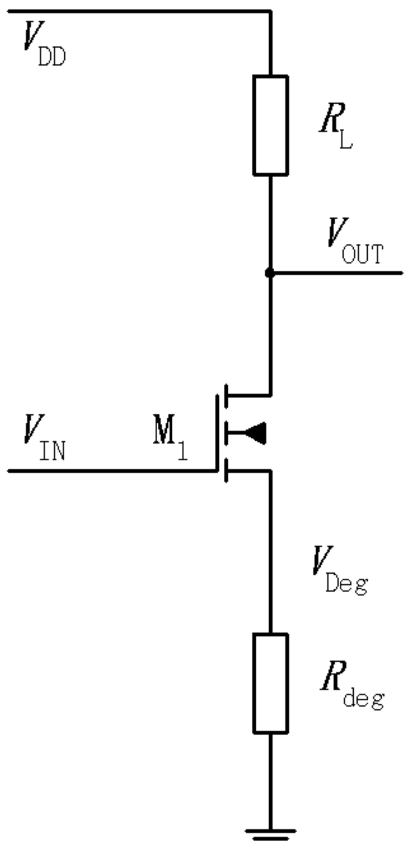
$$I = \frac{\mu_n C_{ox}}{2}\frac{W}{L}\left(V_{IN} - R_{\deg} I\right)^2$$

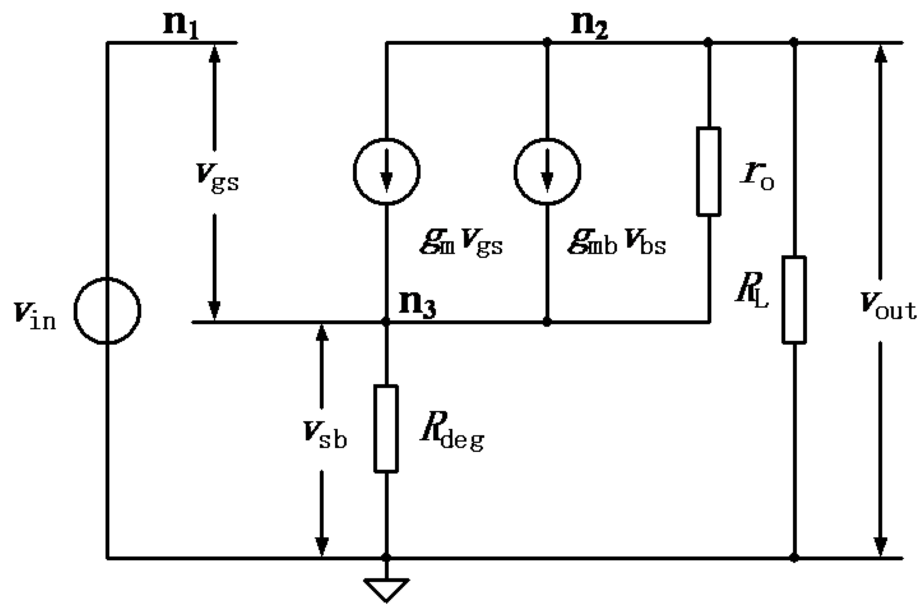
$$\left(\frac{2L}{\mu_n C_{ox} W}\right)^{\frac{1}{2}} I^{\frac{1}{2}} + R_{\deg} I = V_{IN}$$



输入电压主要加在源极负反馈电阻R_{deg}上

直流增益-低频小信号等效电路图





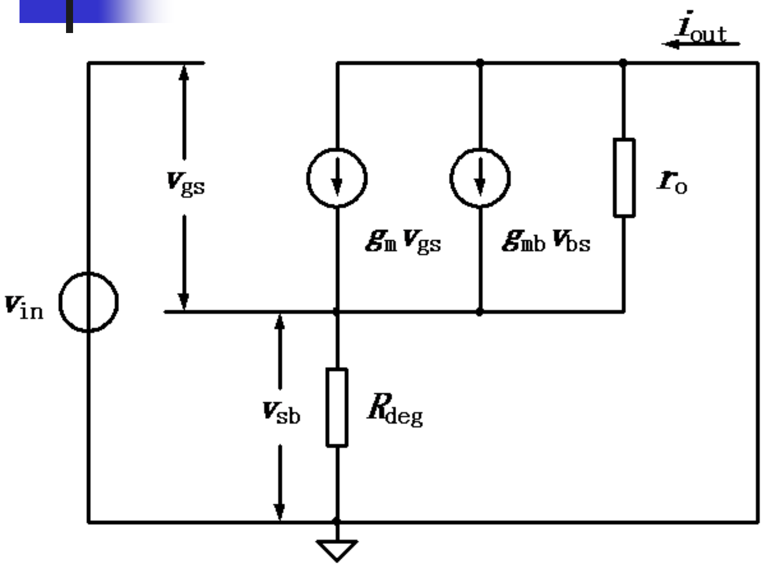
$$\frac{v_{out}}{v_{in}} = \frac{-g_m R_L}{R_{\deg}\left(g_m + g_{mb}\right) + \dfrac{r_o + R_{\deg} + R_L}{r_o}}$$

直流增益-直观分析

- 前面的放大器：
  - 跨导×输出电阻

- 使用源级负反馈电阻的放大器：
  - （等效）跨导×（等效）输出电阻？

等效跨导G_m

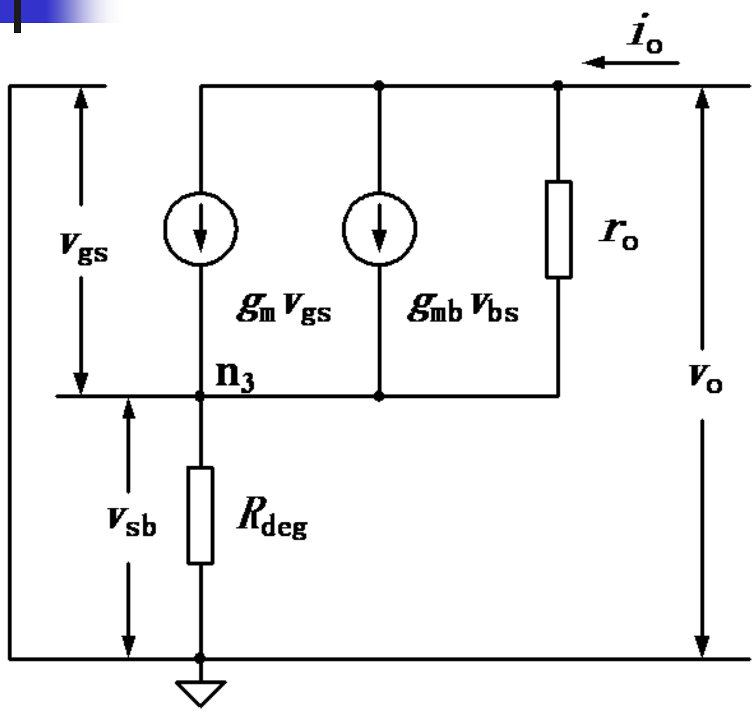


$$G_m = \frac{i_{out}}{v_{in}}$$

$$\boxed{G_m = \frac{g_m}{R_{\deg}\left(g_m + g_{mb} + 1/r_o\right) + 1} \qquad G_m \approx \frac{g_m}{1 + g_m R_{\deg}} \approx \frac{1}{R_{\deg}}}$$

- 等效跨导的计算—短路跨导：
  - 令输出短路到地
  - 在一定的信号电压v_{in}激励下，计算流过输出端的电流i_{out}
- 使用源极负反馈电阻的共源放大器的等效跨导是输入管的跨导g_m除以（$1+g_mR_{deg}$）
- 跨导线性化：g_m变成$1/R_{deg}$

等效输出电阻r_{out}

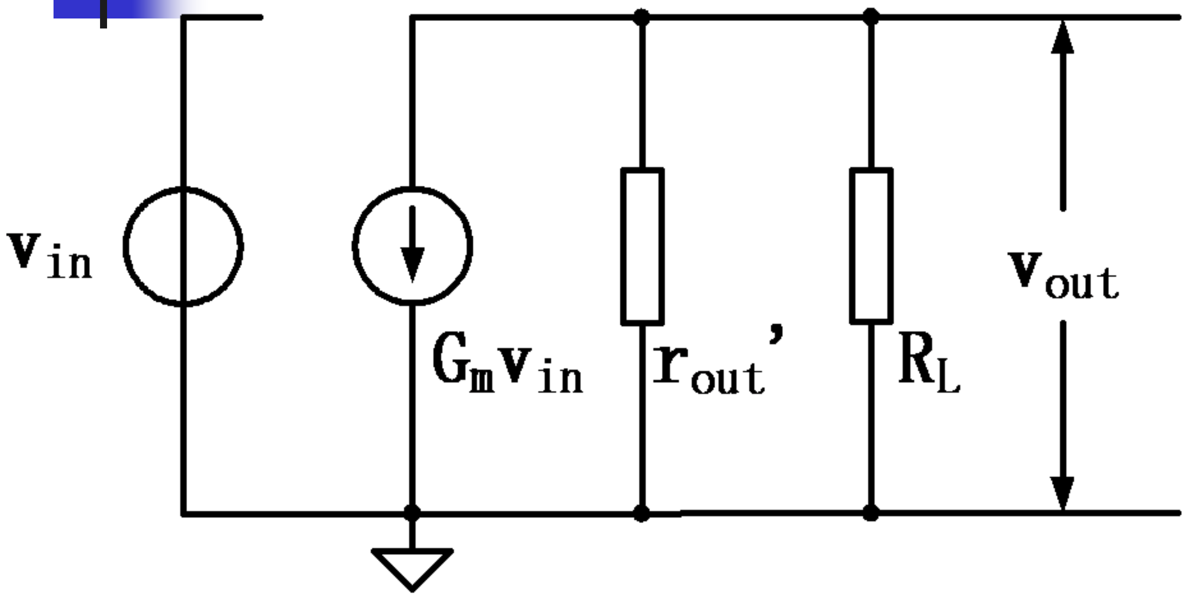


$$r_{out}' = (g_m + g_{mb}) r_o R_{deg} + R_{deg} + r_o$$

$$r_{out}' \approx (g_m r_o) R_{deg}$$

- **电阻R_{deg}大约提高了$g_m r_o$倍，这正是MOS管本征增益**
- 在后面共源共栅电路中还要提到
- $R_{out} = R_L // r_{out}'$
- 如图计算r_{out}'

直流增益



增益=等效跨导×等效输出电阻

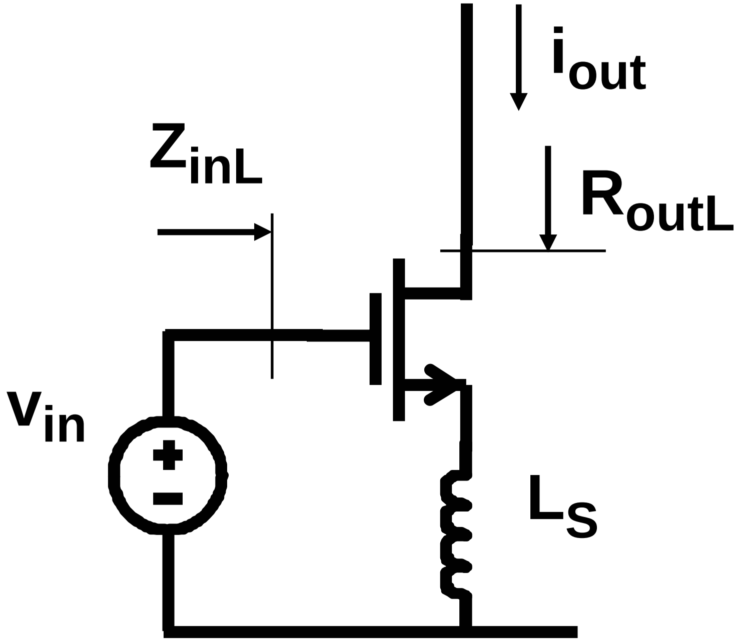
$$\frac{v_{out}}{v_{in}} = G_m\left(r_{out}'//R_L\right) = \frac{-g_m R_L}{R_{\deg}\left(g_m + g_{mb}\right) + \dfrac{r_o + R_{\deg} + R_L}{r_o}} \approx \frac{R_L}{R_{\deg}}$$

增益近似等于两个电阻的比值

源极负反馈结构的总结

- 关于跨导：
  - 等效跨导的推导：短路跨导
  - 等效跨导的公式
  - 跨导线性化
- 关于输出电阻：
  - 输出电阻的增大：提高MOS管本征增益倍
  - 输出电阻的直观计算（向上看、向下看）
- 关于电路增益：
  - 等效跨导×等效输出电阻（直接得结果）
  - 近似等于电阻之比

Amplifier with local L- feedback



$$g_{mL} = \frac{g_m}{1 + g_m L_S s}$$

$$R_{outL} = r_{DS}\,(1 + g_m L_S s)$$

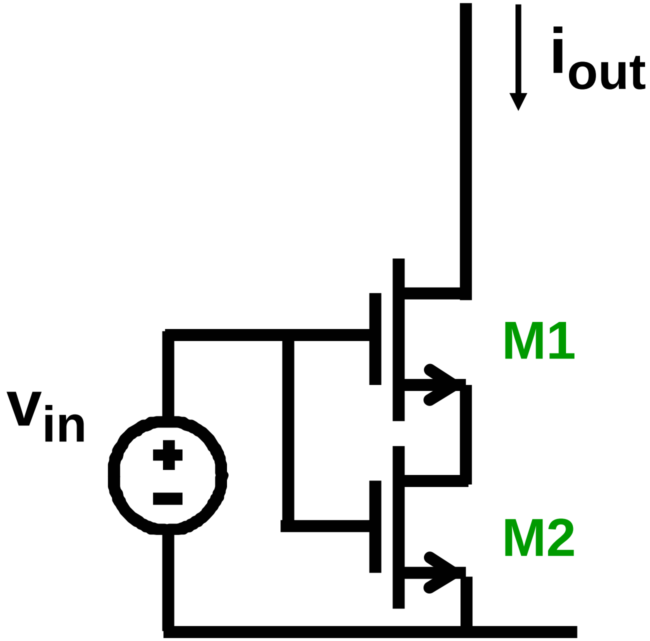
$$Z_{inL} = g_m \frac{L_S}{C_{GS}} + \frac{1 + L_S C_{GS}\, s^2}{s\, C_{GS}}$$

No extra noise !

$$Z_{inL} = L_S\,\omega_T + L_S s + \frac{1}{s\, C_{GS}}$$

补充文章的说明

Amplifier with local MOST-R- Feedback



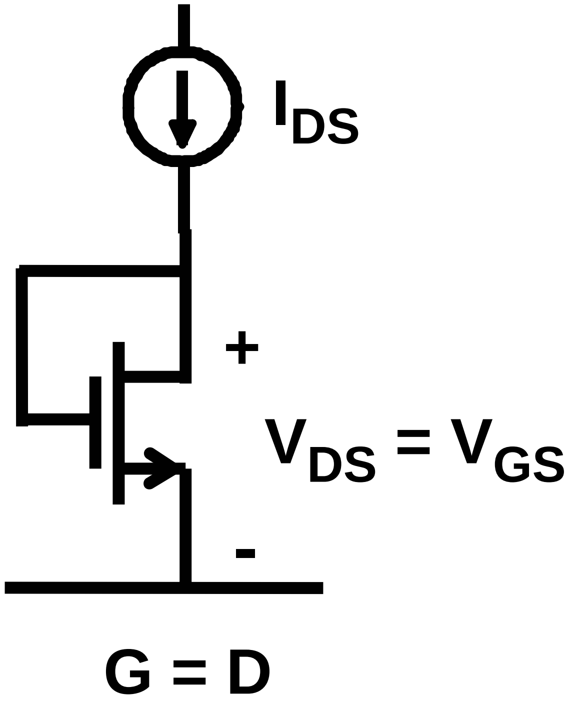
$$V_{DS2} = V_{GS2} - V_{GS1} \approx 0.2\ V$$

$$r_{DS2} = \frac{1}{KP\ W_2/L_2\ (V_{GS2}-V_T)}$$

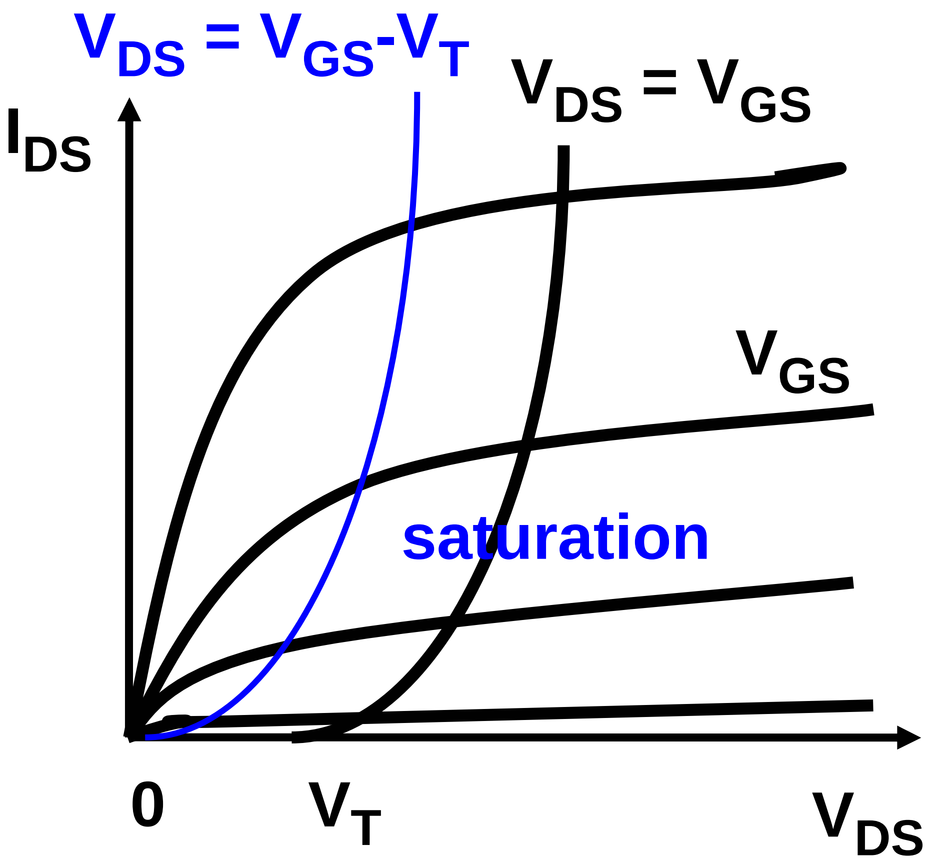
$$R_{outR} = r_{DS1}\ (1+ g_{m1}\ r_{DS2})$$

$$C_{inR} = \frac{C_{GS1}+C_{GS2}}{1 + g_{m1}r_{DS2}}$$

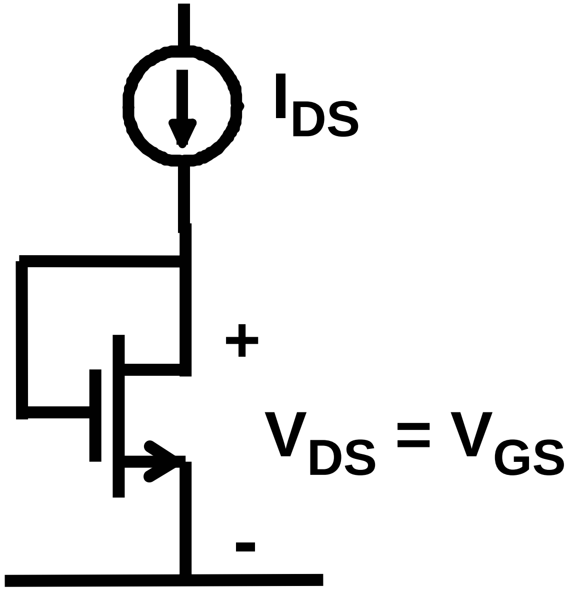
Diode-connected MOST : parallel Feedback



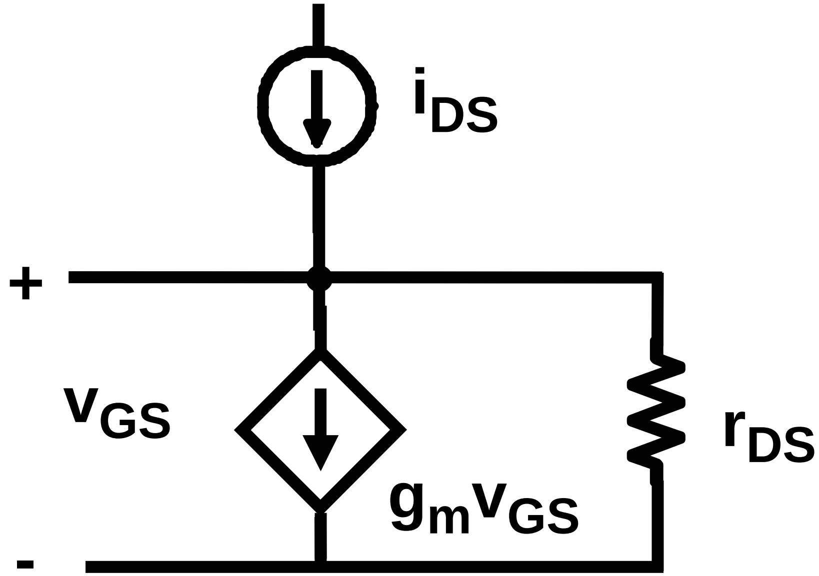
$$I_{DS} = K'_n \frac{W}{L} (V_{DS}-V_T)^2$$



Diode-connected MOST: small-signal

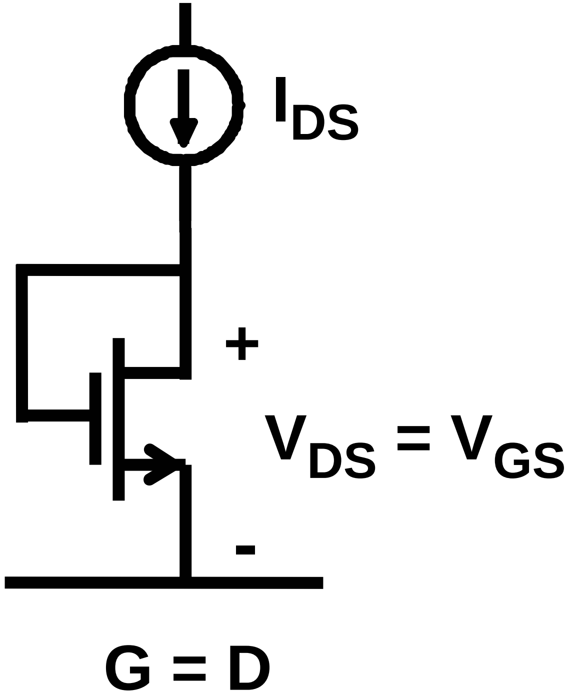


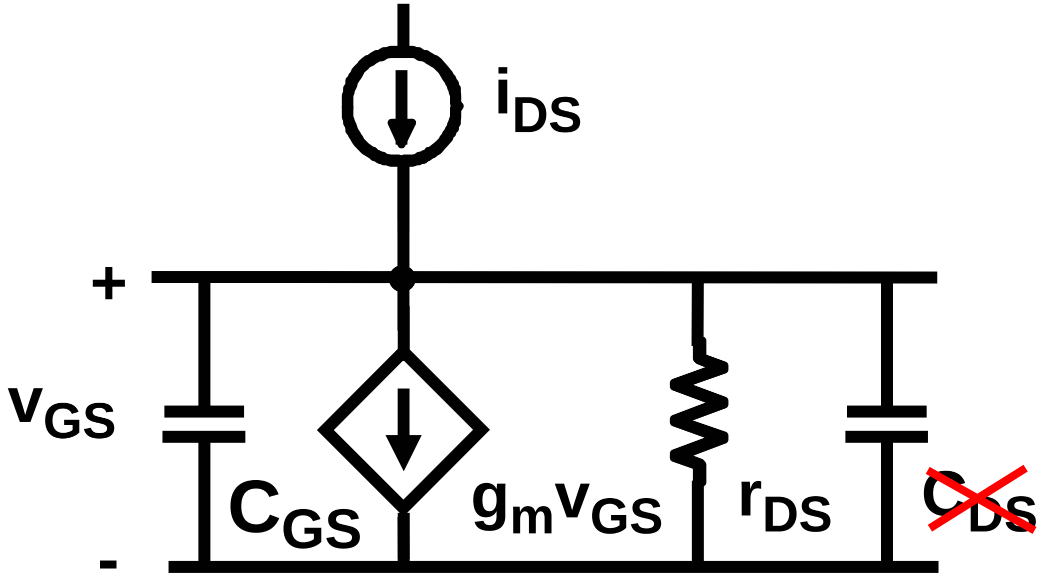
G = D



$$\boxed{r_{ds} = 1/g_m \; // \; r_{DS} \approx 1/g_m}$$

Diode-connected MOST at high frequencies

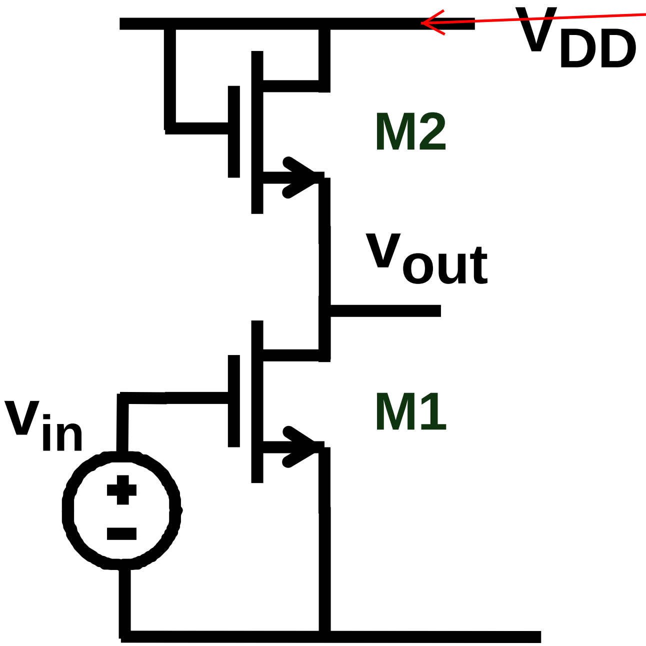




$$\boxed{BW = \frac{g_m}{2\pi\,(C_{GS}+C_{DS})} \approx \frac{f_T}{2}}$$

Wideband amplifier

注意结构上的特点

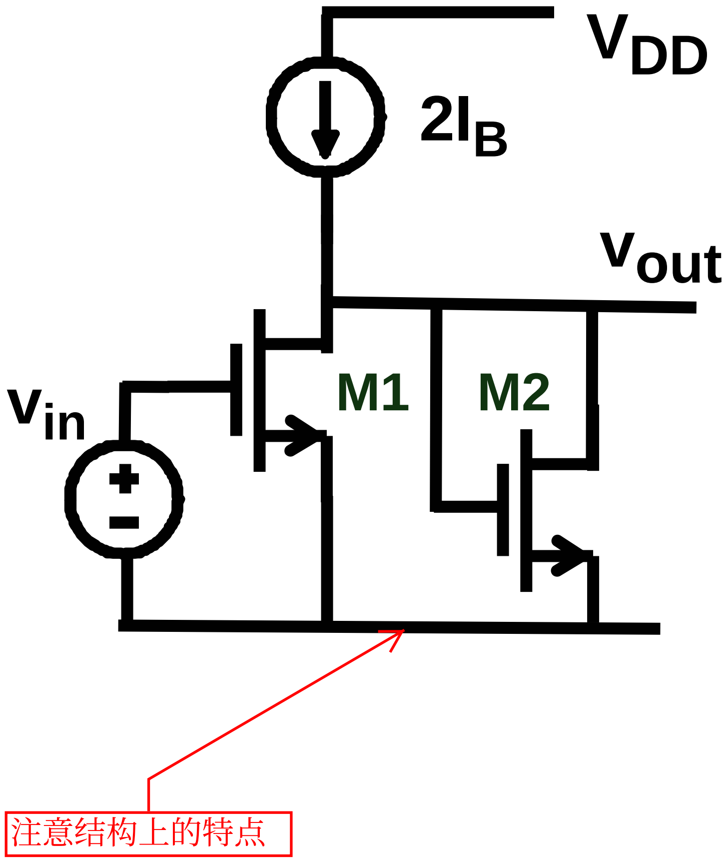


$$V_{OUT} = V_{DD} - V_{GS2}(V_{OUT})$$

$$A_{v0} = \frac{g_{m1}}{g_{m2}} = \sqrt{\frac{(W/L)_1}{(W/L)_2}} = \frac{V_{GS2} - V_T}{V_{GS1} - V_T}$$

$$R_{OUT} = 1/g_{m2}$$

Linear wideband amplifier



$$V_{OUT} = V_{GS2}$$

$$A_{v0} = \frac{g_{m1}}{g_{m2}} = \sqrt{\frac{(W/L)_1}{(W/L)_2}} = \frac{V_{GS2} - V_T}{V_{GS1} - V_T}$$

$$R_{OUT} = 1/g_{m2}$$

Current mirror with only nMOSTs

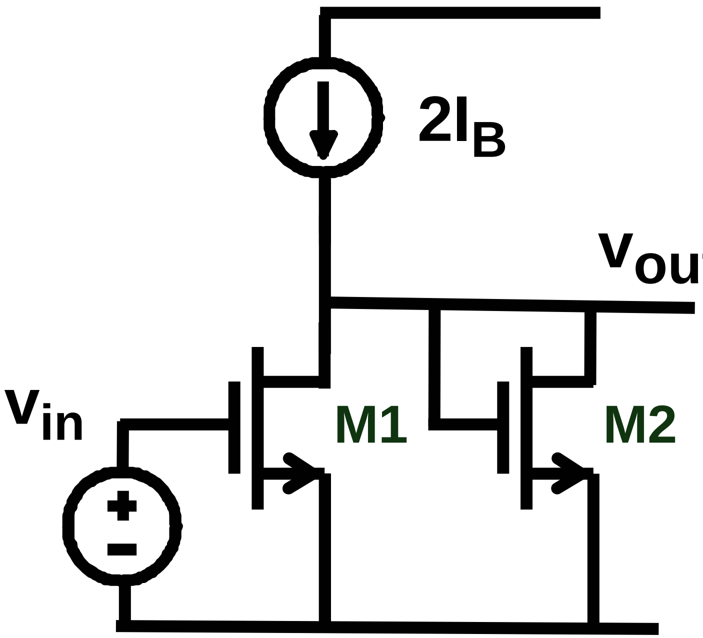
Same V_{OUTDC} as V_{INDC}

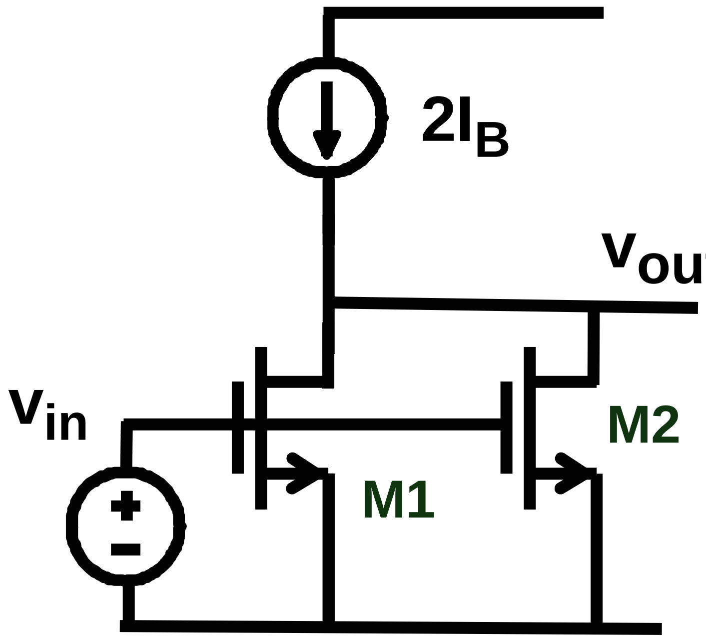
No body bias effect

Good PSRR

Double power consumption

Wideband amplifiers





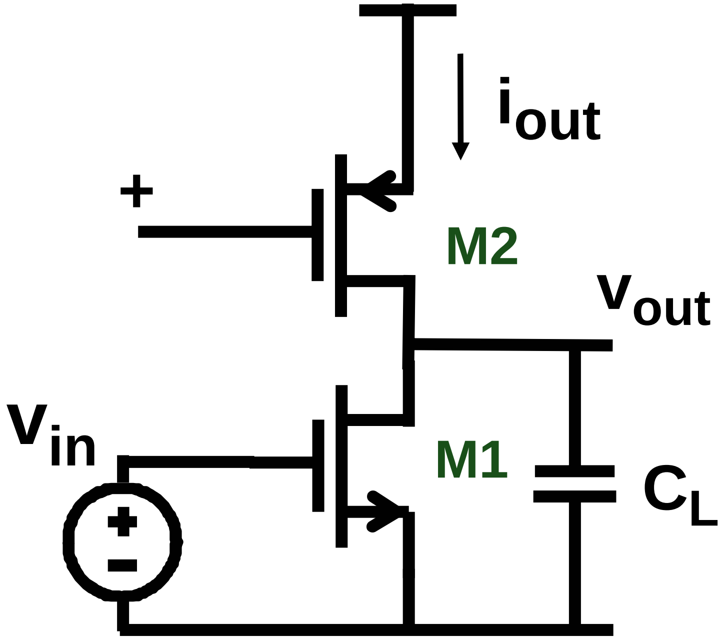
$$A_{v0} = \frac{g_{m1}}{g_{m2}} = \sqrt{\frac{(W/L)_1}{(W/L)_2}} = \frac{V_{GS2} - V_T}{V_{GS1} - V_T}$$

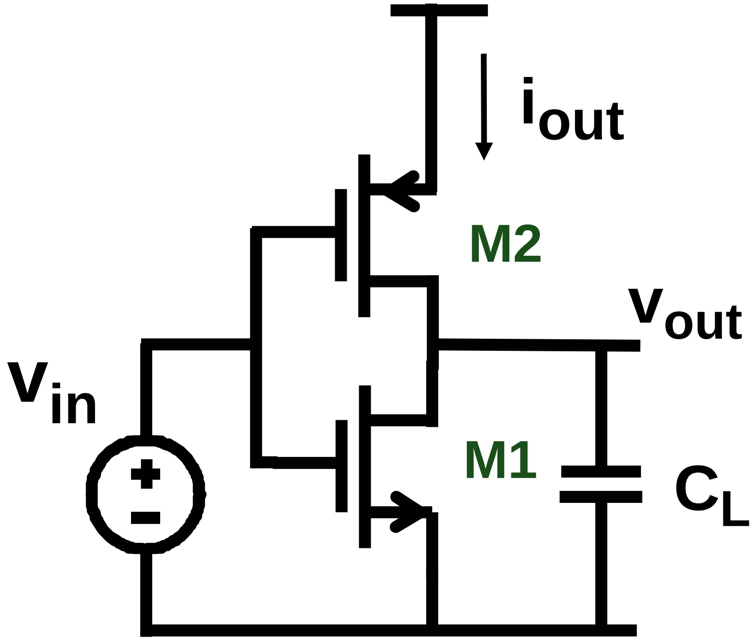
$$R_{out} = 1/g_{m2}$$

$$A_{v0} = g_m R_{out}$$

$$R_{out} = r_{DS1}//r_{DS2}$$

Class A versus class AB amplifier

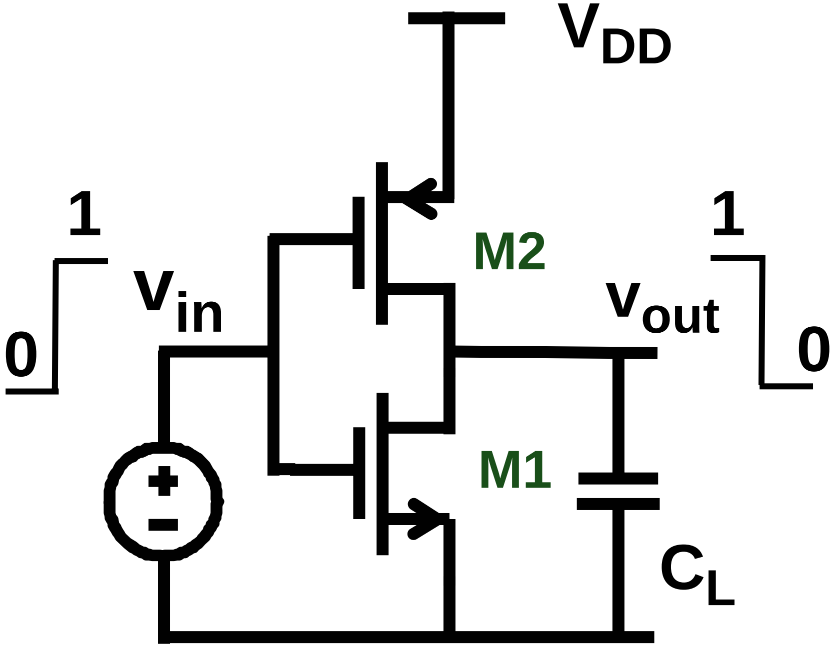


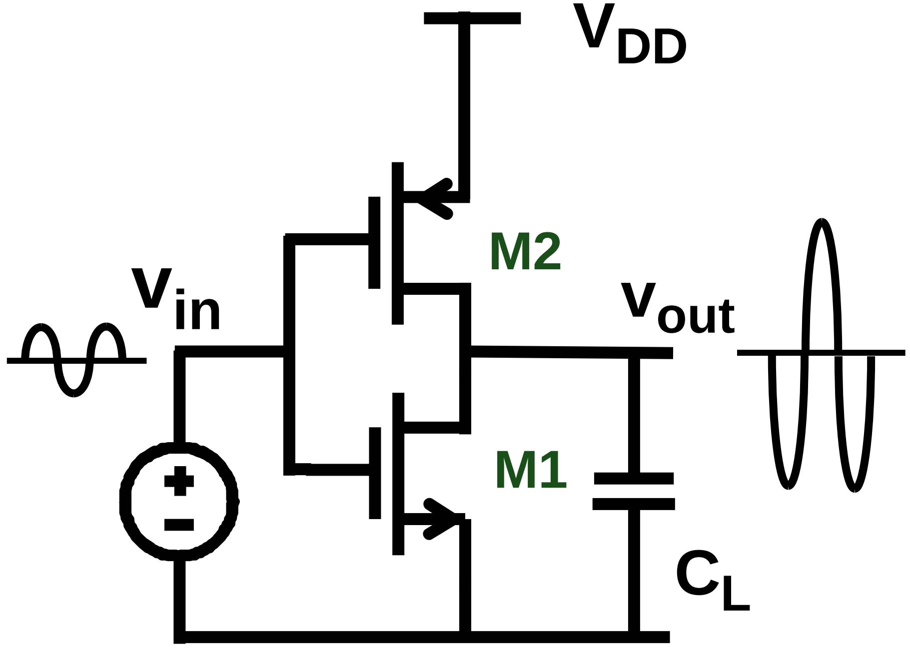


$v_{out} = A_v \, v_{in}$ $v_{out} = A_v \, v_{in}$

Class A stage **Class AB stage**

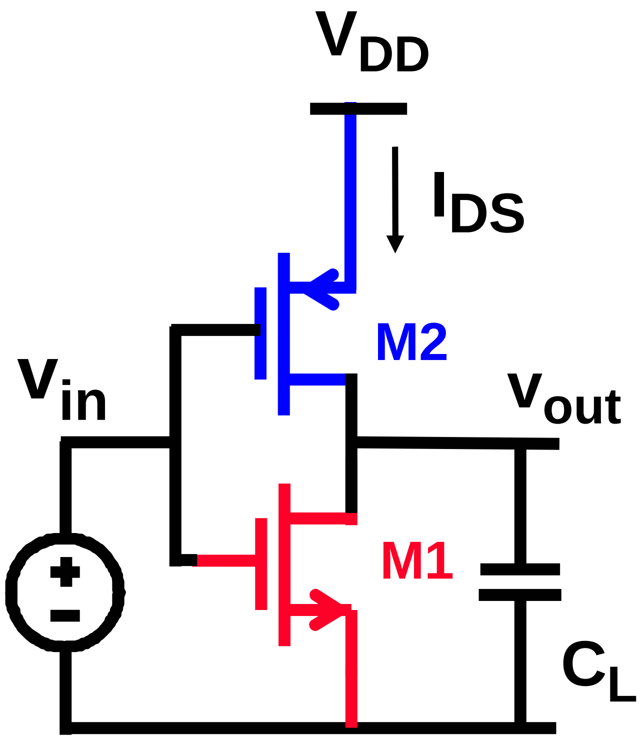
CMOS inverter-amplifier

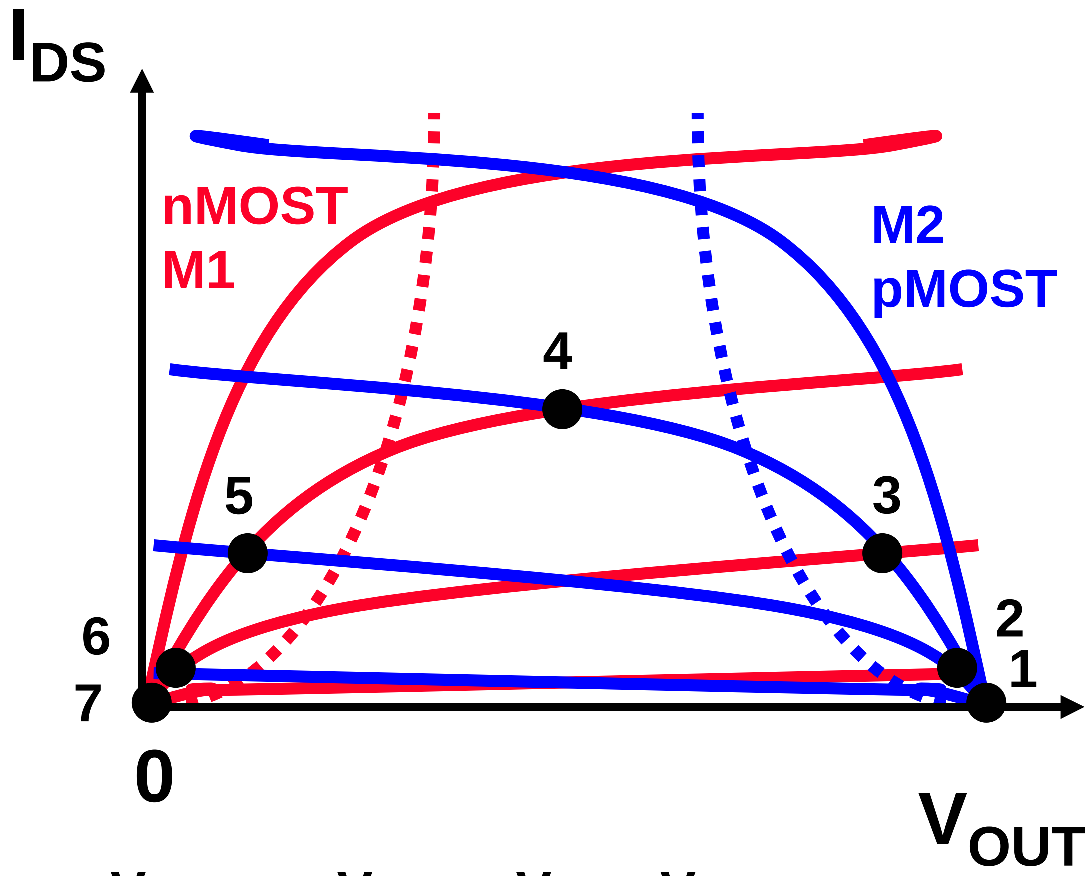




Digital invertor **Analog amplifier**

Operating points nMOST & pMOST





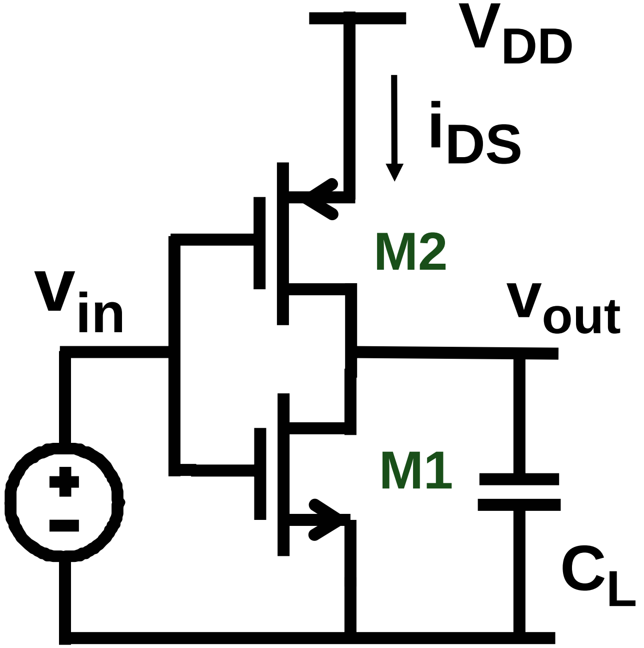
$V_{DD} = V_{DSn} + V_{DSp}$

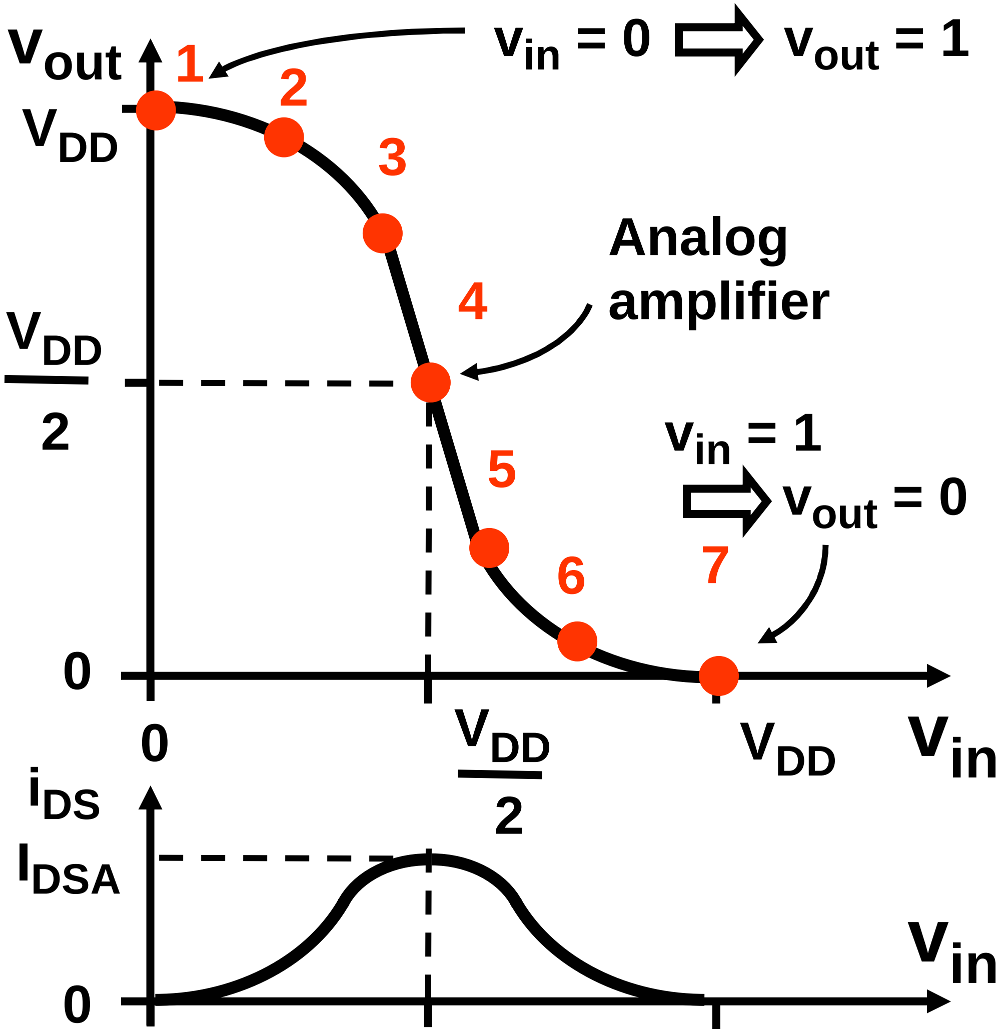
$= V_{GSn} + V_{GSp}$

$V_{DSn} = V_{OUT}$ $\quad V_{DSp} = V_{DD} - V_{OUT}$

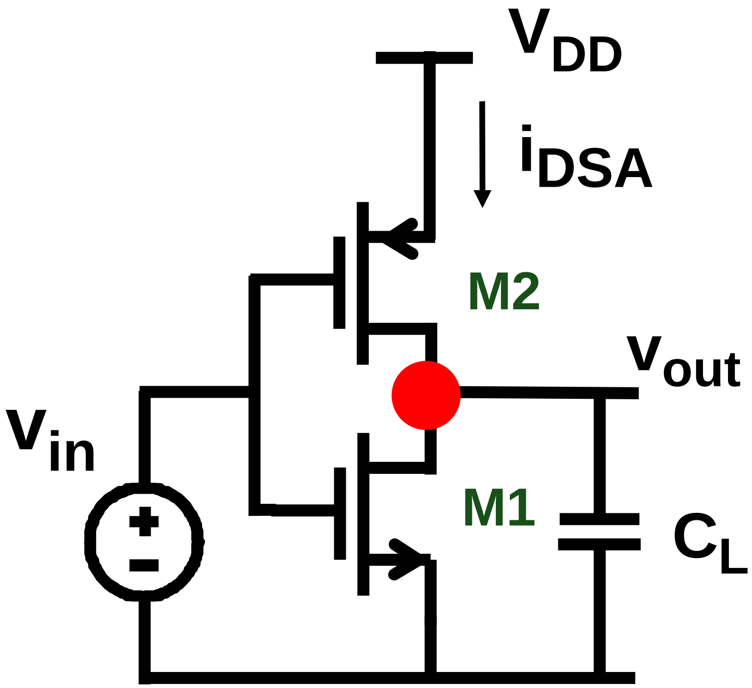
$V_{GSn} = V_{IN}$ $\quad V_{GSp} = V_{DD} - V_{IN}$

Transfer characteristic





Analog amplifier : DC



$$V_{in} = \frac{V_{DD}}{2} \Rightarrow V_{out} = \frac{V_{DD}}{2}$$

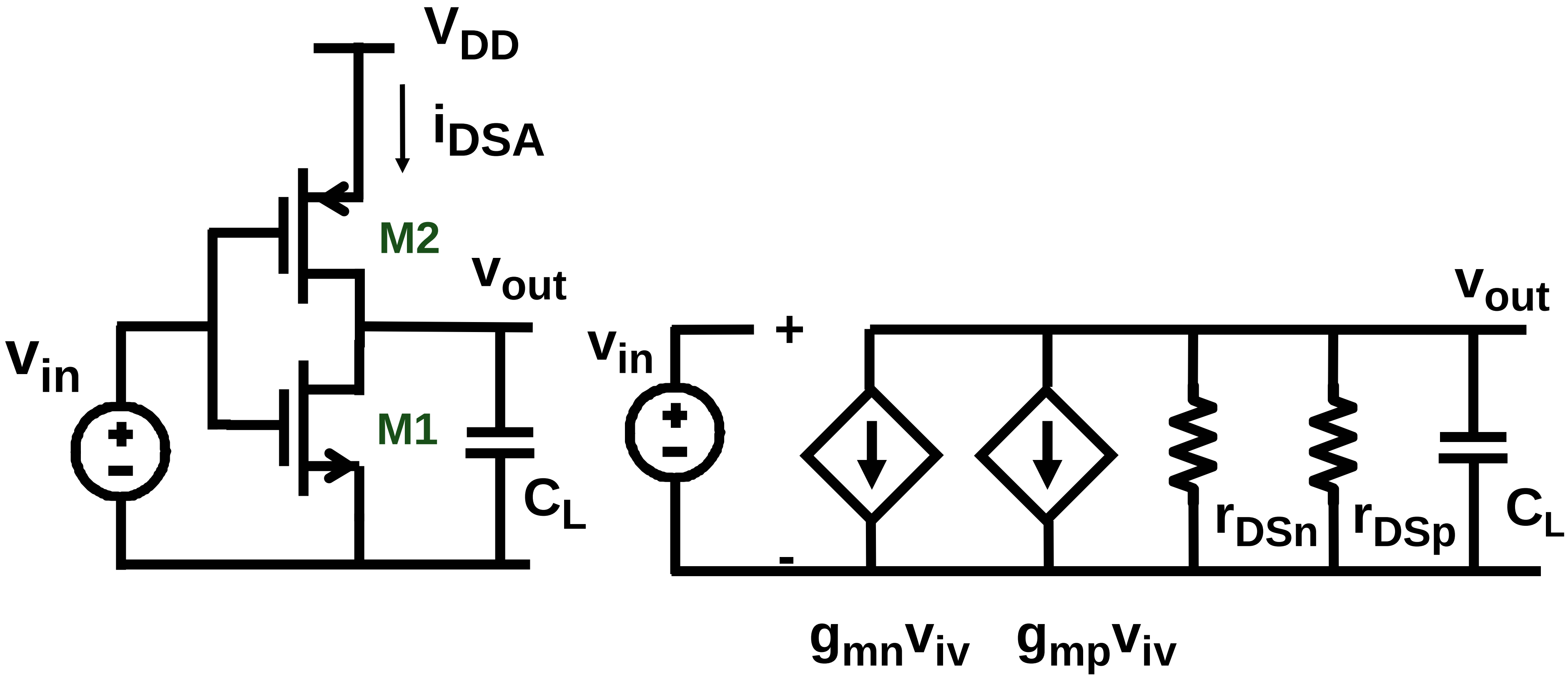
$$I_{DSn} = K'_n \frac{W_n}{L_n} (V_{in}-V_T)^2$$

$$I_{DSp} = K'_p \frac{W_p}{L_p} (V_{DD}-V_{in}-V_T)^2$$

$$\Rightarrow K'_n \frac{W_n}{L_n} = K'_p \frac{W_p}{L_p}$$

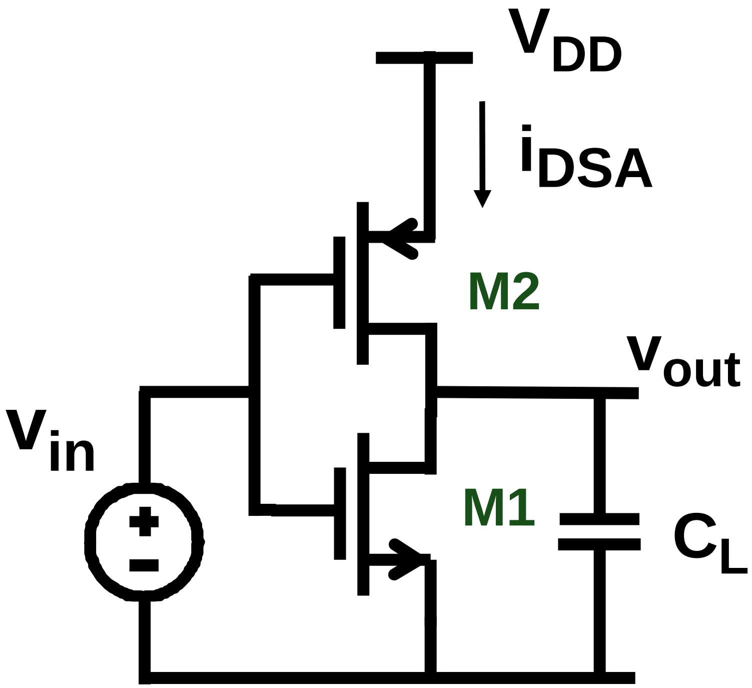
$$I_{DS} = K'_n \frac{W_n}{L_n} (\frac{V_{DD}}{2} -V_T)^2$$

Analog amplifier : AC model



For the same I_{DS} en V_{GS}-V_T : $g_{mn} = g_{mp} = g_m$

Analog amplifier: AC gain Av



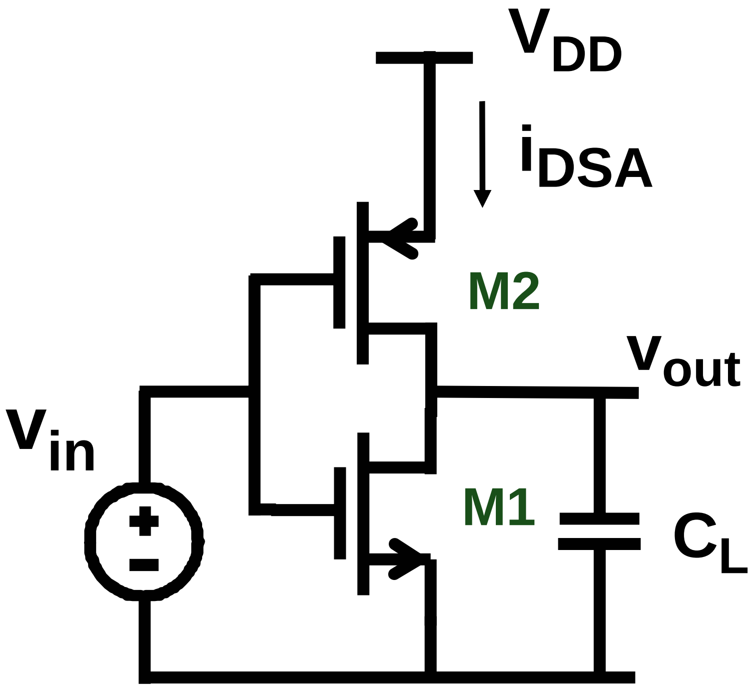
If $V_{En}L_n = V_{Ep}L_p = V_E$

$$g_{DSn} = g_{DSp} = g_{DS}$$

$$(g_{DS} = 1/r_{DS})$$

$$\boxed{A_{v0} = -\frac{2g_m}{2g_{DS}}} = -\frac{2V_E}{\frac{V_{DD}}{2} - V_T}$$

Analog amplifier : BW & GBW



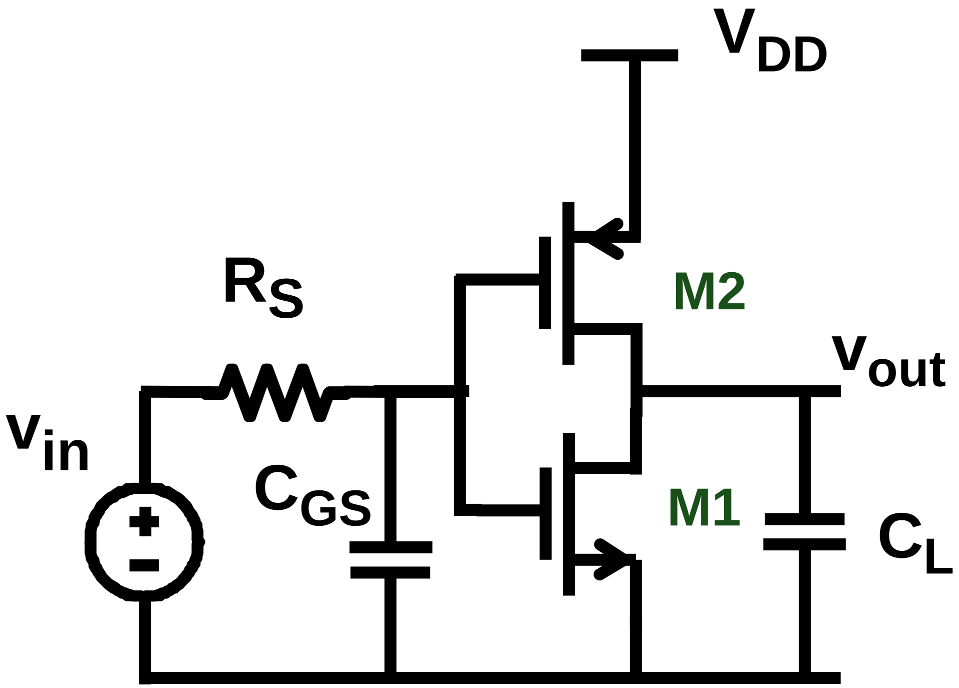
$$A_{v0} = 2g_m R_{out}$$

$$R_{out} = \frac{r_{DS}}{2}$$

$$BW = \frac{1}{2\pi R_{out} C_L}$$

$$\boxed{GBW = \frac{2g_m}{2\pi C_L}}$$

Analog amplifier: poles due to C_{GS}



$$A_{v0} = 2g_m R_{out}$$

主极点在输出

$$GBW = \frac{2g_m}{2\pi C_L}$$

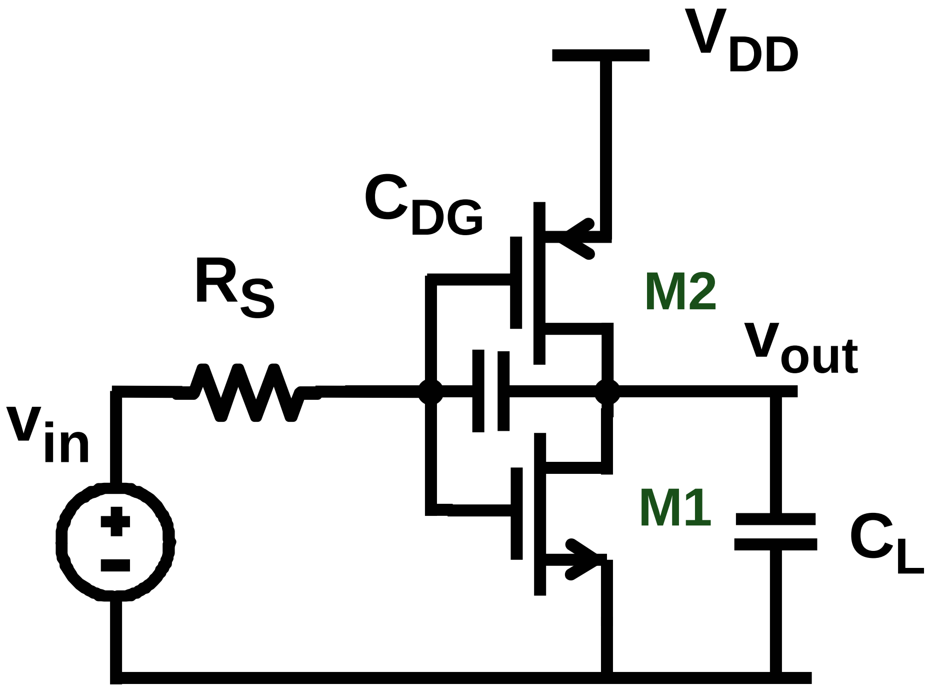
$$C_{GSt} = C_{GS1} + C_{GS2}$$

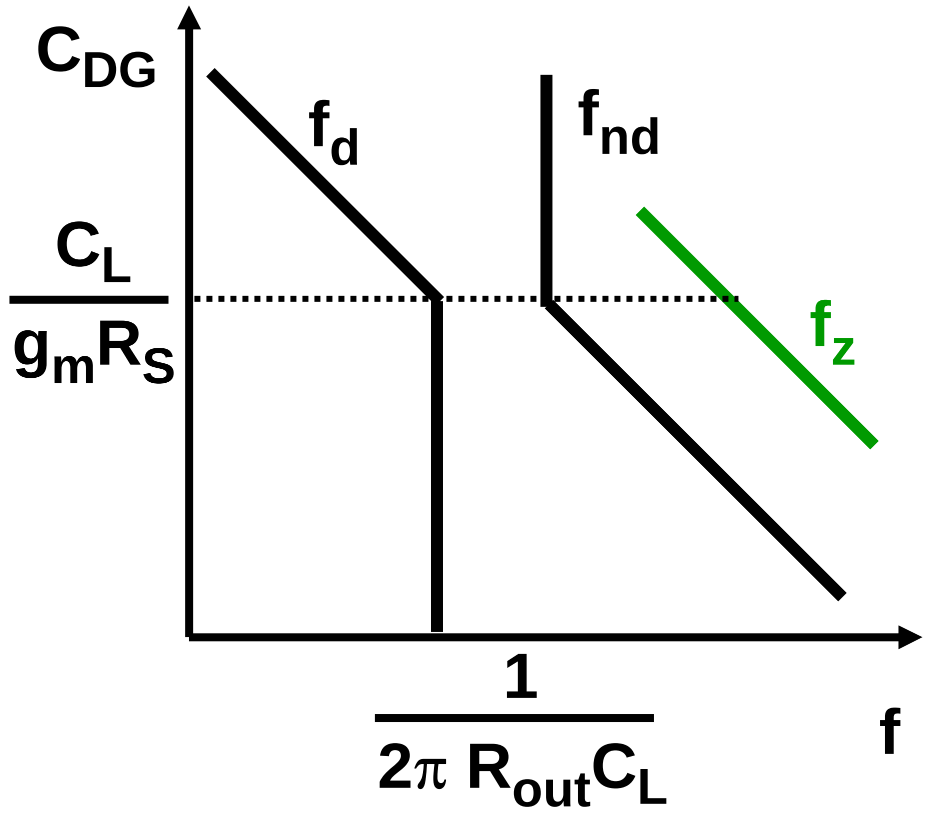
主极点在输入

But if $R_S C_{GSt} > r_{DS} C_L$:

$$GBW = f_T \frac{r_{DS}}{R_S}$$

Analog amplifier: poles due to C_{DG}





$$\frac{v_{out}}{v_{in}} = \frac{A_{v0}\,(1 - sC_{DGt}/g_m)}{1 + s\,(R_{out}C_L + A_{v0}R_SC_{DGt}) + s^2\,R_SR_{out}C_{DGt}C_L}$$

CC	p1	p2	z1
0.01p	-953.234k	-173.517x	6.57585g
0.02p	-936.042k	-159.742x	5.33966g
0.05p	-887.915k	-130.747x	3.41559g
0.1p	-817.635k	-103.439x	2.13459g
0.2p	-705.474k	-77.6965x	1.21987g
0.5p	-498.854k	-53.4500x	533.769x
1p	-334.643k	-42.9319x	275.513x
2p	-201.569k	-37.0767x	140.020x
5p	-91.8480k	-33.3556x	56.5660x
10p	-48.1491k	-32.0797x	28.3772x
20p	-24.6712k	-31.4351x	14.2123x
50p	-10.0173k	-31.0462x	5.69061x
100p	-5.03392k	-30.9162x	2.84626x

subckt

element	0:m1	0:m2
model	0:mn	0:mp
region	Saturati	Saturati
id	904.0387u	-904.0387u
ibs	-7.106e-21	5.533e-21
ibd	-6.5052a	9.8044a
vgs	2.5000	-2.5000
vds	2.4856	-2.5144
vbs	0.	0.
vth	851.3079m	-957.5862m
vdsat	1.0619	-1.1649
vod	1.6487	-1.5424
beta	1.0566m	810.7718u
gam eff	730.2303m	375.5742m
gm	844.9518u	949.0140u
gds	10.6160u	51.5473u
gmb	362.1461u	234.1548u
cdtot	13.1470f	28.5552f
cgtot	23.3467f	61.3342f
cstot	34.4337f	80.1351f
cbtot	38.0011f	75.1086f
cgs	19.7554f	55.7527f
cgd	1.8358f	5.0229f

从公式角度的分析

$$\omega_{p1}=\frac{1}{\left[\left(1+g_m r_{out}\right)C_{DG}+C_S\right]R_S+\left(C_{DG}+C_L\right)r_{out}}$$

$$\omega_{p2}=\frac{1}{\omega_{p1}R_S r_{out}K}=\frac{\left[\left(1+g_m r_{out}\right)C_{DG}+C_S\right]R_S+\left(C_{DG}+C_L\right)r_{out}}{R_S r_{out}\left(C_S C_{DG}+C_S C_L+C_{DG}C_L\right)}$$

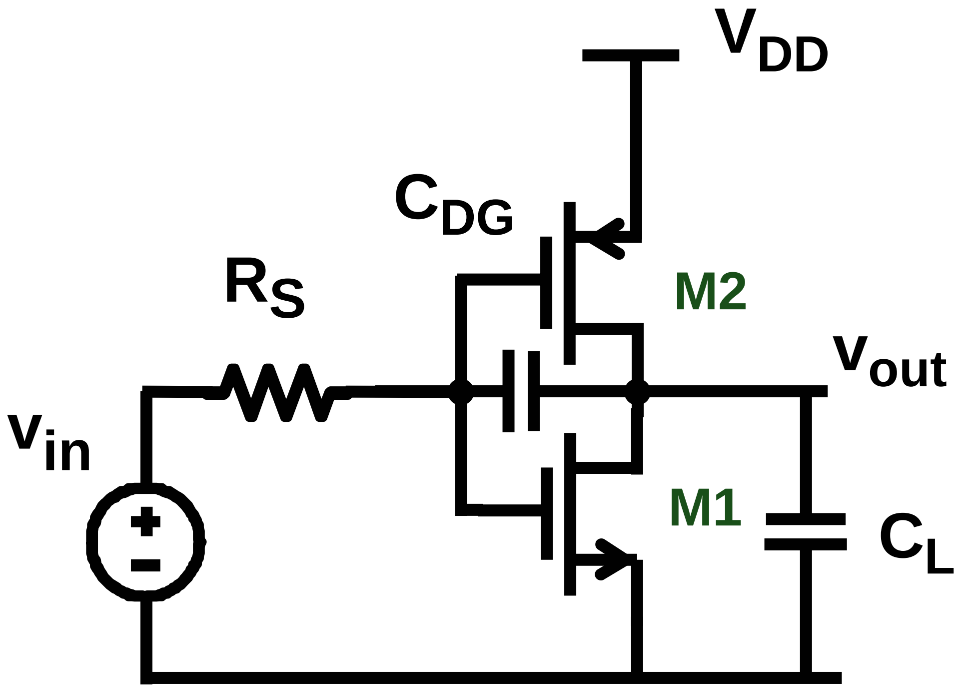
$$\omega_{pd}=\frac{1}{\left(C_{DG}+C_L\right)r_{out}}$$

$$\omega_{pd}=\frac{1}{\left[\left(1+g_m r_{out}\right)C_{DG}+C_S\right]R_S}$$

$$\omega_{pnd}\approx\frac{1}{R_S\left(C_S+C_{DG}\right)}$$

$$\omega_{pnd}\approx=\frac{g_m}{\left(1+\frac{C_S}{C_{DG}}\right)C_L}$$

Analog amplifier: other poles



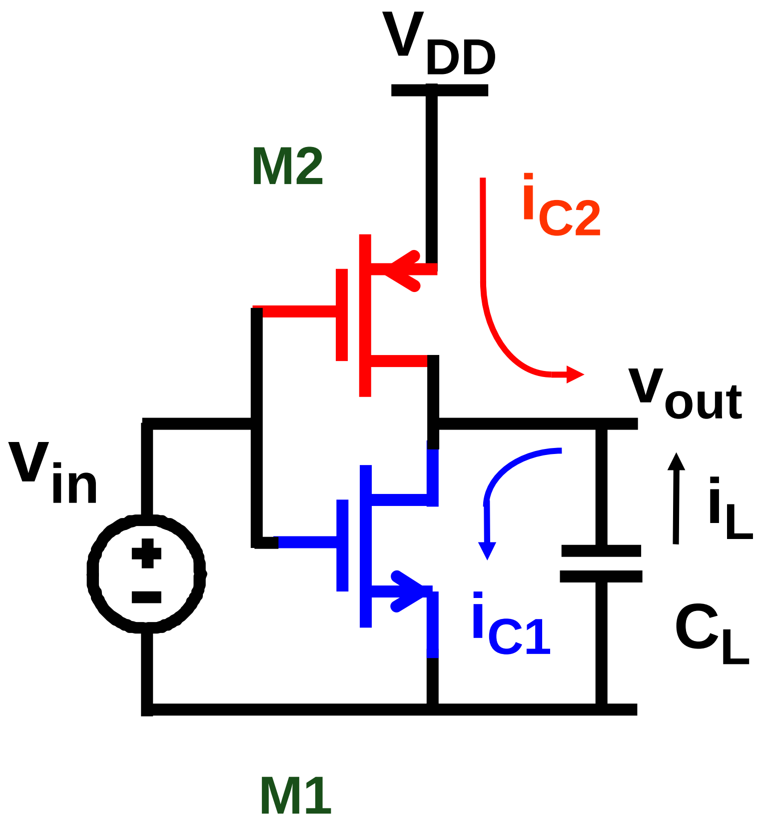
$$A_{v0} = 2g_m R_{out}$$

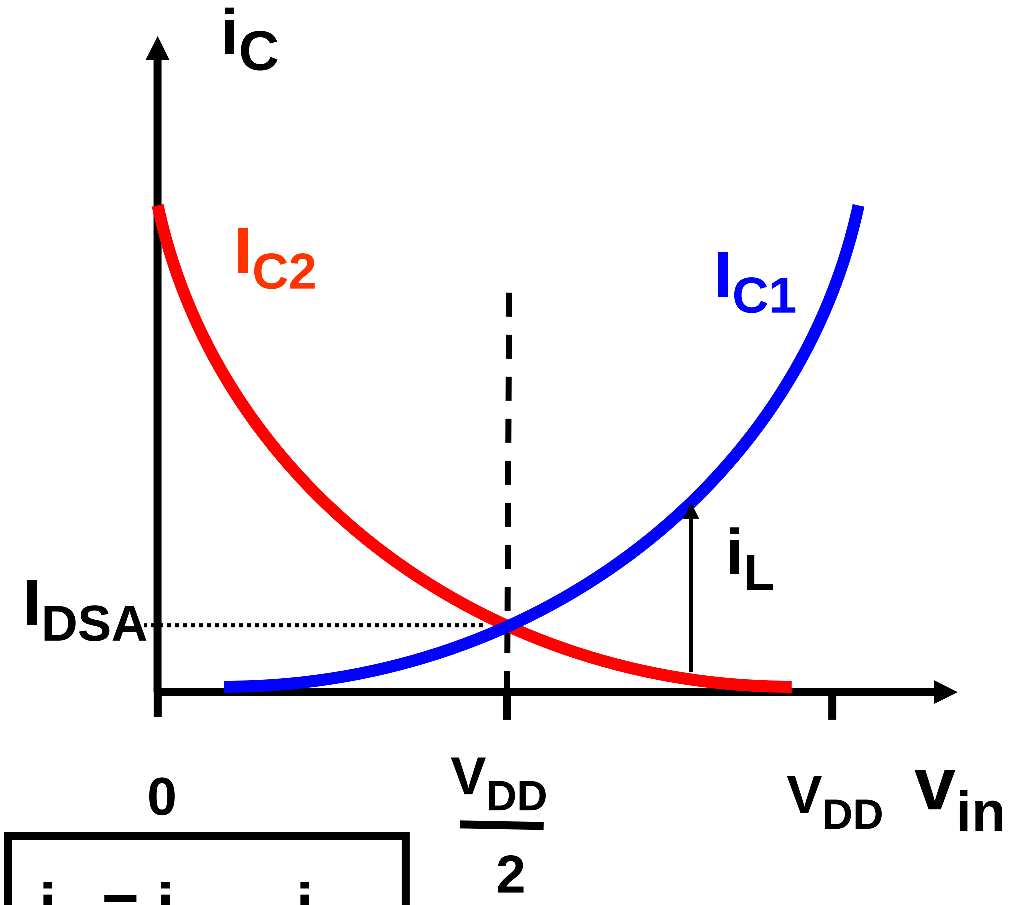
$$GBW = \frac{2g_m}{2\pi\ C_L}$$

$$C_{DGt} = C_{DG1} + C_{DG2}$$

But if $R_S C_{DGt} > \dfrac{1}{2\pi\ GBW}$: $\cancel{G}BW = \dfrac{1}{2\pi\ R_S\ C_{DGt}}$

Class AB operation





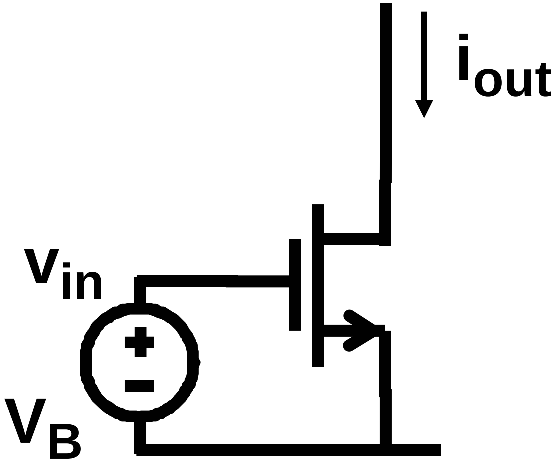
$$i_L = i_{C2} - i_{C1}$$

Table of contents

- ❑ Single-transistor amplifiers
- ❑ Source followers
- ❑ Cascodes

Single-transistor stages

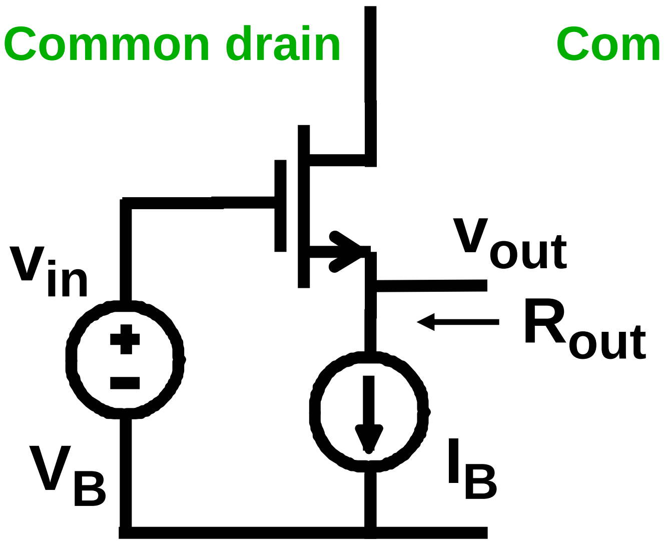
Common source



$$i_{out} = g_m v_{in}$$

Amplifier

Common drain



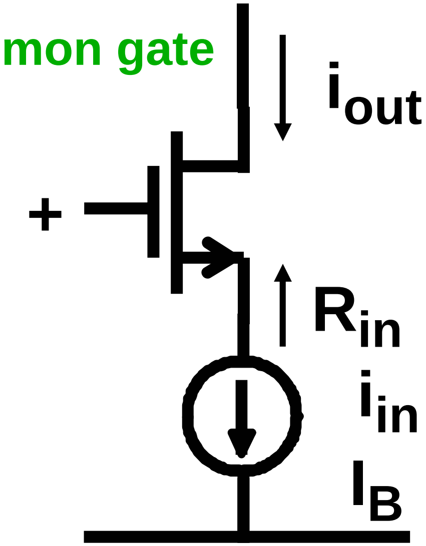
$$v_{out} = v_{in}$$

$$R_{out} \approx 1/g_m$$

Source follower

Voltage buffer

Common gate



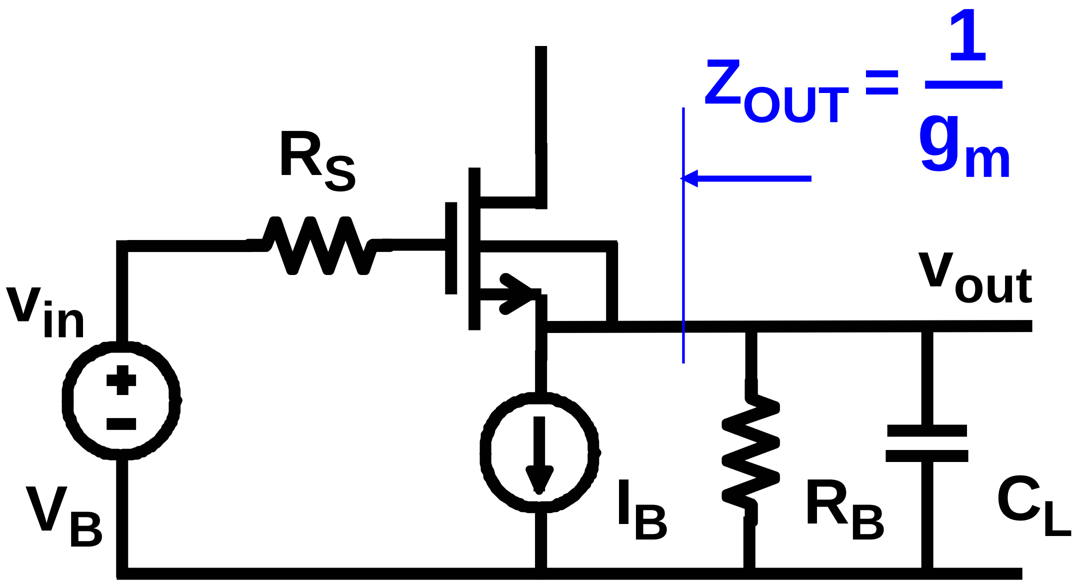
$$i_{out} = i_{in}$$

$$R_{in} \approx 1/g_m$$

Cascode

Current buffer

Source follower with V_{BS} = 0 (p-well)



$$V_{GS} = V_{T0} + \sqrt{\frac{I_B}{K'W/L}}$$

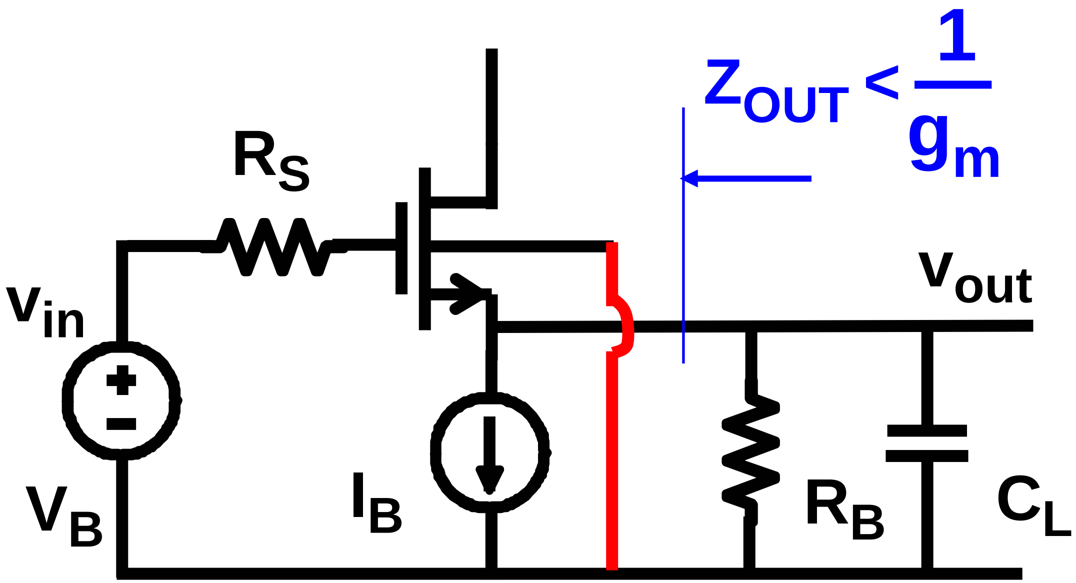
$$V_{GS} = ct \quad \text{if } I_B = ct$$

$$V_{OUT} = V_{IN} - V_{GS}$$

$$\Delta V_{OUT} = \Delta V_{IN}$$

$$\boxed{A_v = 1}$$

Source follower with $V_{BS} \neq 0$ (n-well)



$$V_{GS} = V_T + \sqrt{\frac{I_B}{K'W/L}}$$

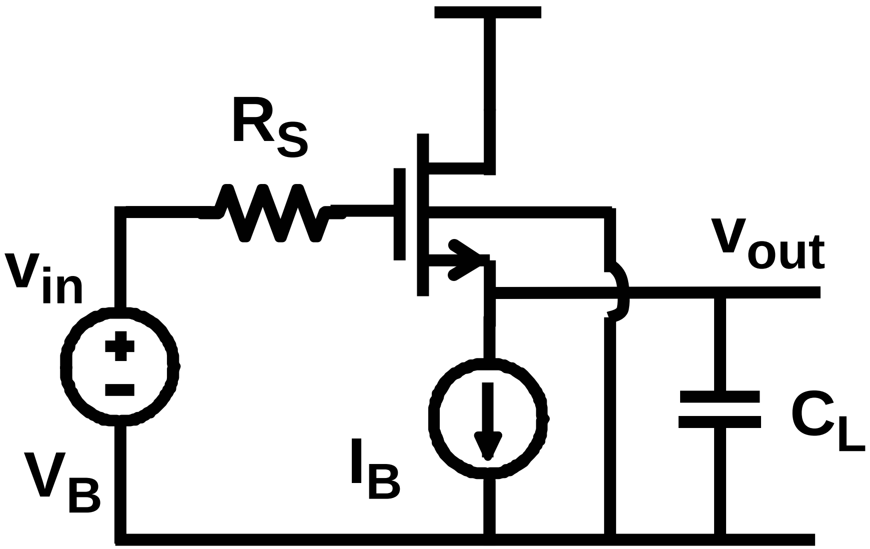
$$V_{GS} \neq ct$$

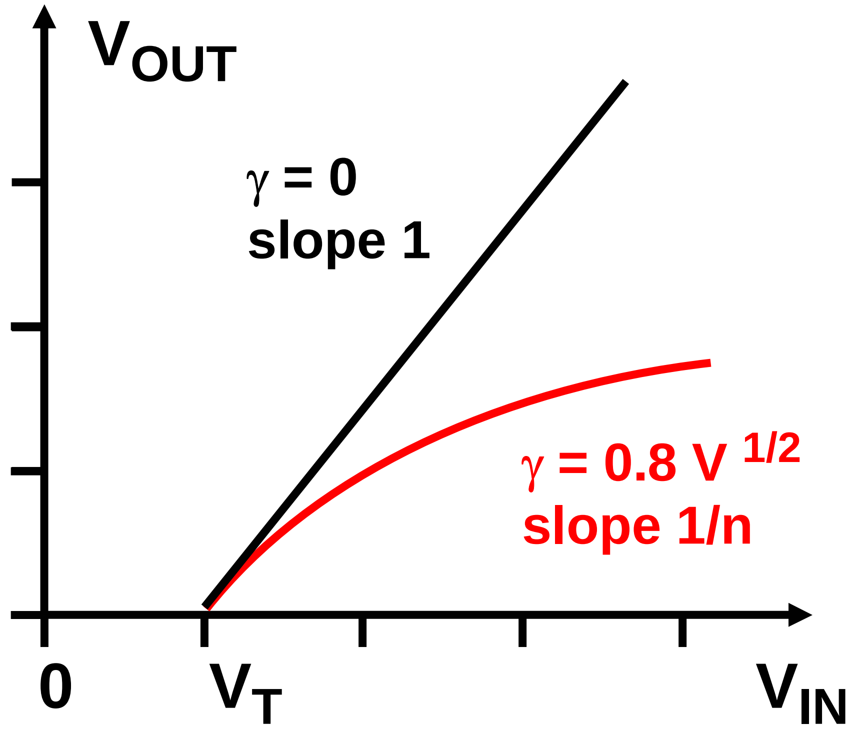
$$V_{OUT} = V_{IN} - V_{GS}$$

$$V_T = V_{T0} + \gamma \left[\sqrt{|2\Phi_F| + V_{OUT}} - \sqrt{|2\Phi_F|} \right]$$

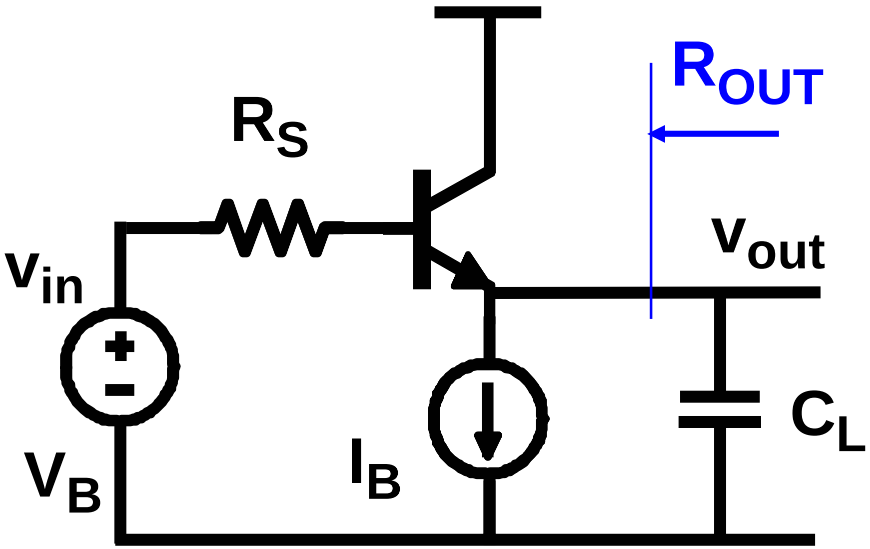
$$\boxed{A_v = \frac{1}{n}}$$

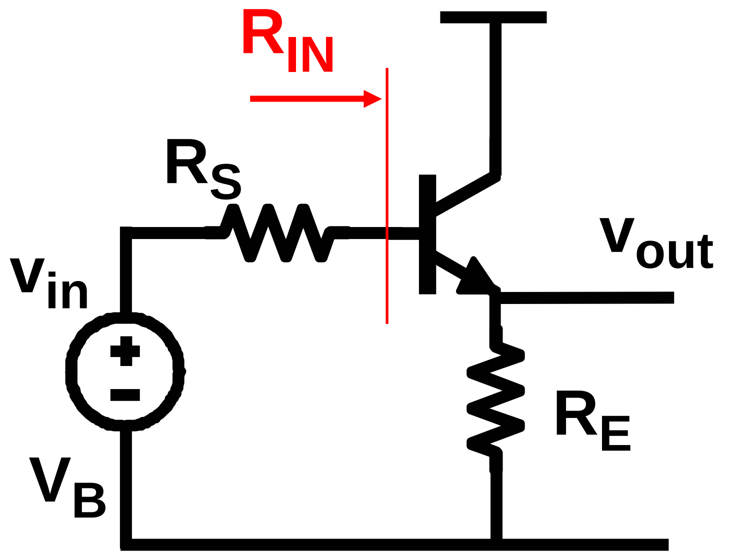
Source follower non-linearity





Emitter follower

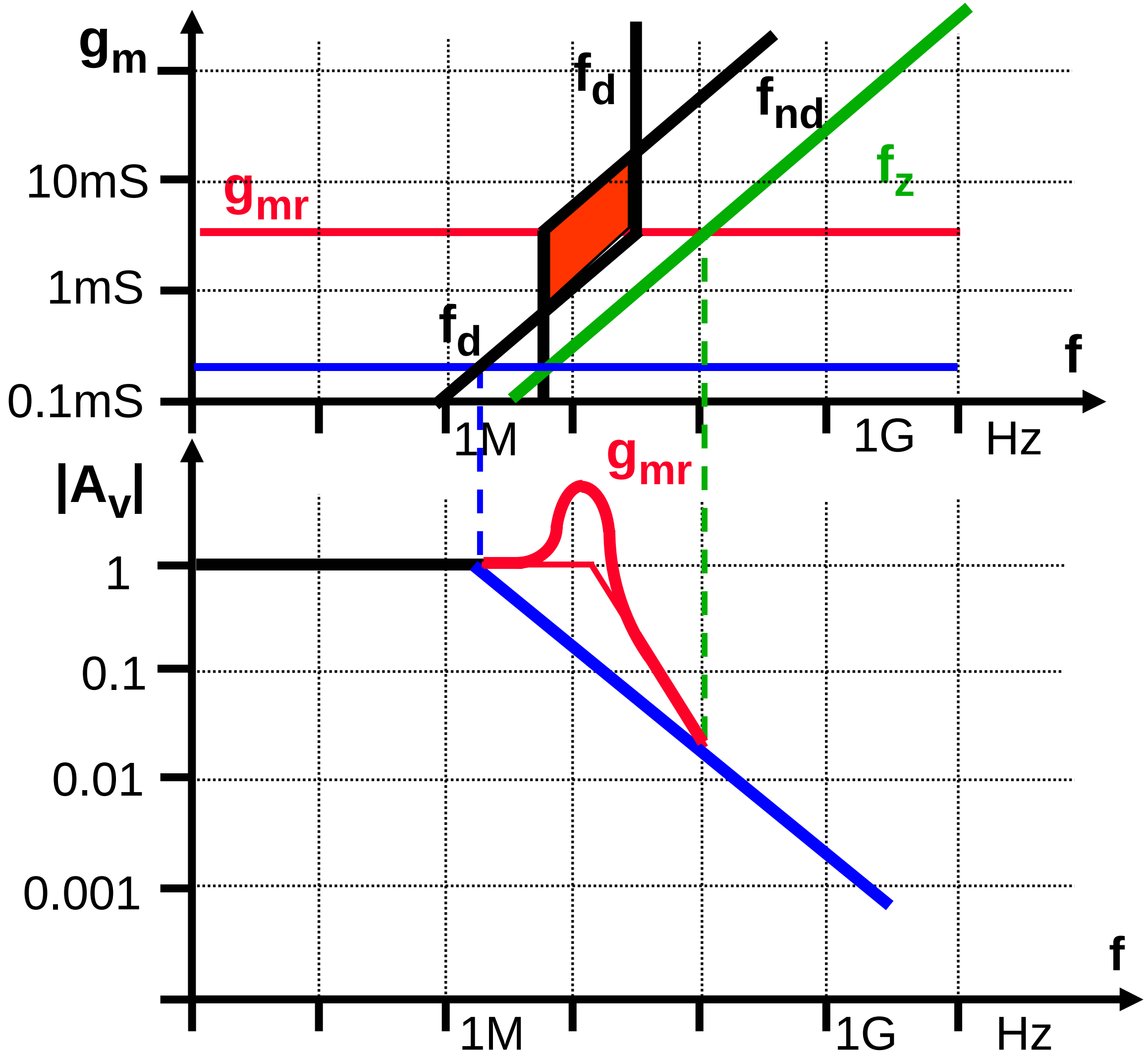




$\boxed{A_v = 1}$ $\quad R_{OUT} = \dfrac{1}{g_m} + \dfrac{R_S + r_B}{\beta + 1}$ $\quad R_{IN} = r_\pi + r_B + (\beta + 1)R_E$

Limited isolation !

Source follower with C_L load



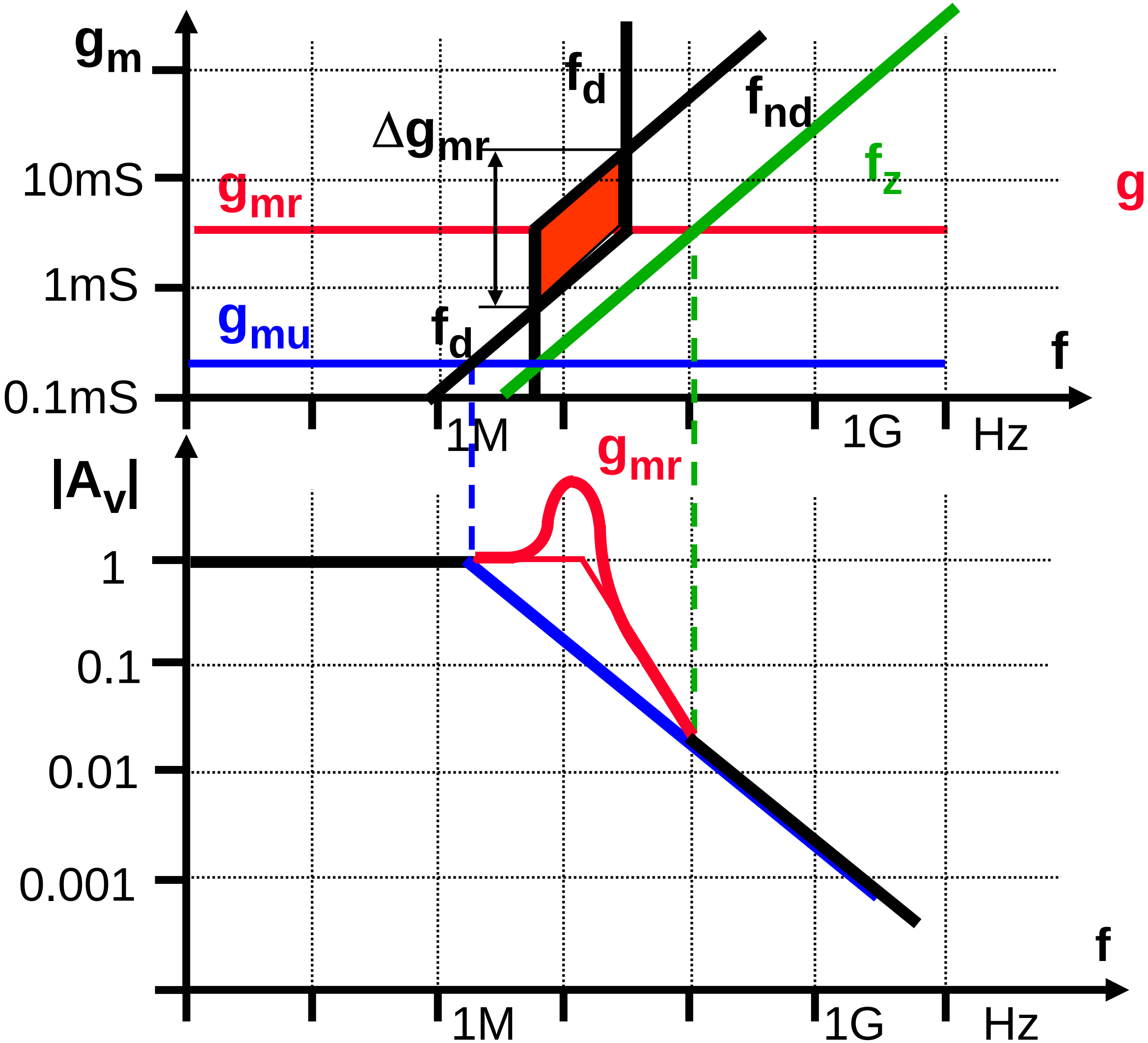
$$A_v = \frac{(1 + s\,C_{GS}/g_m)}{1 + s\,B + s^2\,C^2 R_S/g_m}$$

$$B = R_S C_{DG} + \frac{C'_{DS}}{g_m} + \frac{C_{GS}}{g_m}\left(1 + \frac{R_S}{r_{DS}}\right)$$

$$C^2 = C'_{DS}C_{DG} + C'_{DS}C_{GS} + C_{DG}C_{GS}$$

$$C'_{DS} = C_L + C_{DS}$$

Source follower with C_L load



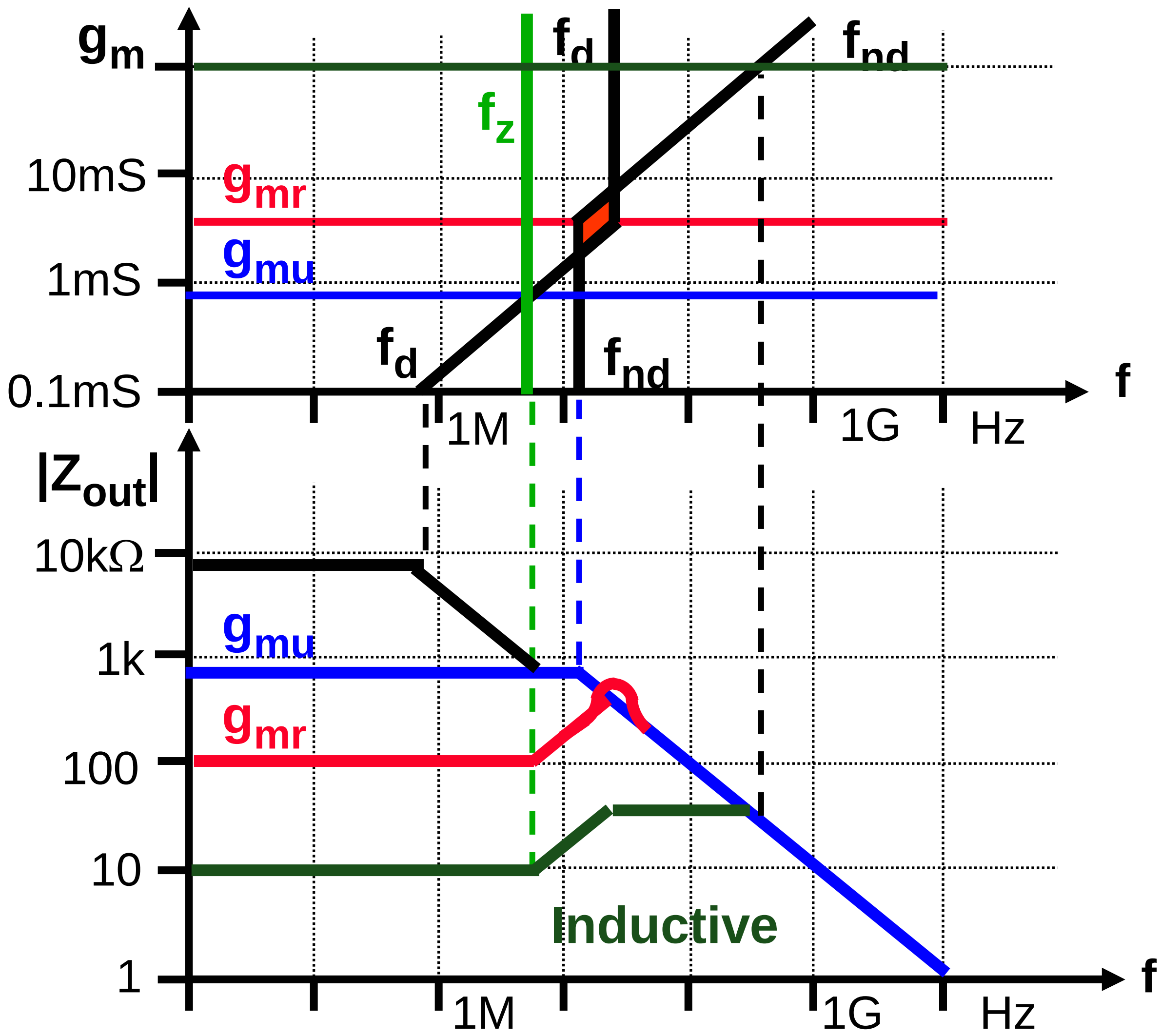
$$g_{mr} = \frac{1}{R_S} \frac{C_L + C_{DS} + C_{GS}}{C_{DG}}$$

$$\Delta g_{mr} = \frac{C_{DGt}}{C_{DG}}$$

$$C_{DGt} = \frac{C'_{DS} C_{GS}}{C'_{DS} + C_{GS}}$$

$$g_{mu} = \frac{1}{R_S}$$

Source follower : Output impedance



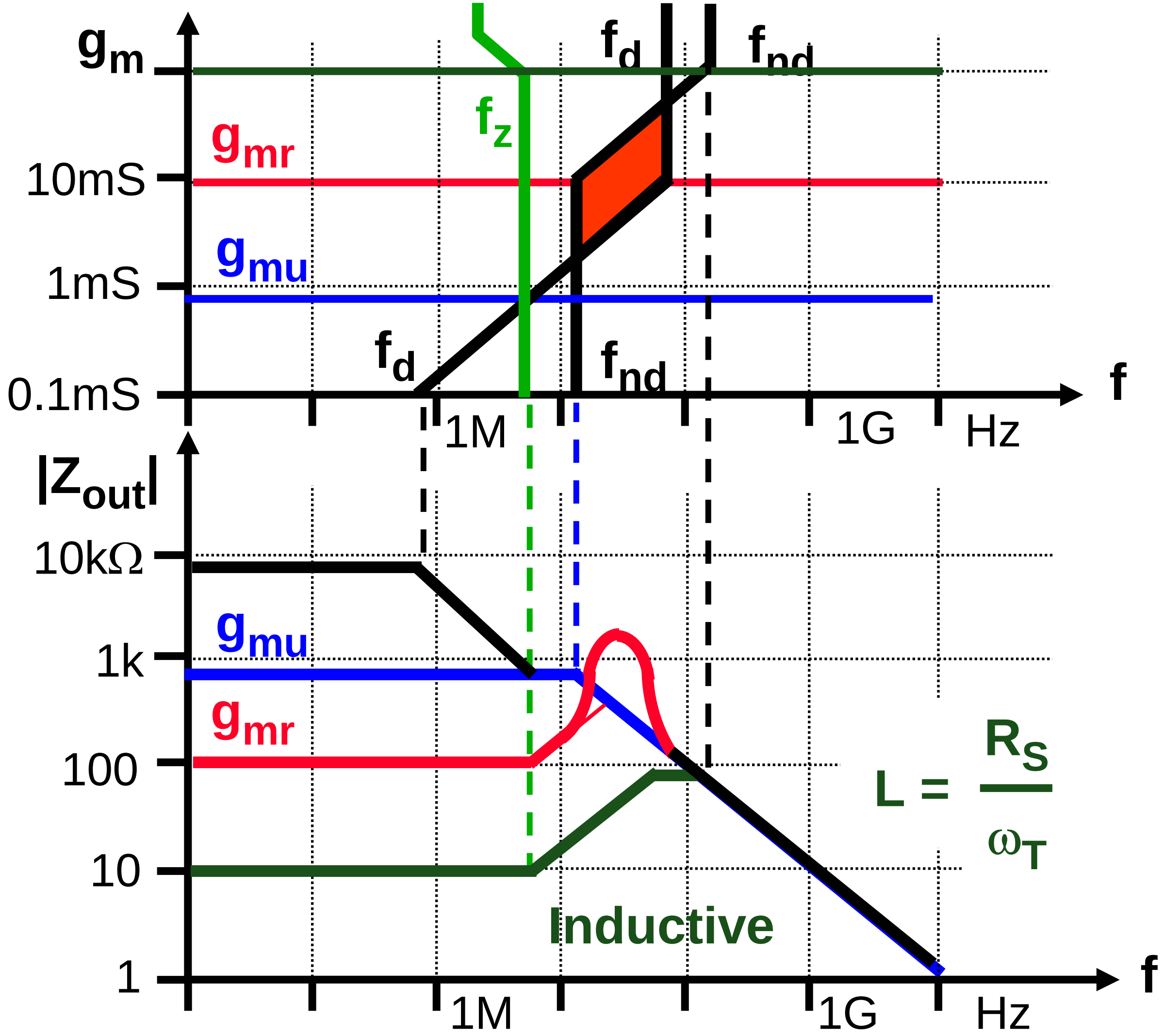
$$g_{mr} = \frac{1}{R_S} \frac{C_{GS} + C_{DS}}{C_{DG}}$$

$$g_{mu} \approx \frac{1}{R_S} \frac{C_{GS} + C_{DS}}{C_{GS} + C_{DG}}$$

$$f_z = \frac{1}{2\pi R_S C_{GS}}$$

$$f_{d,higm} = \frac{1}{2\pi R_S C_{DG}}$$

Emitter follower : Output impedance



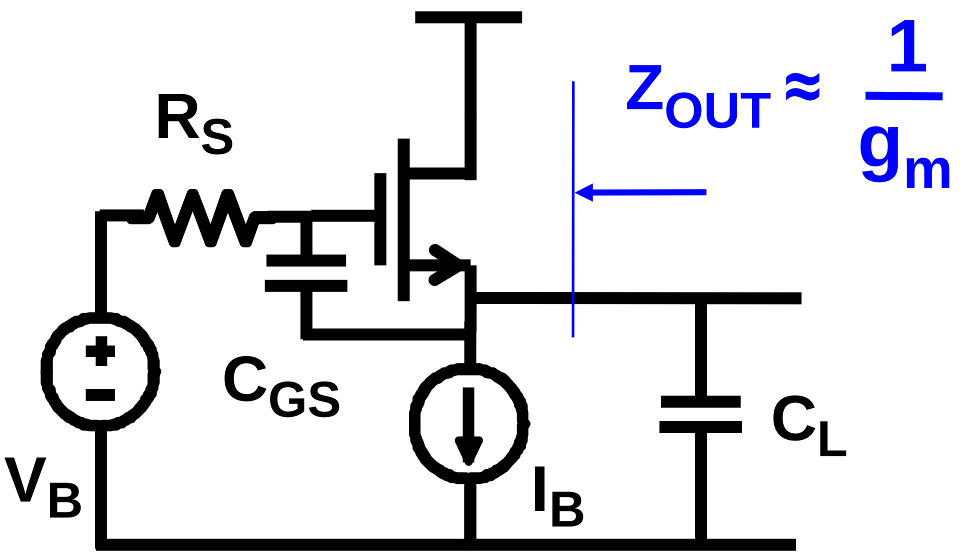
$$L = \frac{R_S}{\omega_T}$$

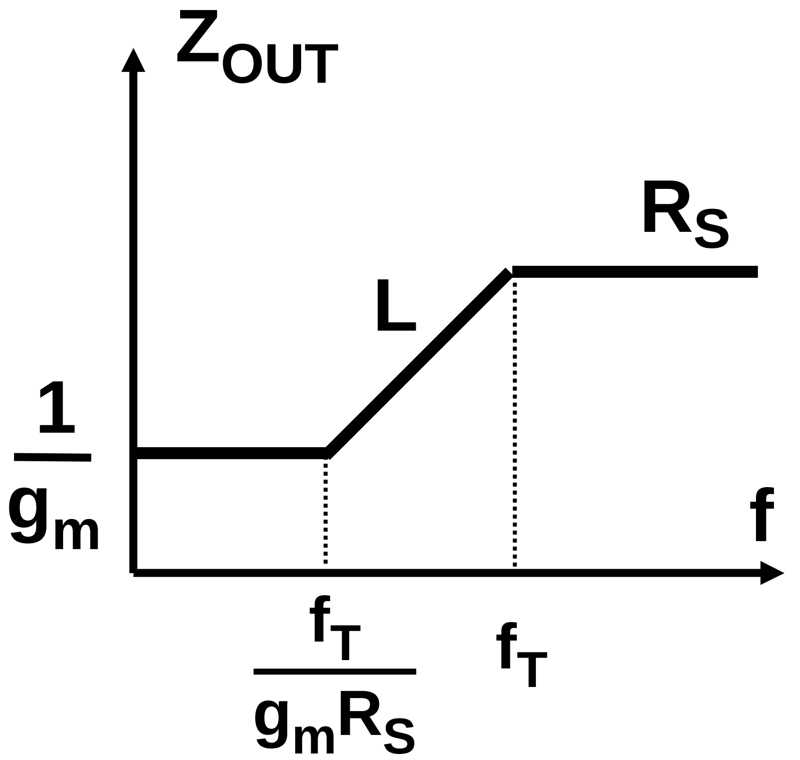
$$g_{mr} = \frac{1}{R_S}\,\frac{C_\pi + C_{CE}}{C_\pi + C_\mu}$$

$$g_{mu} \approx \frac{1}{R_S}\,\frac{C_{jE} + C_{CE}}{C_{jE} + C_\mu}$$

$$f_z = \frac{1}{2\pi\, R_S//r_\pi\,(C_\pi + C_\mu)}$$

Source follower as active L



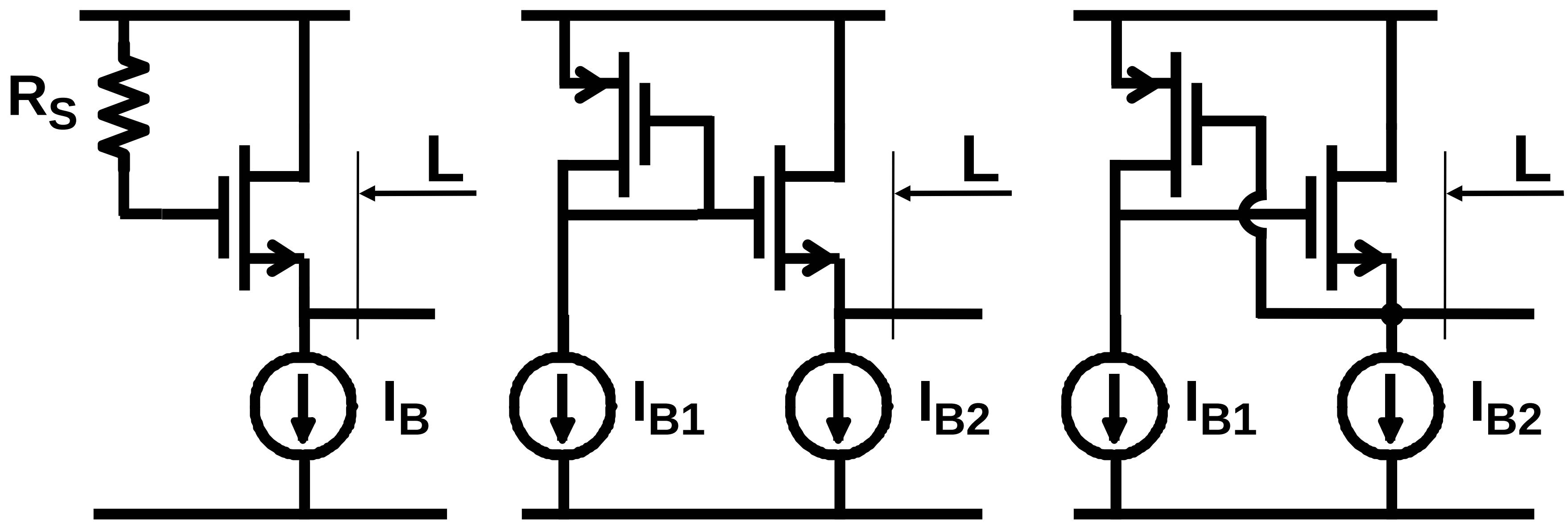


$$\boxed{Z_{out} = \frac{g_s + sC_{GS}}{g_m + sC_{GS}} \cdot R_S}$$

$$L \approx \frac{R_S}{2\pi\, f_T} \quad \text{up to} \quad f_T = \frac{g_m}{2\pi\, C_{GS}}$$

Razavi，6.3节

Source follower as active L



$$L \approx \frac{R_S}{2\pi f_T}$$

$$L \approx \frac{1/g_{mp}}{2\pi f_{Tn}}$$

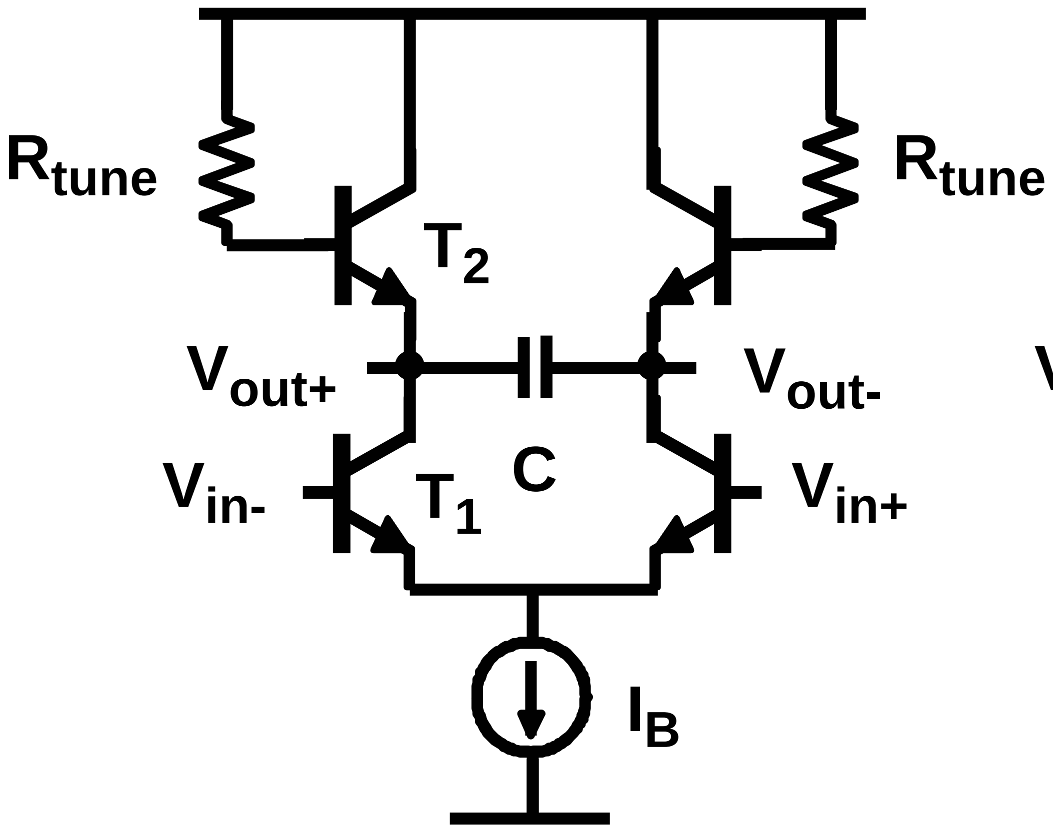
$$L \approx \frac{1/g_{mp}}{2\pi f_{Tn}}$$

$V_{DSn} = V_{GSn}$

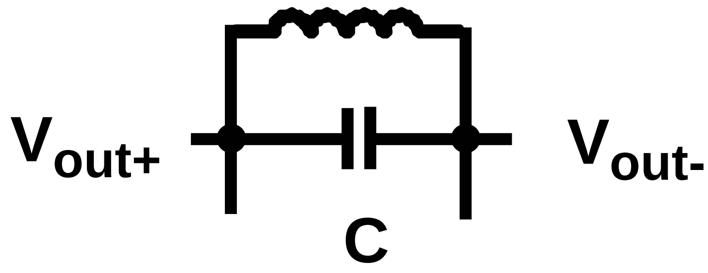
$V_{DSn} = V_{GSn} + V_{GSp}$

$V_{DSn} = V_{GSp}$

Floating inductor with parallel C



$$L = \frac{R_{tune}}{\omega_{T2}}$$



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

with HF peaking !

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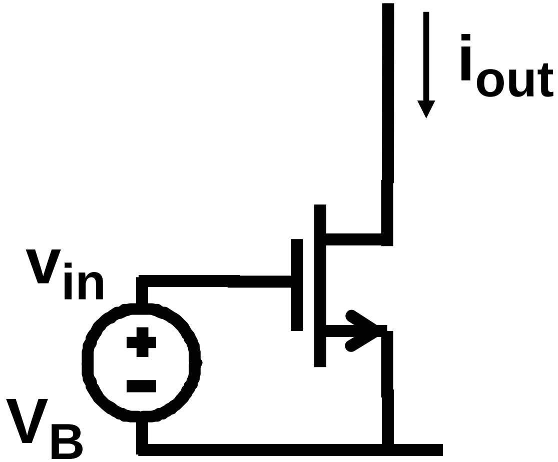
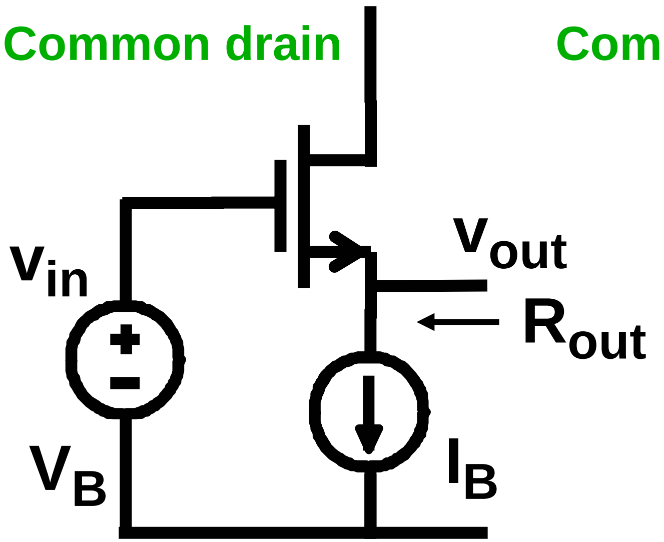
Single-transistor stages

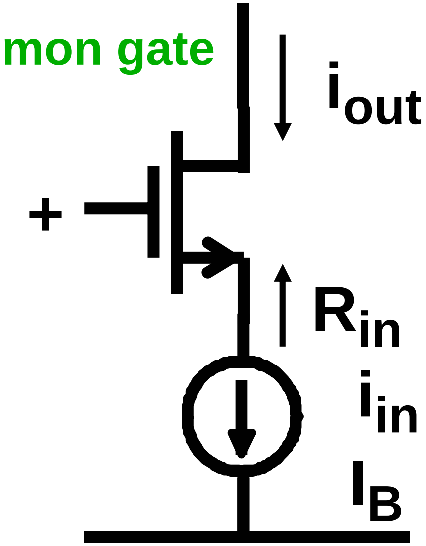
Common source



Common drain



Common gate



$i_{out} = g_m v_{in}$

$v_{out} = v_{in}$

$R_{out} \approx 1/g_m$

$i_{out} = i_{in}$

$R_{in} \approx 1/g_m$

Amplifier

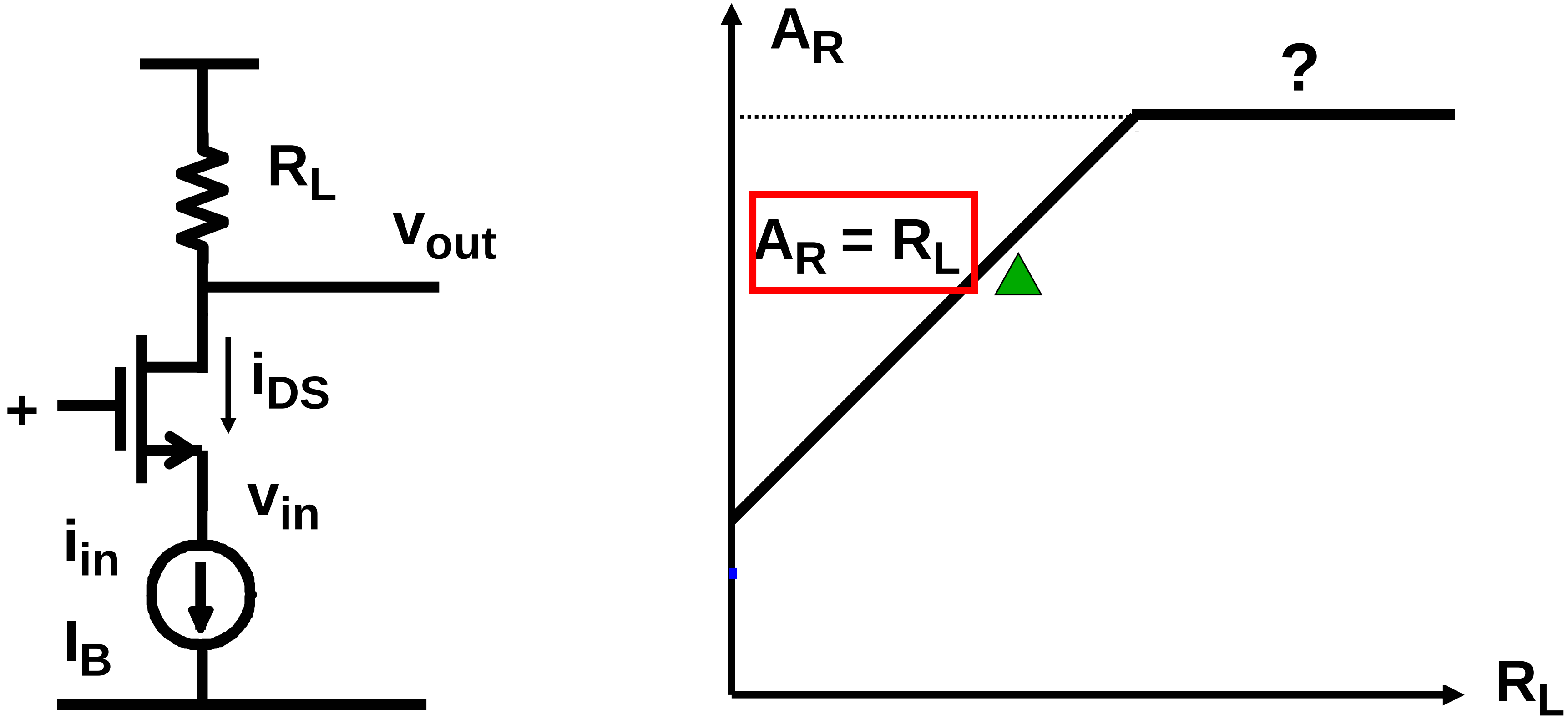
Source follower

Voltage buffer

Cascode

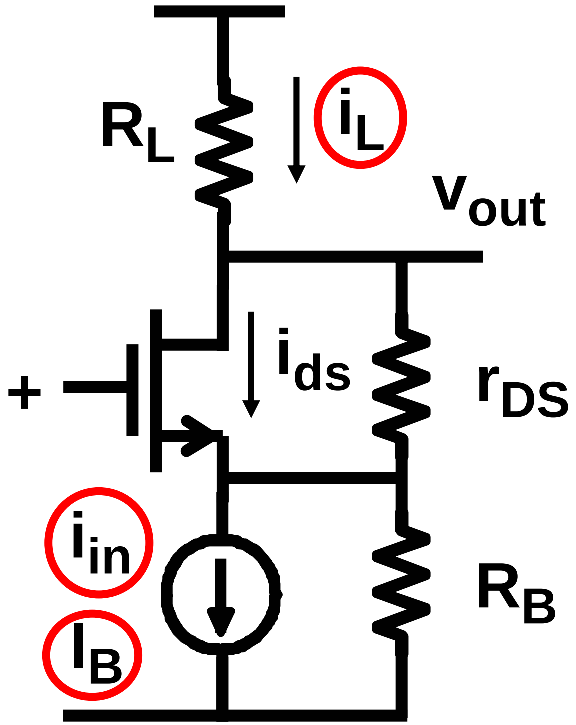
Current buffer

Cascode with resistive load



$$A_R = \frac{v_{out}}{i_{in}}$$

Cascode with resistive load



低频交流指标-输入电阻

$$\boxed{r_{in}=\frac{v_{in}}{i_{in}}=\frac{R_L+r_o}{\left(g_m+g_{mb}\right)r_o+1}}$$

共栅放大器的输入电阻随着负载电阻的增大而增大

$R_L \to 0$

$$r_{in}=\frac{R_L+r_o}{\left(g_m+g_{mb}\right)r_o+1}\approx\frac{1}{g_m+g_{mb}}$$

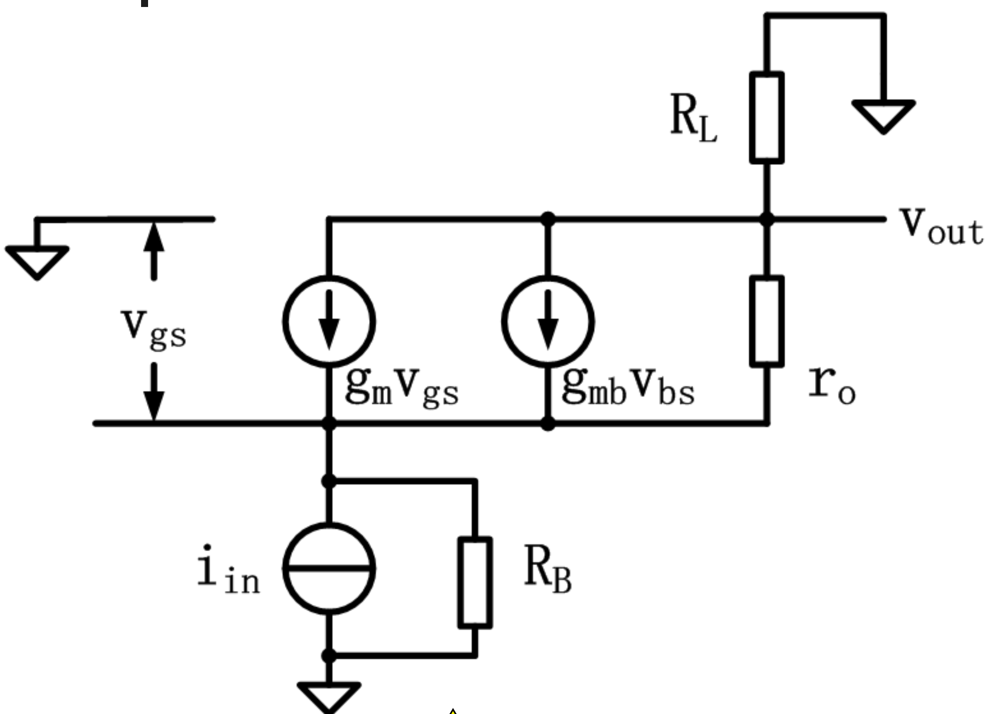
$R_L \to \infty$

$$r_{in}=\frac{R_L+r_o}{\left(g_m+g_{mb}\right)r_o+1}\to\infty$$

$R_L = Kr_o(0<K<\infty)$

$$\boxed{r_{in}=\frac{v_{in}}{i_{in}}=\frac{\left(K+1\right)r_o}{\left(g_m+g_{mb}\right)r_o+1}\approx\frac{K+1}{g_m+g_{mb}}}$$

低频交流指标-直流跨阻增益



- 输入电流源没有内阻

$$\frac{v_{out}}{i_{in}} = R_L$$

- 输入电流源有内阻

$$\frac{v_{out}}{i_{in}} = \frac{R_B}{R_B + r_{in}} R_L$$

低频交流指标-直流电压增益

- 输入电流源没有内阻

$$\frac{v_{out}}{i_{in}} = R_L \quad v_{in} = r_{in} \cdot i_{in} \quad \frac{v_{out}}{v_{in}} = \frac{R_L}{r_{in}} = \frac{g_m r_o + 1}{1 + \frac{r_o}{R_L}}$$

- 输入电流源有内阻
  - 戴维南-诺顿变换

$$\frac{v_{out}}{i_{in}} = \frac{R_B}{R_B + r_{in}} R_L \quad v_{in} = R_B \cdot i_{in} \quad \frac{v_{out}}{v_{in}} = \frac{R_L}{R_B + r_{in}} = \frac{R_L\left(g_m r_o + 1\right)}{R_B\left(g_m r_o + 1\right) + R_L + r_o}$$

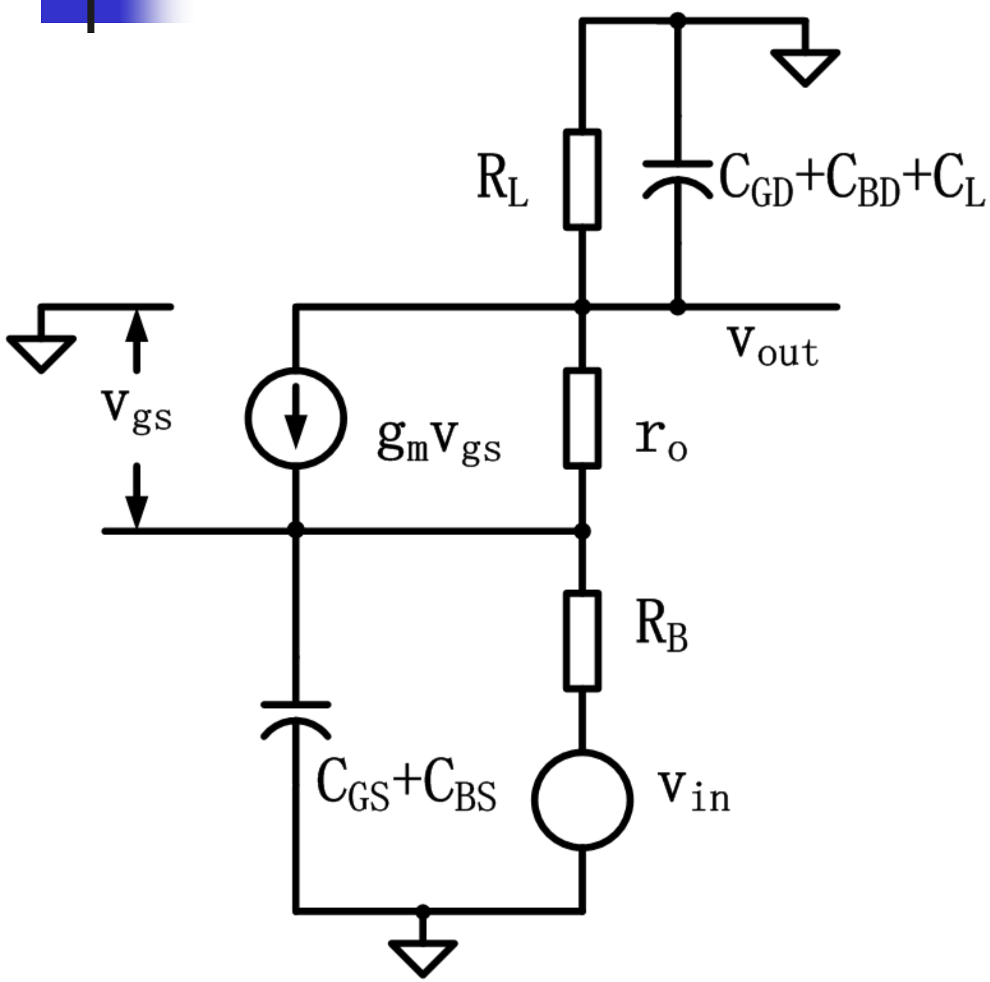
低频交流指标-输出电阻（电压放大器）

- 输入电流源没有内阻
  - 向上看：R_L
  - 向下看：r_o
- 输入电流源有内阻
  - 向上看： R_L
  - 向下看：

$$\boxed{r_{out}=\left[\left(g_m+g_{mb}\right)r_o R_B+R_B+r_o\right]}$$

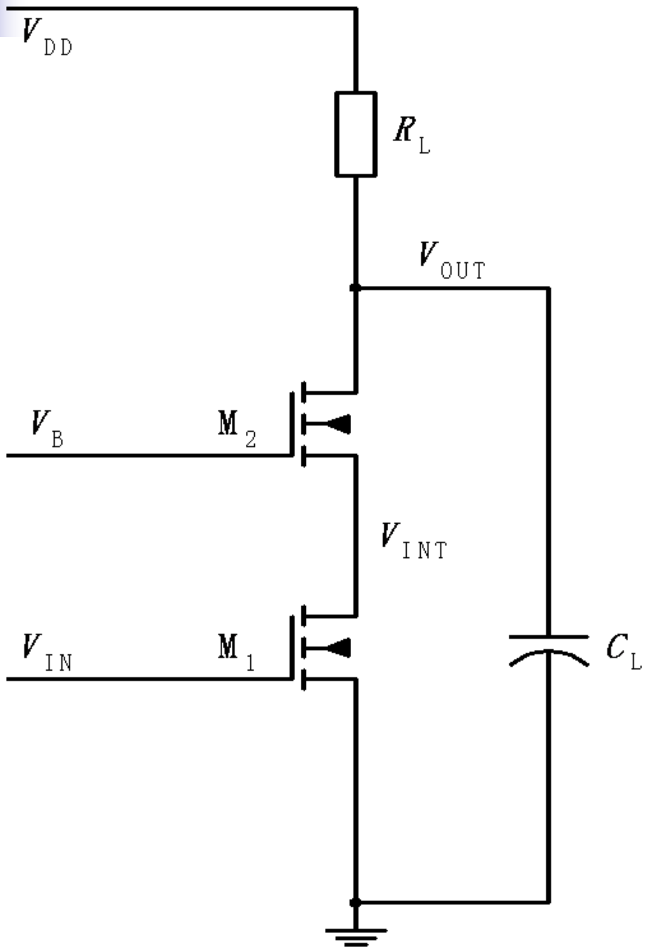
输出电阻大约提高$g_m r_o$倍，正是MOS管本征增益

高频交流指标-BW、GBW、PZ



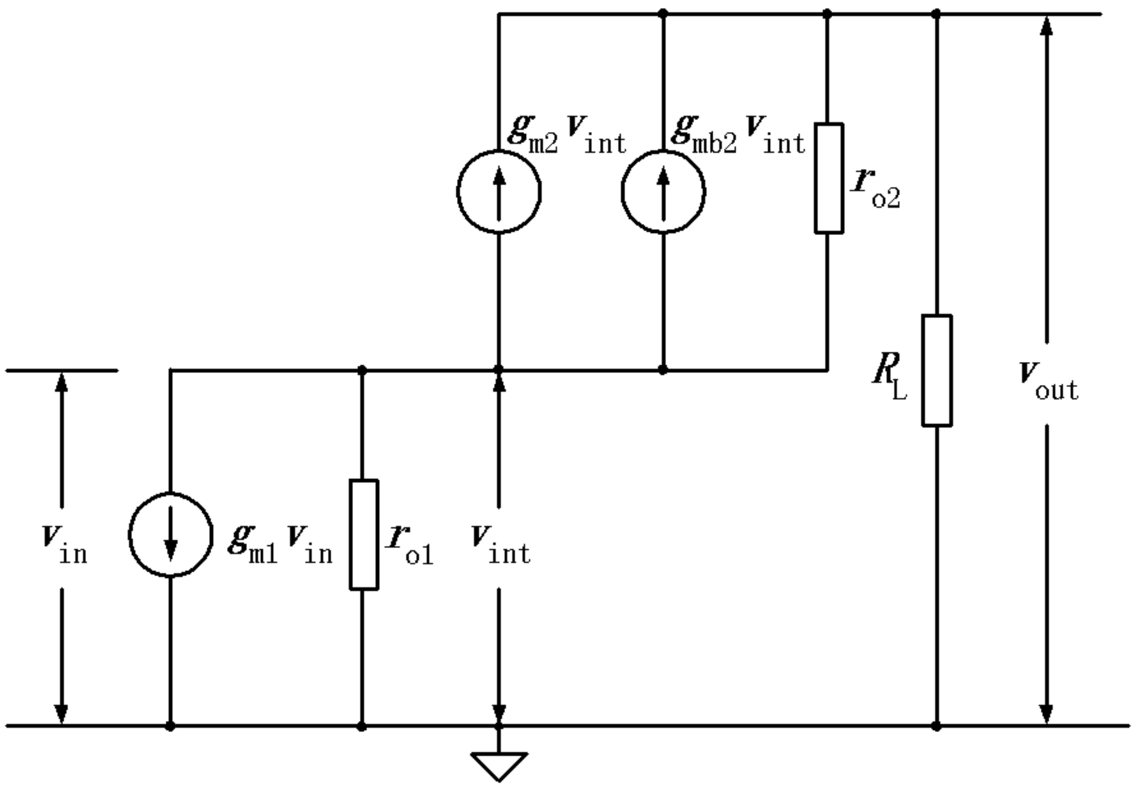
- 没有信号源内阻时
  - 单极点
- 有信号源内阻
  - 两个极点
  - 谁是主极点
  - 相互之间的影响
  - “到地电阻（在某一频点的到地阻抗”的计算

cascode电路结构



- 工作原理的直观解释
- 第一种：
  - 共源：vin-vint
  - 共栅：vint-vout
- 第二种：
  - 共源：vin-iout
  - 共栅：iout(s)-iout(d)
  - 输出：iout×rout

低频交流参数-直流增益



- 求解方法：
  - 1、低频小信号等效电路图
  - 2、增益=等效跨导×输出电阻
  - 3、分段求解
    - 共源+共栅

等效跨导×等效输出电阻

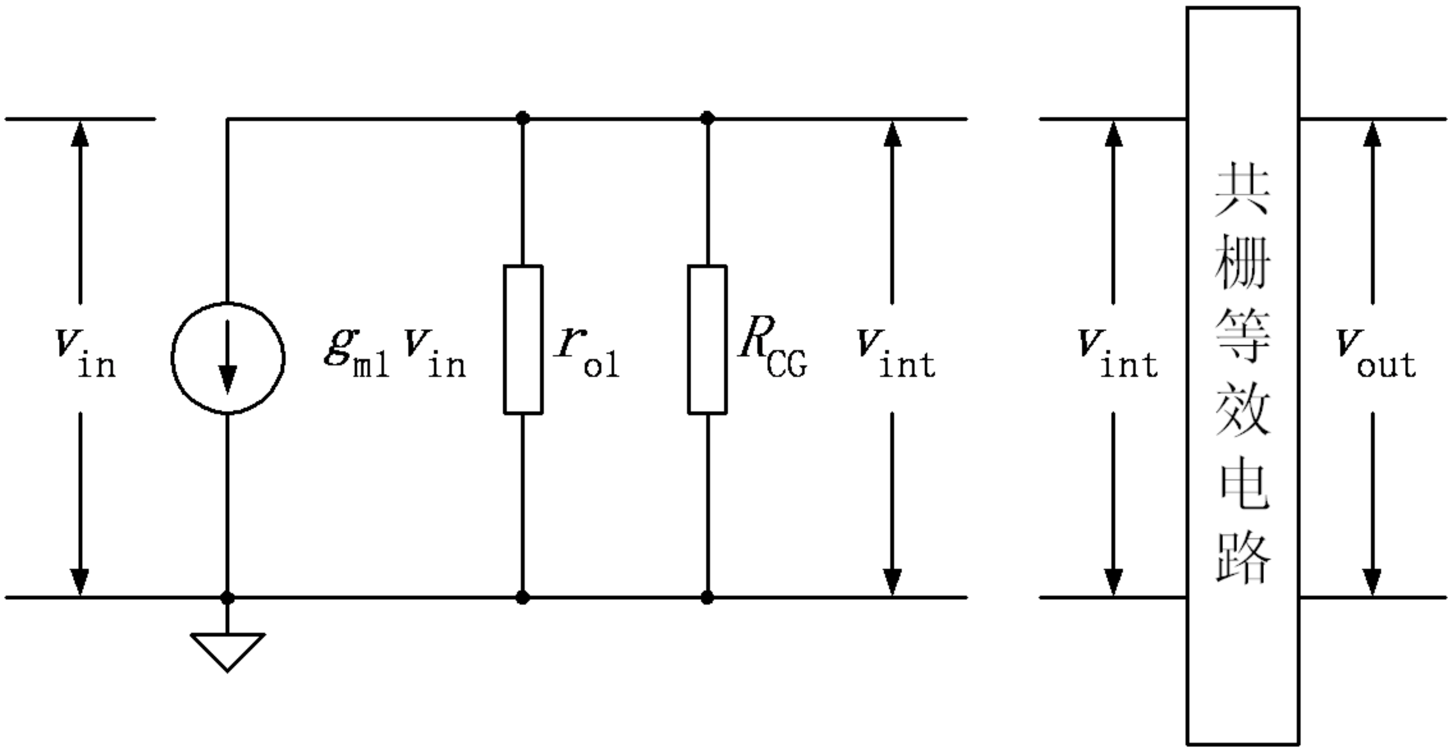
$$G_m = -g_{m1}\frac{\left(g_{m2}+g_{mb2}\right)r_{o1}r_{o2}+r_{o1}}{\left(g_{m2}+g_{mb2}\right)r_{o1}r_{o2}+r_{o1}+r_{o2}} \approx -g_{m1}$$

$$r_{out} = \left(g_{m2}+g_{mb2}\right)r_{o1}r_{o2}+r_{o1}+r_{o2} \approx \left(g_{m2}r_{o2}\right)r_{o1}$$

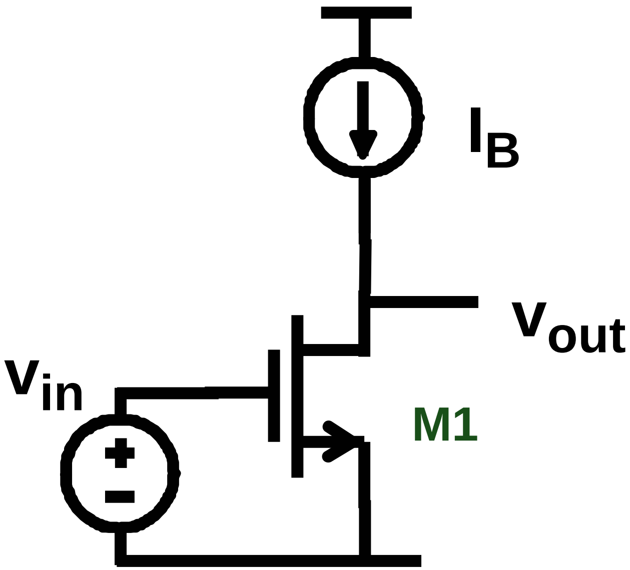
$$A = G_m\left(r_{out}//R_L\right) \approx g_{m1}\left(g_{m2}r_{out2}r_{out1}//R_L\right) = \left(g_{m1}r_{out1}\right)\left(g_{m2}r_{out2}\right)\Big|_{R_L\to\infty}$$

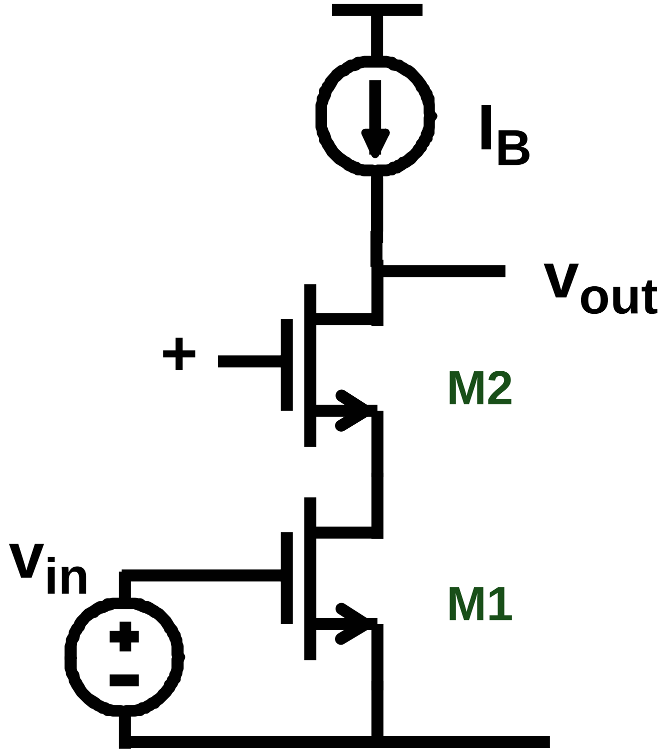
共源+共栅

- 利用共源和共栅放大器的增益表达式



Cascode versus single-transistor





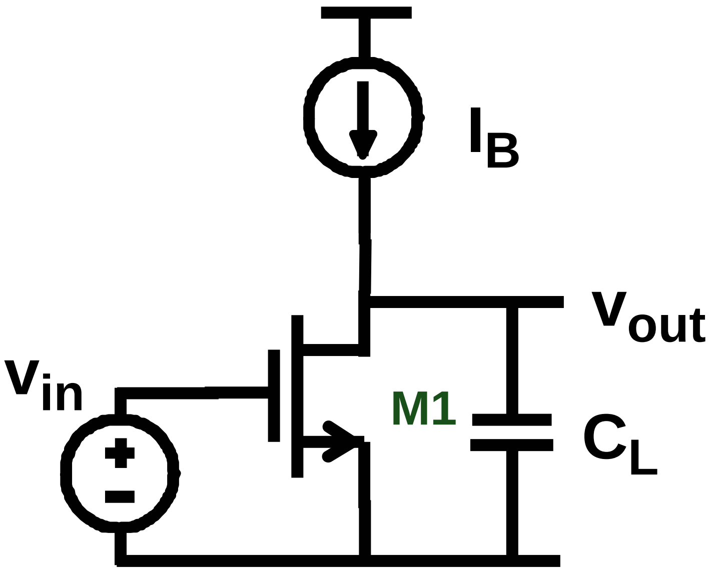
$A_v = (g_m r_{DS})_1$

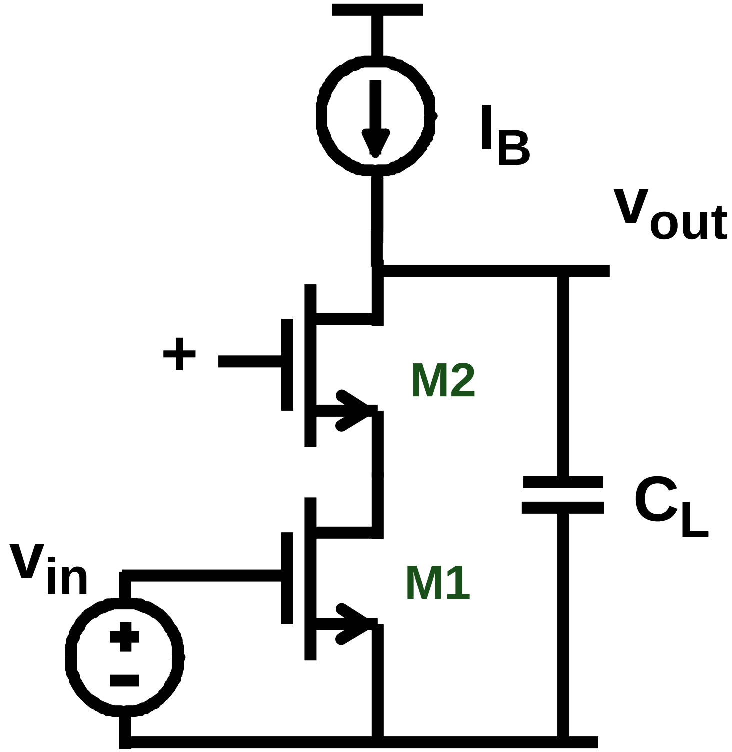
$R_{out} = r_{DS1}$

$A_v = (g_m r_{DS})_1 (g_m r_{DS})_2$

$R_{out} = r_{DS1} (g_m r_{DS})_2$

Cascode versus single-transistor



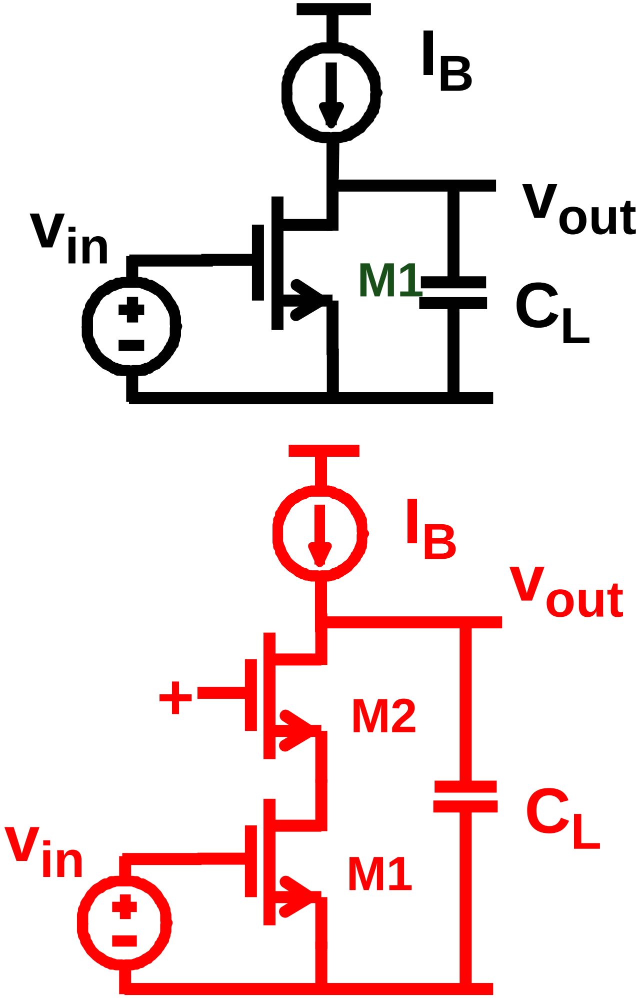


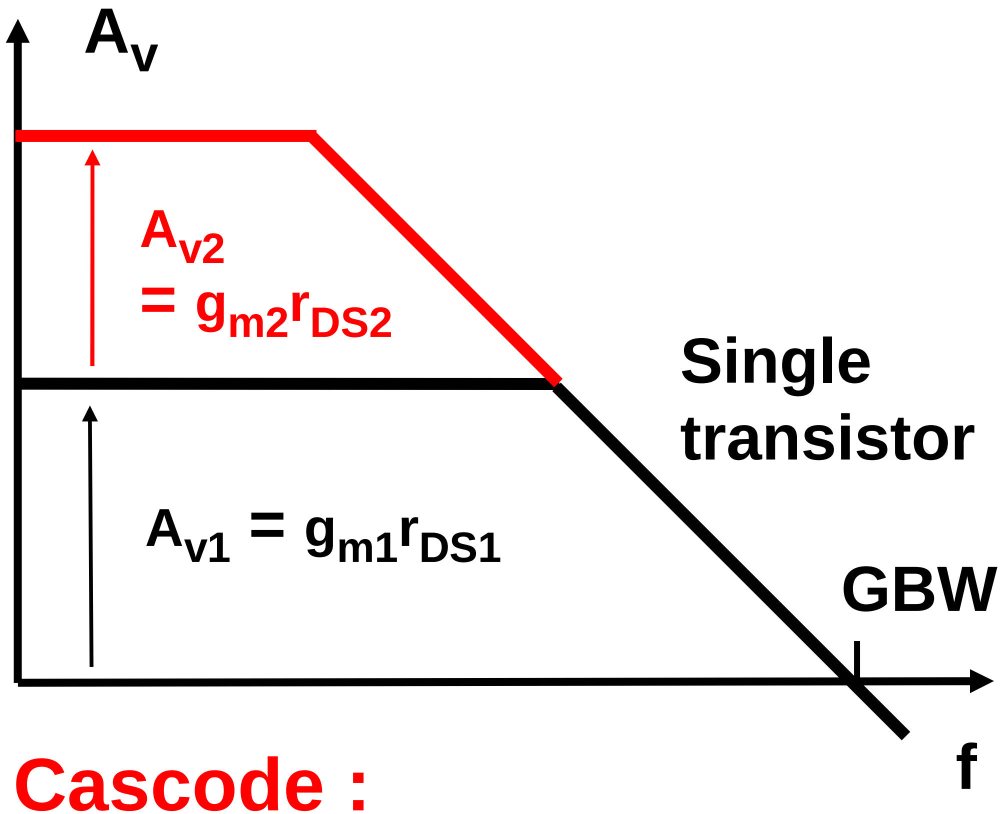
$$BW = \frac{1}{2\pi R_{out} C_L}$$

$$GBW = \frac{g_{m1}}{2\pi C_L}$$

for both !

Cascode versus single-transistor





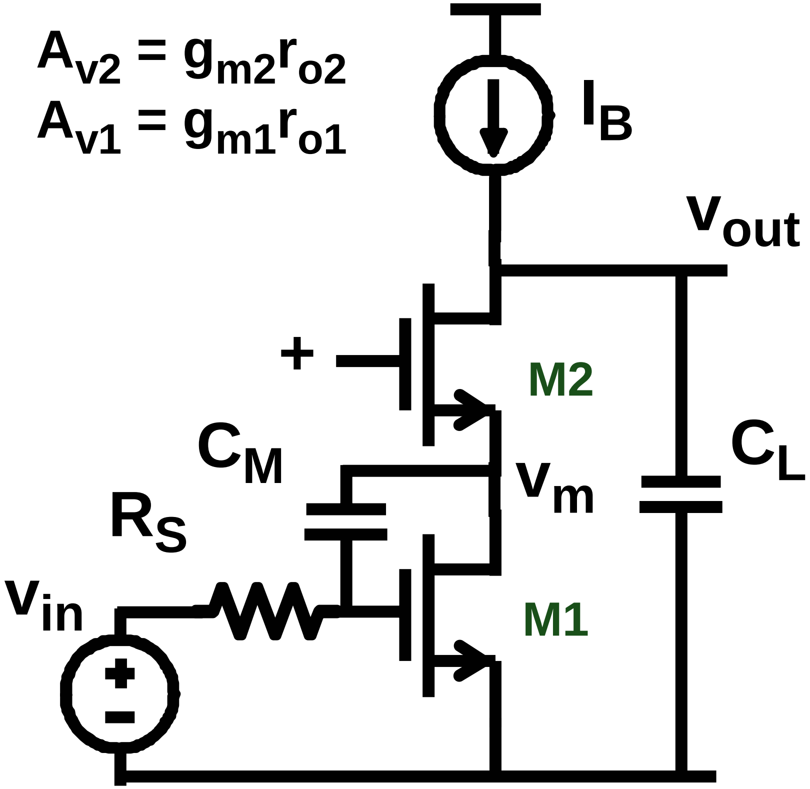
Cascode :
High gain
At low freq.

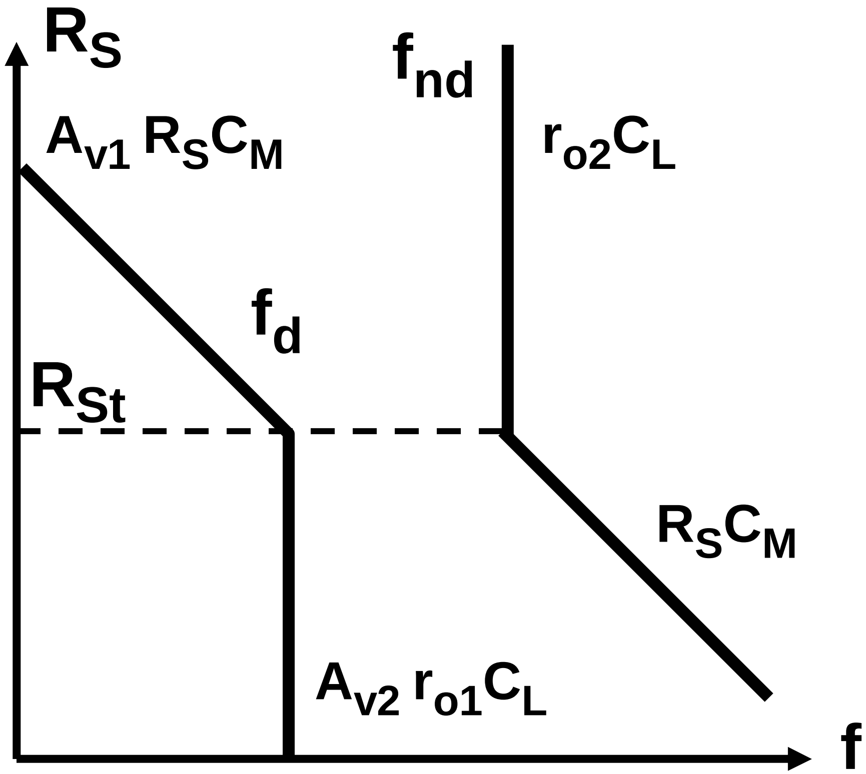
$$GBW = \frac{g_{m1}}{2\pi\, C_L}$$

Miller effect in cascode ?

$A_{v2} = g_{m2} r_{o2}$

$A_{v1} = g_{m1} r_{o1}$





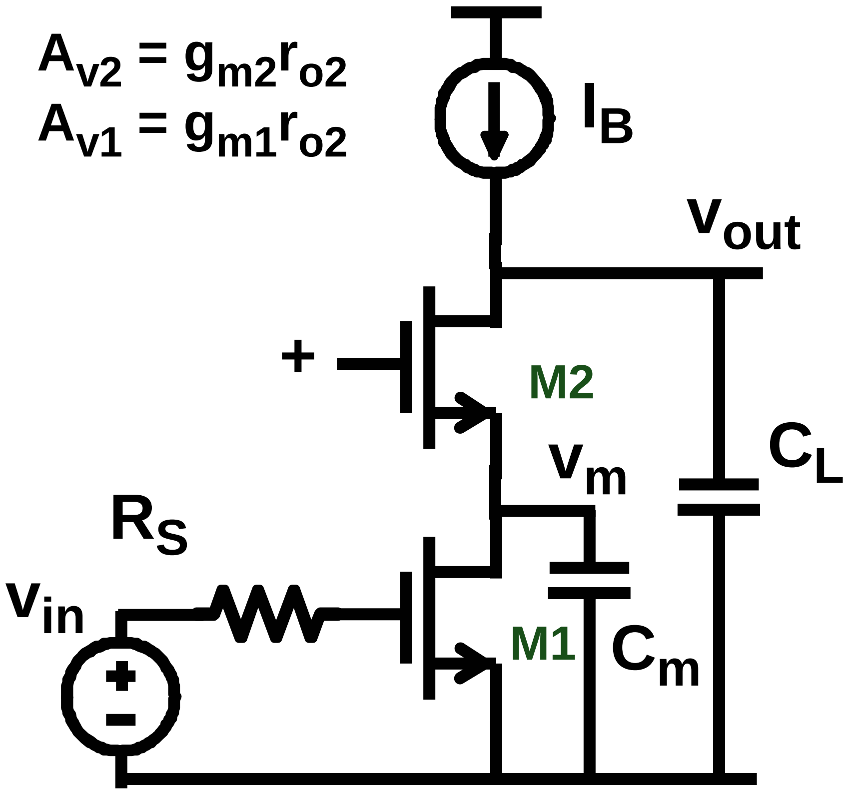
$$GBW = \frac{g_{m1}}{2\pi C_L}$$

No Miller if $R_S < R_{St} = r_{o2} \dfrac{C_L}{C_M} \dfrac{g_{m2}}{g_{m1}}$

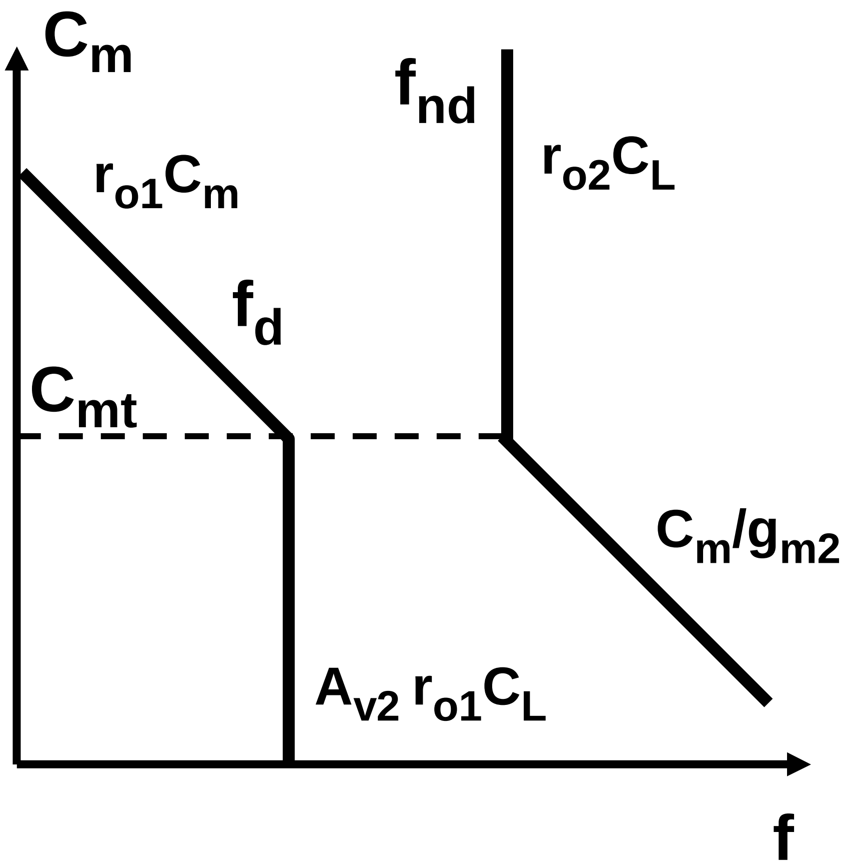
Cascode with capacitance Cm at middle point

$A_{v2} = g_{m2}r_{o2}$

$A_{v1} = g_{m1}r_{o2}$

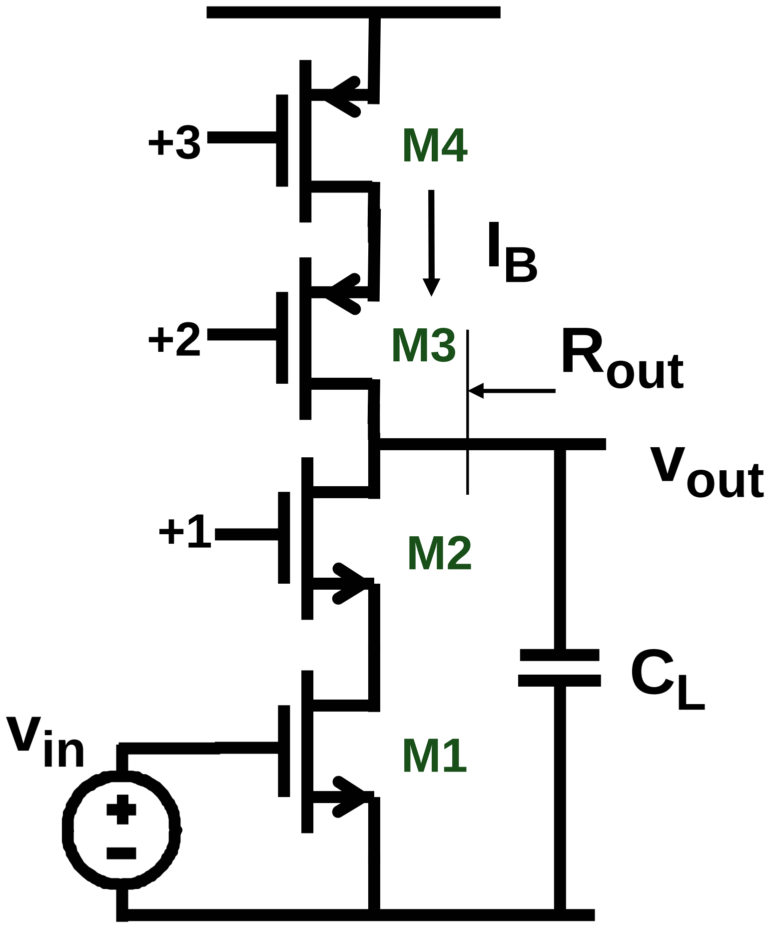


$$GBW = \frac{g_{m1}}{2\pi\, C_L}$$



$$C_{mt} = g_{m2}r_{o2}\,C_L = A_{v2}\,C_L$$

Telescopic Cascode



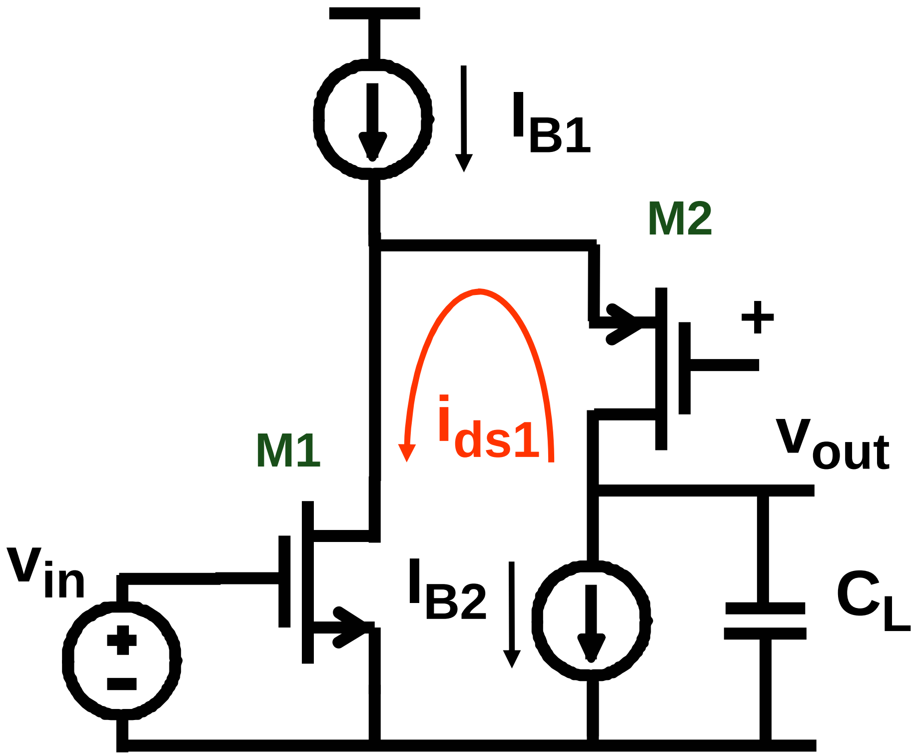
$$A_v = g_{m1} R_{out}$$

$$R_{out} = \frac{1}{2} r_{DS1}\, g_{m2}\, r_{DS2}$$

$$BW = \frac{1}{2\pi R_{out} C_L}$$

$$GBW = \frac{g_{m1}}{2\pi C_L}$$

Folded Cascode



$$I_{DS1} = I_{B1} - I_{B2} \approx I_{B1} / 2$$

$$A_v = g_{m1} R_{out}$$

$$R_{out} = r_{DS1} g_{m2} r_{DS2}$$

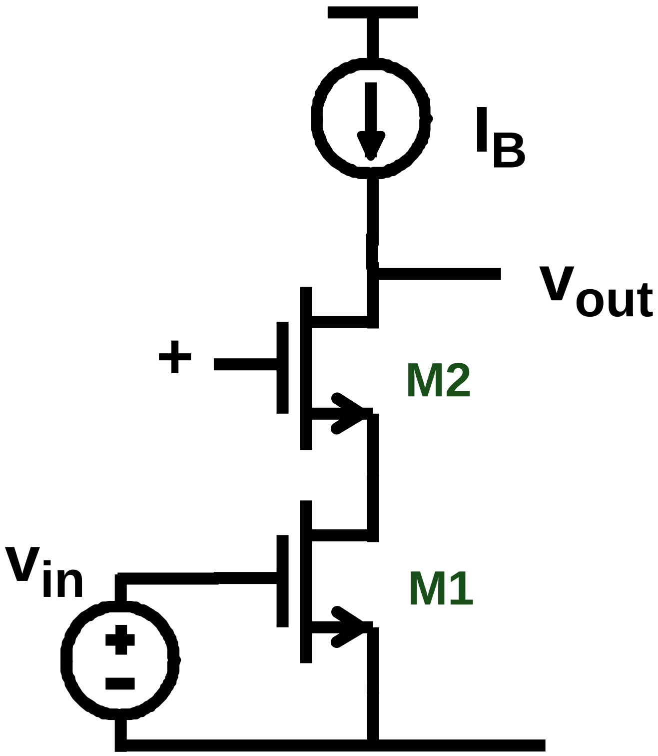
$$BW = \frac{1}{2\pi R_{out} C_L}$$

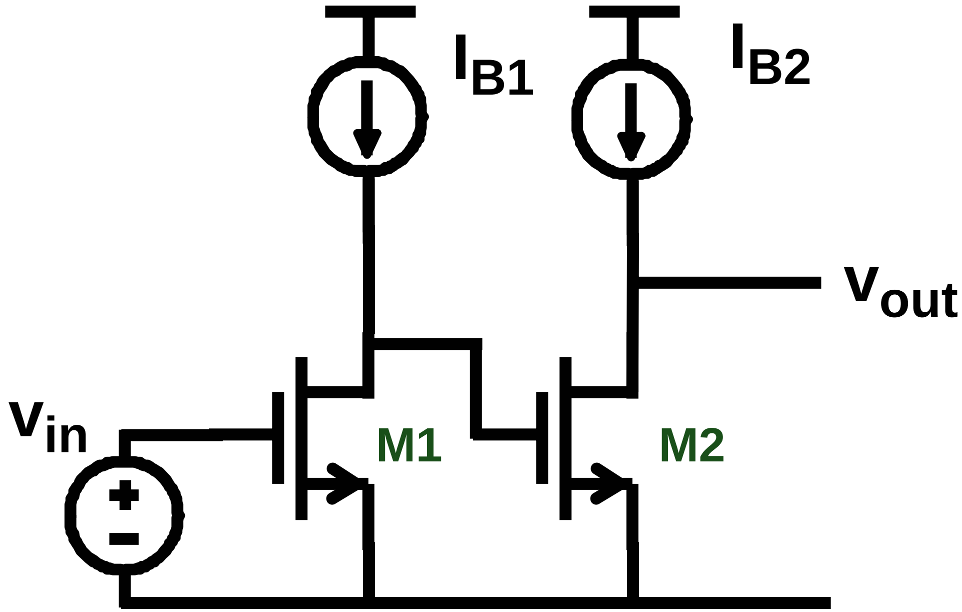
$$GBW = \frac{g_{m1}}{2\pi C_L}$$

套筒式和折叠式cascode的比较

	Telescopic	folded
功耗	低	高
偏置电压个数	4	5
单管输出电压范围	V_{DD}-4$V_{ds,sat}$	V_{DD}-4$V_{ds,sat}$
小信号增益	高	低
差分输出信号范围		
噪声		
构成反馈时输出范围		

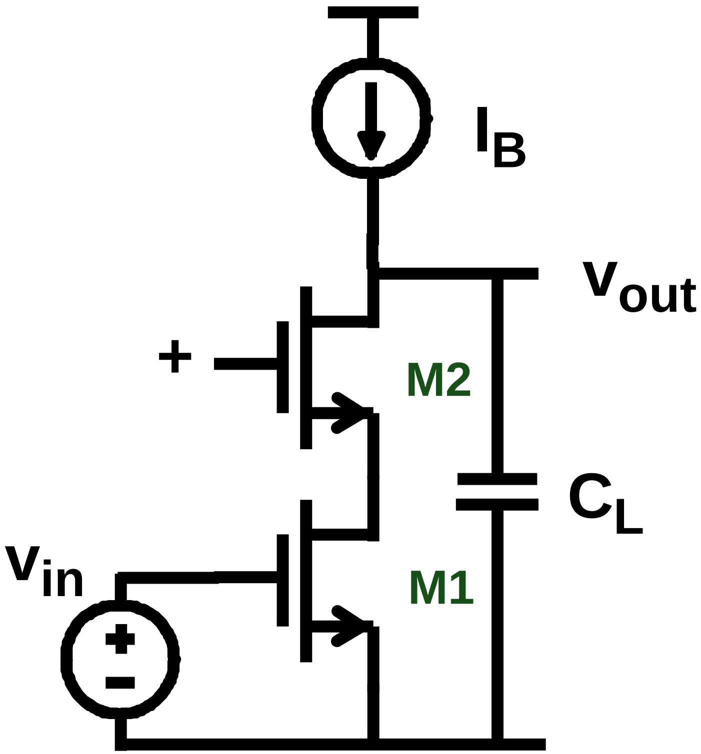
Cascode versus cascade

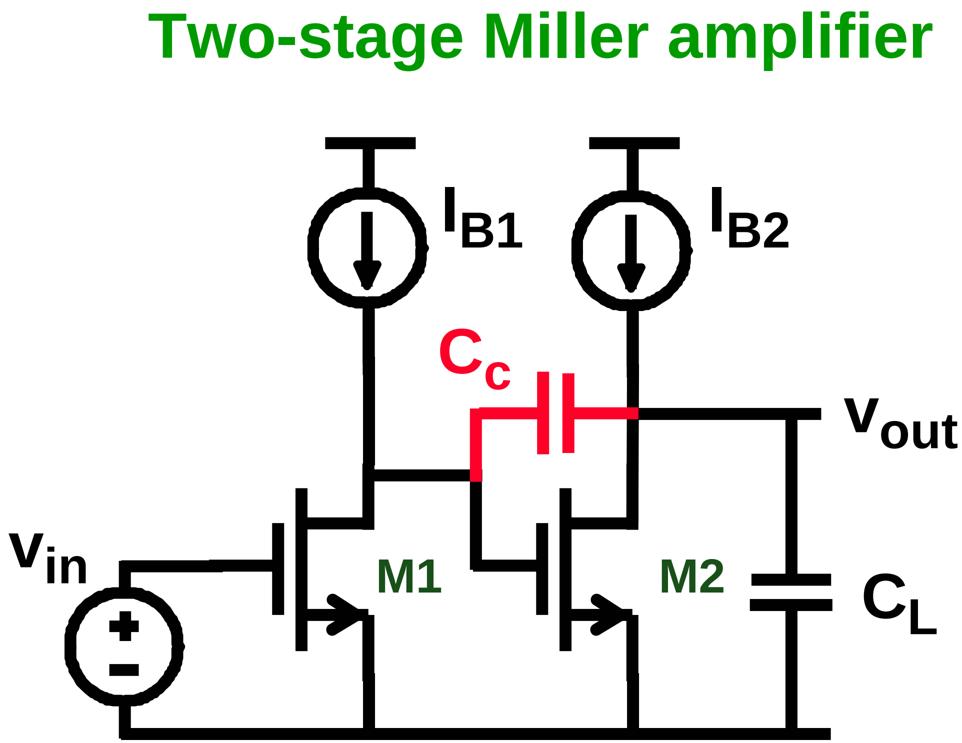




$$A_v = (g_m r_{DS})_1 (g_m r_{DS})_2 \qquad A_v = (g_m r_{DS})_1 (g_m r_{DS})_2$$

Cascode versus cascade

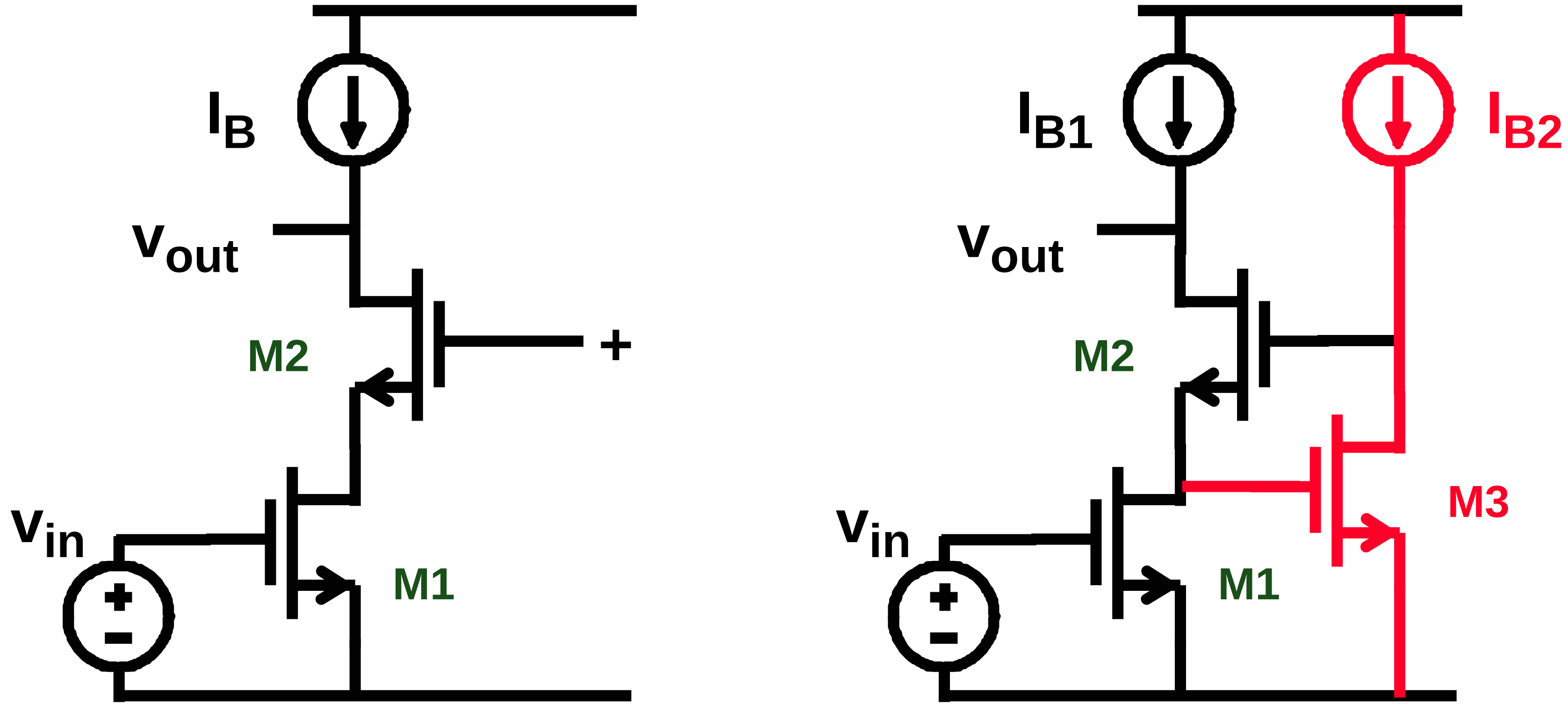




$$GBW = \frac{g_{m1}}{2\pi C_L}$$

$$GBW = \frac{g_{m1}}{2\pi C_c}$$

Regulated cascode or gain boosting

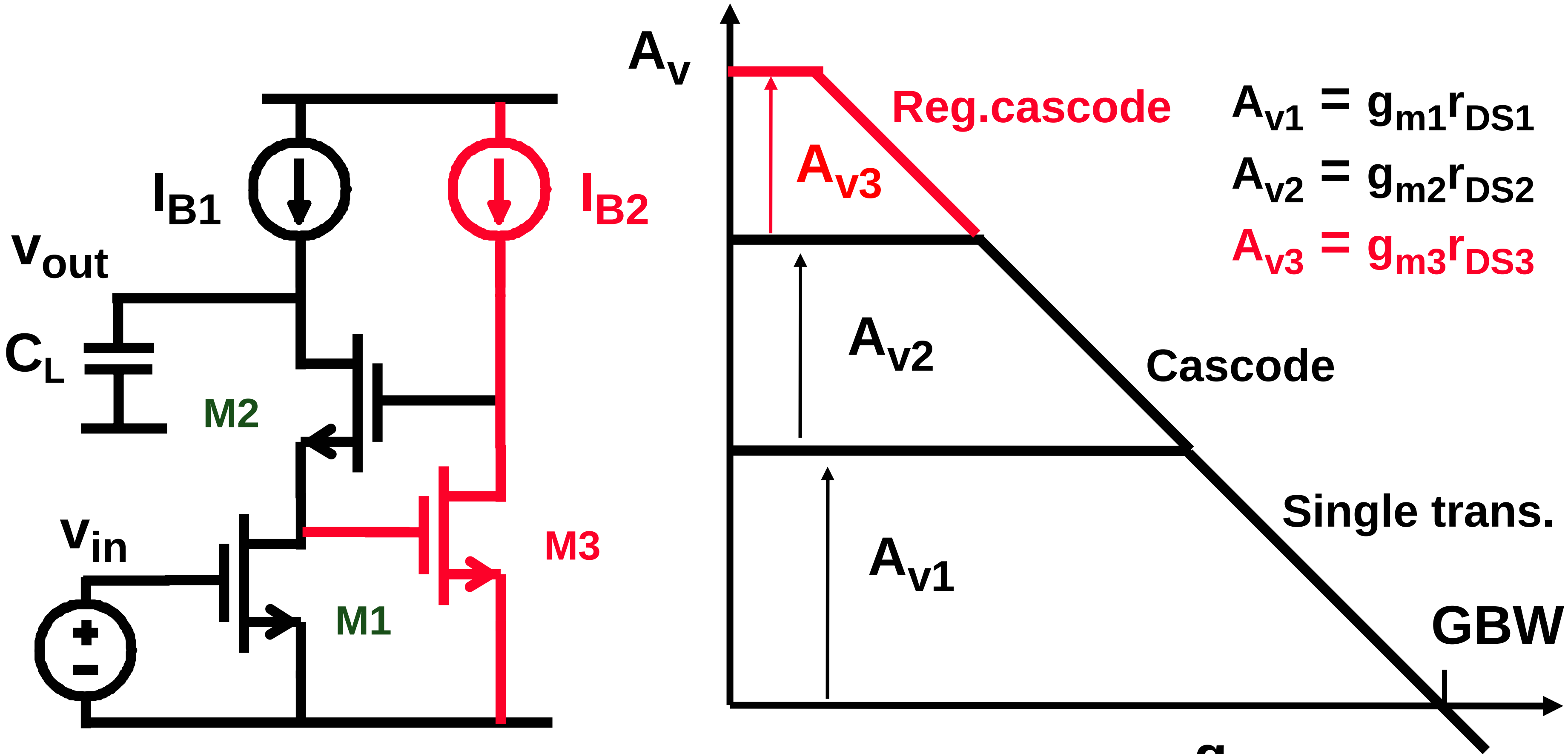


$$A_v = (g_m r_{DS})_1 (g_m r_{DS})_2$$

$$\boxed{A_v = (g_m r_{DS})_1 (g_m r_{DS})_2 (g_m r_{DS})_3}$$

Hosticka, JSSC Dec.79, pp. 1111-1114; Sackinger, JSSC Febr.90, pp. 289-298;
Bult JSSC Dec.90, pp. 1379-1384

Regulated cascode, Cascode & single-transistor



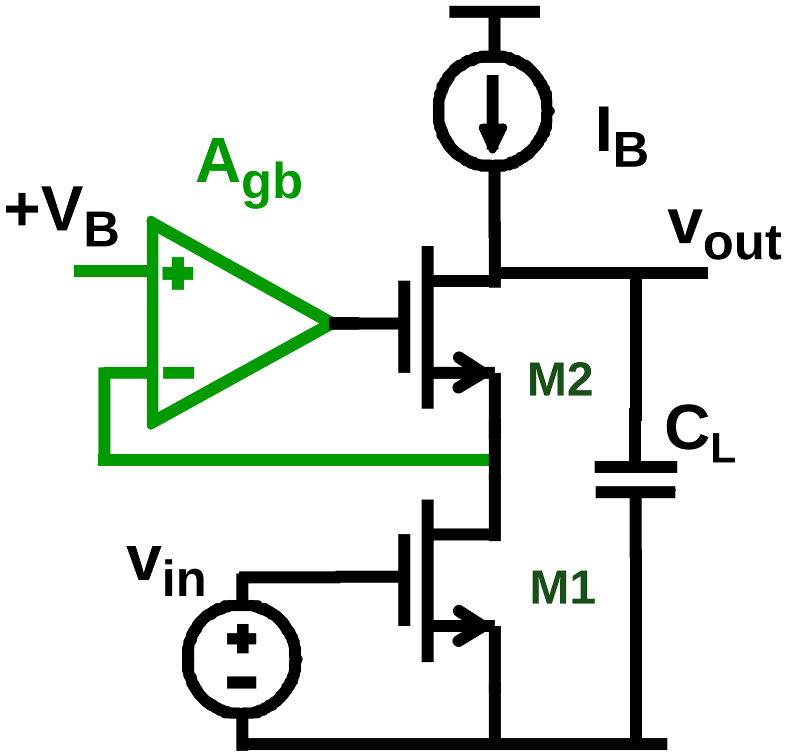
$A_{v1} = g_{m1}r_{DS1}$

$A_{v2} = g_{m2}r_{DS2}$

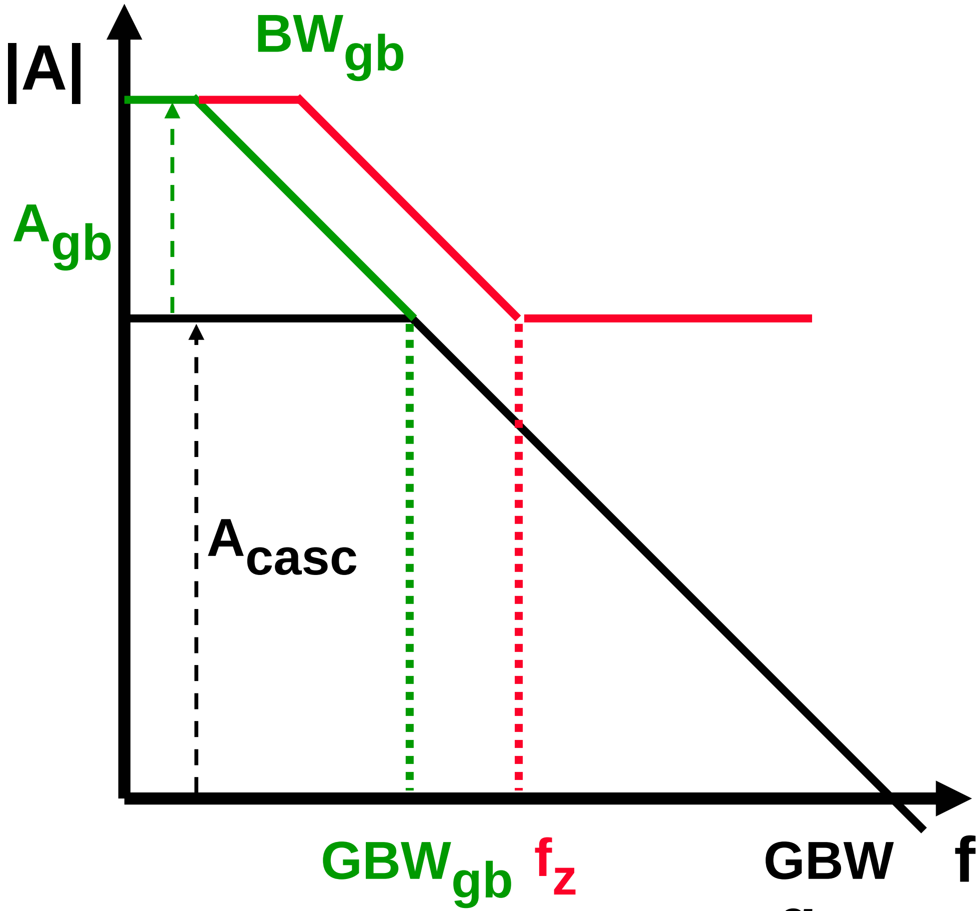
$A_{v3} = g_{m3}r_{DS3}$

$$GBW = \frac{g_{m1}}{2\pi\, C_L}$$

Gain boosting

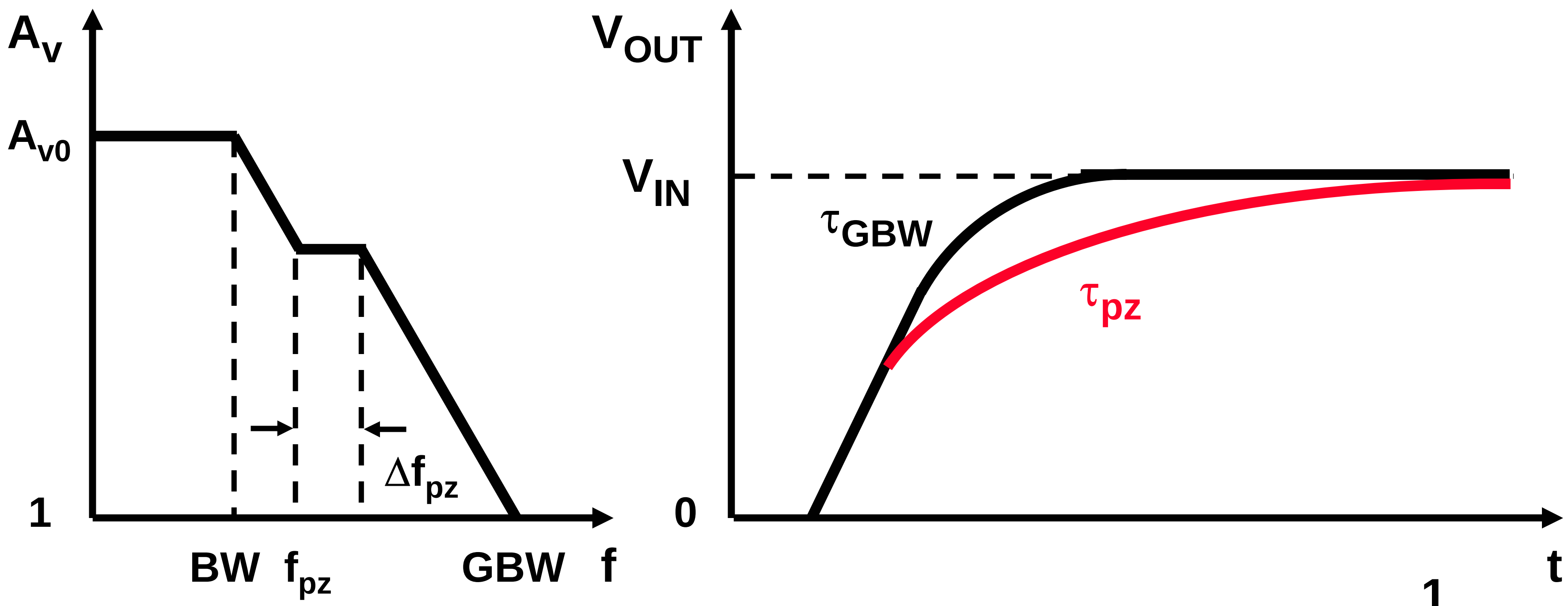


$$A_v = A_{gb}(g_m r_{DS})_1 (g_m r_{DS})_2$$



$$= \frac{g_{m1}}{2\pi\, C_L}$$

Pole-zero doublet and settling time



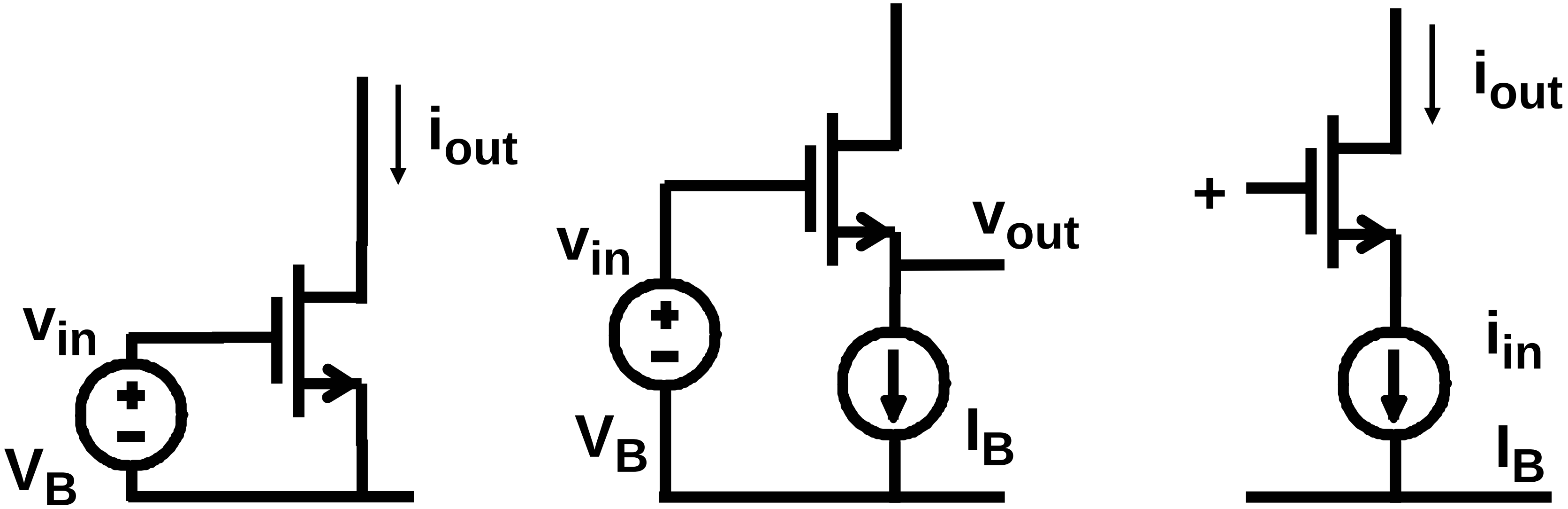
$$f_{pz} = \frac{1}{2\pi\, \tau_{pz}}$$

$$V_{OUT} = V_{IN} \left[1 - \exp\left(- \frac{t}{\tau_{GBW}}\right) - \frac{\Delta f_{pz}}{GBW} \exp\left(- \frac{t}{\tau_{pz}}\right)\right]$$

$$GBW = \frac{1}{2\pi\, \tau_{GBW}}$$

Kamath, etal, JSSC Dec.74, pp. 347-352

Single-transistor stages



$i_{out} = g_m v_{in}$

$v_{out} = v_{in}$

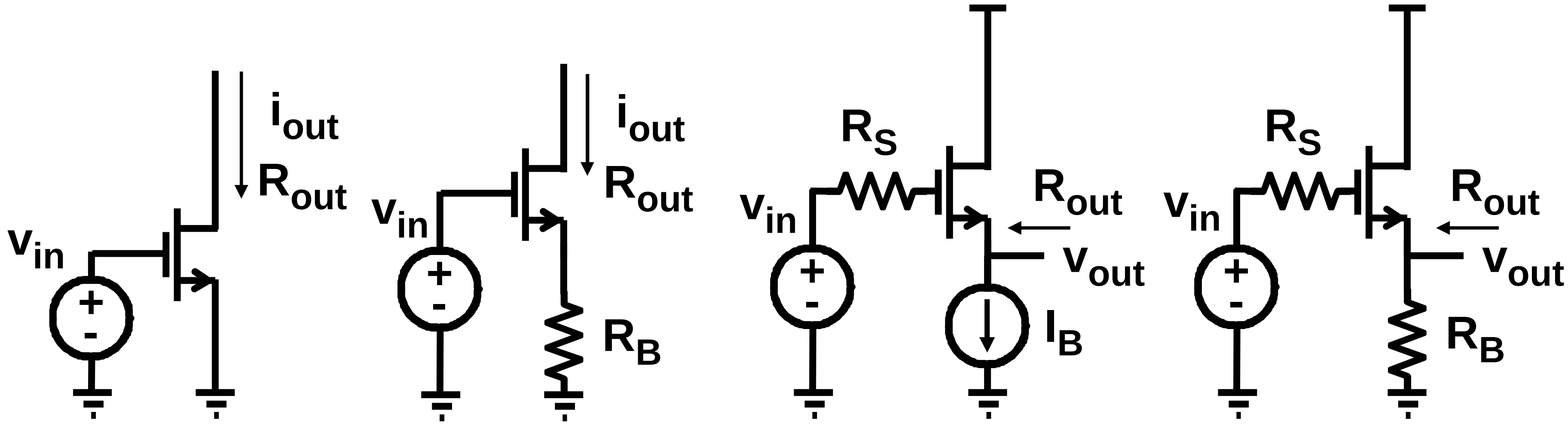
$Z_{out} \approx 1/g_m$

$i_{out} = i_{in}$

$Z_{in} \approx 1/g_m$

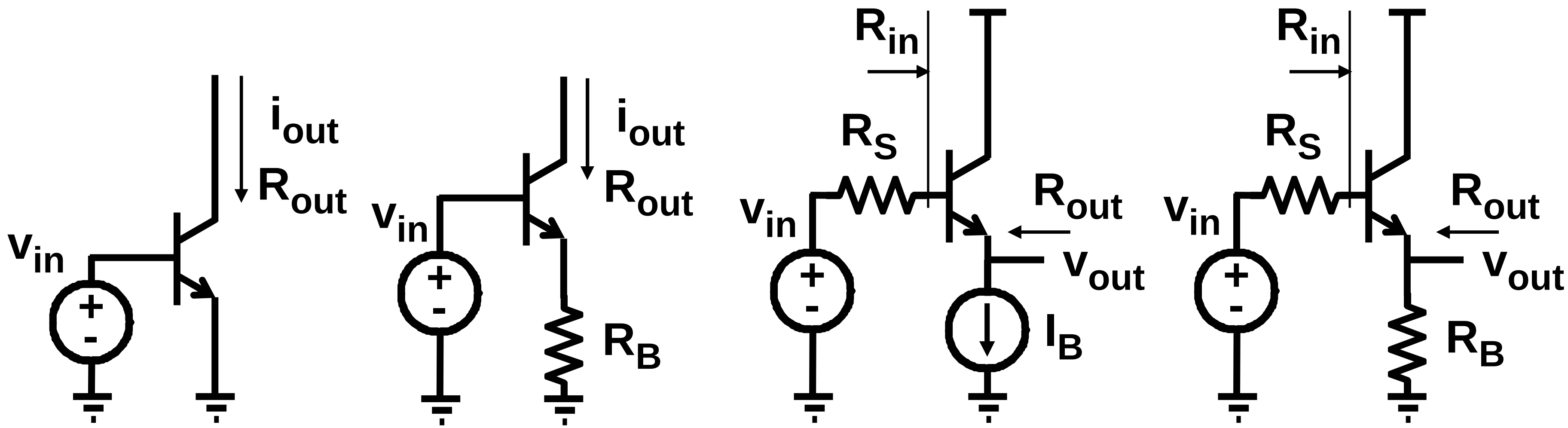
Amplifier **Source follower** **Cascode**

MOST amplifier & follower



		$R_B > 1/g_m$			$R_B > 1/g_m$
A_G	g_m	$1/R_B$	A_V	1	1
R_{in}	∞	∞		∞	∞
R_{out}	r_o	$g_m R_B r_o$		$1/g_m$	$1/g_m$

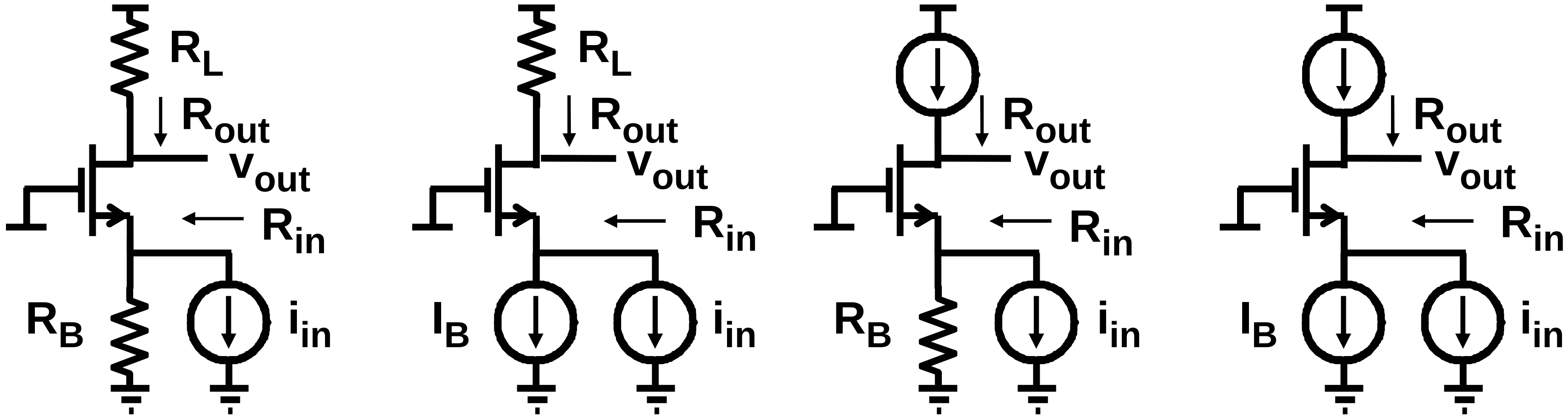
Bipolar transistor ($\beta >> 1$)



$R_B > 1/g_m$ $\qquad\qquad\qquad\qquad$ $R_B > 1/g_m$

A_G	g_m	$1/R_B$	A_V	1	1
R_{in}	r_B+r_π	$r_B+r_\pi+\beta R_B$		$r_B+r_\pi+\beta r_o$	$r_B+r_\pi+\beta R_B$
R_{out}	r_o	$g_m R_B r_o$		$1/g_m + R_S/\beta$	$1/g_m + R_S/\beta$

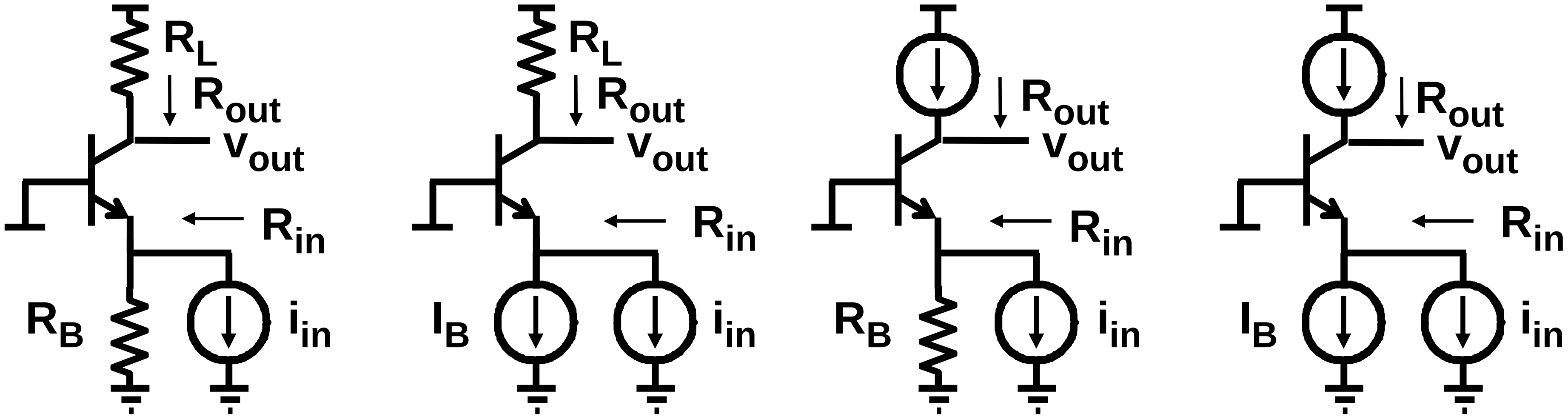
In- & output resistances MOST cascode



$R_B > 1/g_m$ $R_B > 1/g_m$

A_R	R_L	R_L	$g_m r_o R_B$	-
R_{in}	$1/g_m$	$1/g_m$	R_B	∞
R_{out}	$g_m r_o R_B$	∞	$g_m r_o R_B$	∞

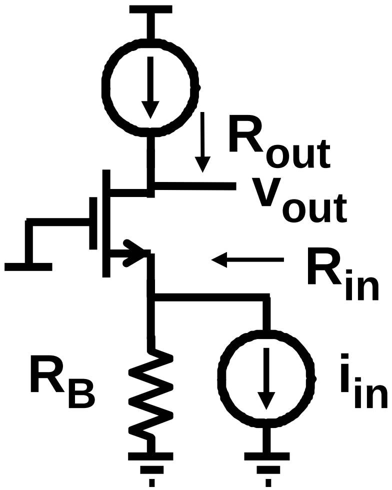
In- & output resistances Bipolar trans. cascode



$R_B > 1/g_m$ $R_B > 1/g_m$

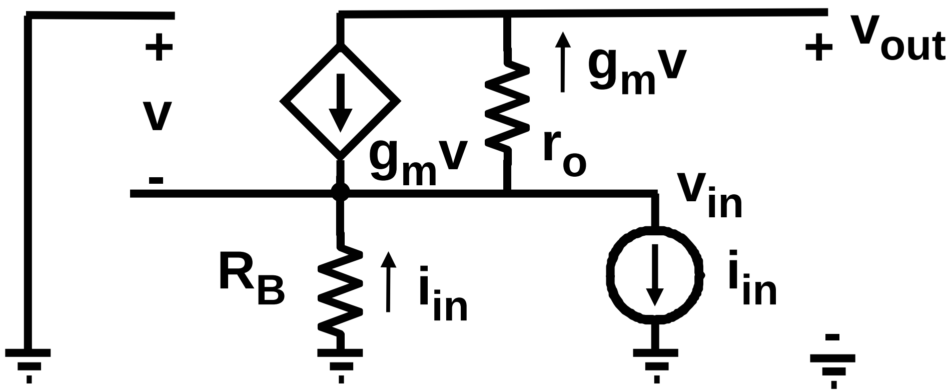
A_R	R_L	R_L	$g_m r_o R_B$	-
R_{in}	$1/g_m$	$1/g_m$	$R_B//(r_B+r_\pi)$	r_B+r_π
R_{out}	$g_m r_o R_B$	$\approx \beta r_o$	$g_m r_o(R_B //(r_B+r_\pi))$	$\approx \beta r_o$

Calculation of A_R for a MOST cascode



$R_B > 1/g_m$

$$A_R = g_m r_o R_B$$



$$\left\{ \begin{array}{l} v = -\,v_{in} \\ v_{out} = v_{in} - g_m v\, r_o \\ v_{in} = -\,R_B\, v_{in} \end{array} \right.$$

yields $\quad v_{out} = -\,R_B\, i_{in}\,(1 + g_m r_o\,)$

and $\quad g_m r_o >> 1$

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