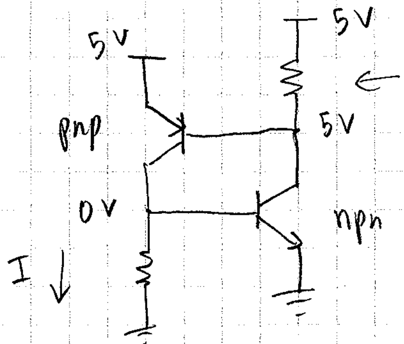
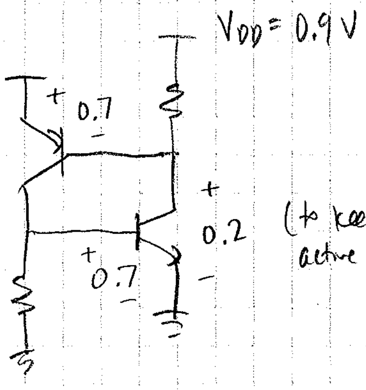
1

- Consider the circuit formed by the parasitic bipolars:



← no current flow, both transistors are off

- Consider the case when for some reason a current flows in the substrate (I)
- Voltage at base of npn rises, turning it on.
- This causes current to flow in other resistor, dropping voltage at base of pnp, turning it on.
- positive feedback, until:



(to keep in active region)

→ Power supply is shorted out to 0.9 V
- chip catches on fire
- not good.

Moral of the story: - Need to keep gain around the feedback loop low
- remember the feedback with gain < 1 will die out.
- do this by keeping gain of two common emitter stages low
  - make parasitic resistors (R_p & R_n) small

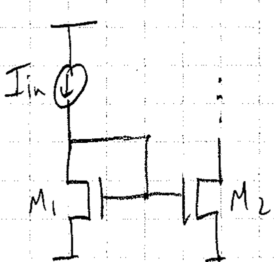
- How to do this?
  - Resistance is a function of two things: doping and size
    - have p+ layer under p− substrate (controlled by process guys)
    - reduce distance with lots of substrate contacts
    - modern processes have design rules with minimum substrate contact distances.

5720/6720: Lecture #6 Jan. 24

Ch.3: Circuit Design!

- Start with current mirrors: Read Ch.3 sections 3.1, 3.5, 3.6

Basic Current Mirror [3.1]



- Consider M_1
  - as I_{in} flows, gate of M_1 charges up until it turns on
  - M_1 will be in saturation, since $V_{GS} = V_{DS}$ ∴ $V_{DS} > V_{GS} - V_T$

→ M_1 configuration (gate shorted to drain) is called diode connected since I-V characteristics resemble those of a diode.

→ For M_1 (neglecting channel length modulation):

$$I_{D_1} = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)_1 (V_{GS_1} - V_T)^2$$

2

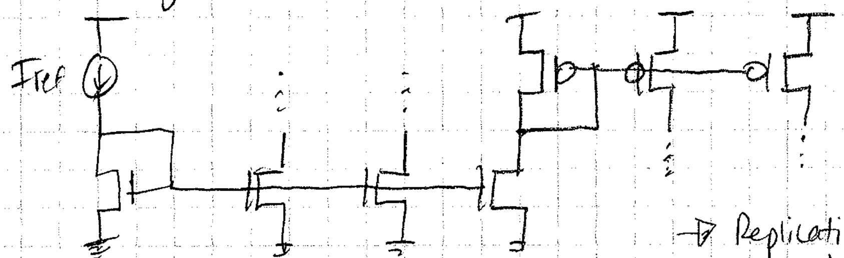
- Can solve this to find V_{GS_1}

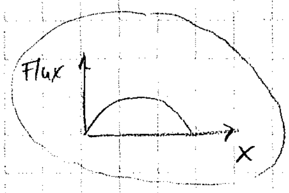
→ Now, $V_{GS_1} = V_{GS_2}$, so if we assume V_{DS_2} is high enough to put it in saturation and ignore C.L.M., we have:

$$I_{D_2} = I_{D_1} \cdot \frac{(W/L)_2}{(W/L)_1}$$

- current in both devices is same, scaled by their (W/L) ratios

→ Using this circuit, a bias current can be replicated and scaled across many devices:



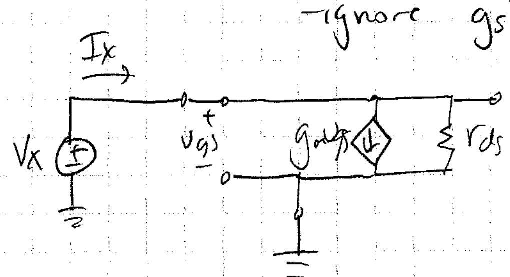


→ Replicating devices vs. scaling widths
- better current matching
- edge effects.

→ layout ...

→ What about small signal behavior?

- First consider diode connected transistor:
  - usually we analyze current mirrors for low frequency signals
    - ignore parasitic capacitors (open circuit)
  - sources are at ground (no body effect)
    - ignore g_s



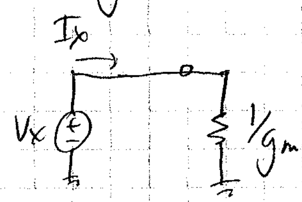
- Connect test source V_x, find I_x.

$$I_x = g_m v_x + \frac{v_x}{r_{ds}}$$

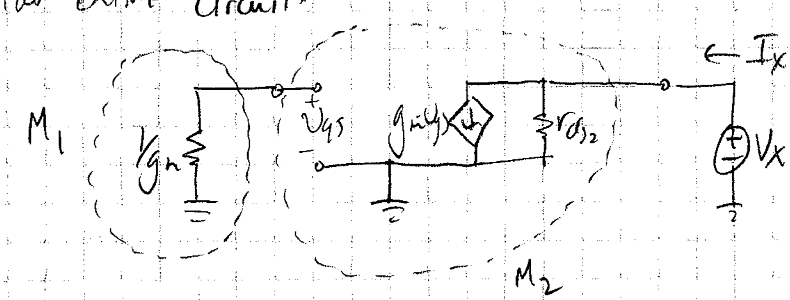
$$\therefore R_{in} = \frac{V_x}{I_x} = \left(g_m + \frac{1}{r_{ds}}\right)^{-1} = \frac{1}{g_m} \parallel r_{ds}$$

→ Usually $r_{ds} \gg \frac{1}{g_m}$, $\therefore R_{in} \approx \frac{1}{g_m}$

So, equivalent S.S. model is:



Now consider entire circuit:



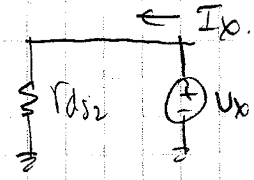
- Open circuited current source.

- We used the simplified model for M_1 (valid since it is diode connected)

- Let's find output impedance.

3

- $v_{gs_2} = 0$, ∴ gm current source will be off:

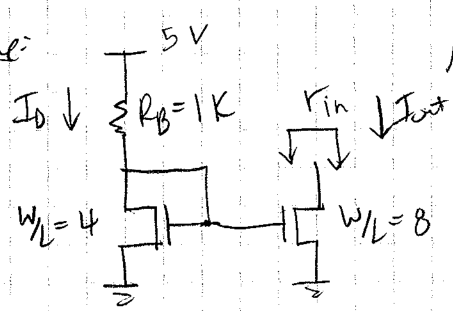


∴ $r_{in} = \frac{V_x}{I_x} = r_{ds_2}$ (easy!)

→ For the assumption neglecting C.L.M., $r_{ds} = \infty$, but in general r_{ds} will have an effect and we need to consider it.

In general, $r_{ds} \propto \frac{L}{I_D}$.

Example:



$\mu_n C_{ox} = 100 \frac{\mu A}{V^2}$, $V_T = 1V$, $r_{ds} = \frac{8000 \cdot L}{I_D}$ ← μm, ← mA, $L = 1\ \mu m$

Find r_{in}.

- Need to solve large signal equations first
  - DON'T use $1/g_m$ for any large signal calculations!!!

$V_{GS} = 5 - I_D \cdot R_B$.

$I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_T)^2 \rightarrow I_D = \frac{100E\text{-}6}{2} \cdot 4\,(5 - I_D \cdot 1000 - 1)^2$

→ quadratic equation with I_D: $I_D^2 \cdot 1E6 - I_D \cdot 13E3 + 16 = 0$

solve using $x = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$

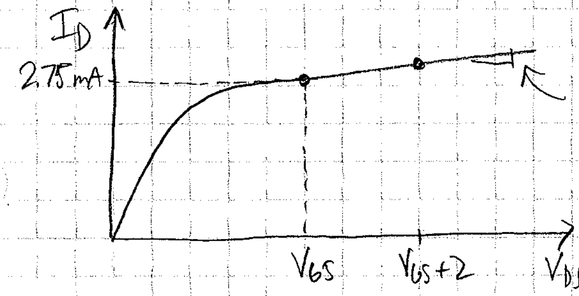
→ $\underline{I_D = 1.376\ mA}$ or $I_D = 11.6\ mA$ ← leads to −ve V_{GS}

∴ $I_{out} = 2 \cdot I_D = 2.752\ mA$

∴ $r_{in} = r_{ds} = \frac{8000 \cdot 1}{2.752} = 2.907\ k\Omega$ (small signal resistance)

assuming it stays in saturation.

Example: In prior example, assume Vout increases by 2V above V_{GS}, how much error is introduced into the output current?



$\frac{\Delta I_D}{\Delta V_{DS}} = \frac{1}{r_{ds}}$

$\Delta V_{DS} = 2\,V$

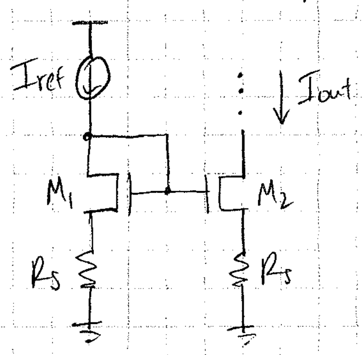
$\therefore \Delta I_D = \frac{\Delta V_{DS}}{r_{ds}} = \frac{2}{2.907k} = 0.69\ mA$

→ This corresponds to a $\frac{0.69}{2.75} \times 100\% = 25\%$ error (significant!!)

- The finite output resistance of the standard current mirror is their biggest weakness, as it can lead to large errors in the output current.
- to address this, we need some way of increasing the output resistance of the current mirror.

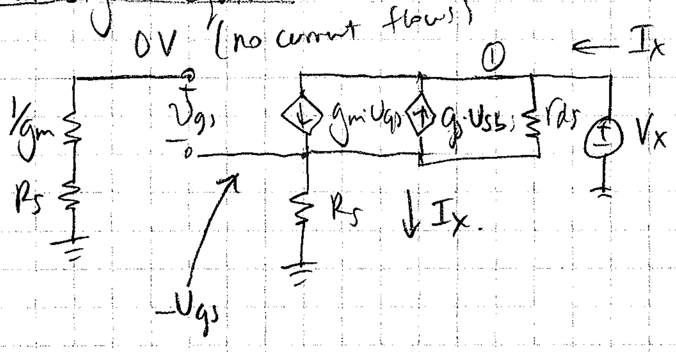
Source Regenerated Current Mirrors [3.5]

- what if we put a resistor in the source of the transistor?



- why include R_s in both legs? (Matching)
- Let's analyze small signal circuit to find output impedance.

Small signal equivalent:



⊛ → $v_{gs} = -I_x \cdot R_s$

→ KCL @ ①: $I_x = g_m v_{gs} + \frac{[V_x - (-v_{gs})]}{r_{ds}} - g_s \cdot (-v_{gs})$

Sub in ⊛: $I_x = -I_x \cdot R_s (g_m + g_s) + \frac{V_x - I_x R_s}{r_{ds_2}}$

$$\rightarrow R_{out} = \frac{V_x}{I_x} = r_{ds_2}\left[1 + R_s\left(g_m + g_s + \frac{1}{r_{ds_2}}\right)\right]$$

$$\approx r_{ds_2}\left[1 + R_s(g_m + g_s)\right]$$

(since $\frac{1}{r_{ds}}$ is small compared to g_m).

→ g_s is usually about $1/5$ of g_m, so sometimes it is neglected to yield the simplified result:

$$R_{out} \approx r_{ds_2}(1 + g_m \cdot R_s)$$

Observations: - The O/p resistance has been increased over the nominal case by a factor of $(1 + R_S(g_m + g_s))$.

- This is a general result that is true when analyzing a source degenerated transistor

- Here the body effect works to our advantage, further increasing the output resistance.

Drawbacks: - Need to keep M_2 in saturation, R_S increases the minimum V_D for which this condition holds.

Example: Revisit prior example, but assume there was a source degeneration resistance of $400\,\Omega$

(a) New R_{out}?
(b) % error for 2 V increase in V_{DS}?
(c) $V_{out,min}$ for saturation?

(a) Need to find g_m (let's ignore g_s for simplicity)

$g_m = \frac{2 I_D}{V_{eff}}$ $\quad$ V_{eff}: $V_{GS} = 5V - I_b \cdot R_B$
$= 5 - 1.376\,mA \cdot 1k$
$= 3.62\,V$

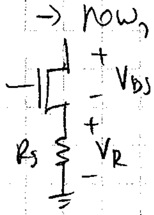
$g_m = \frac{2(2.75\,mA)}{2.62}$ $\quad$ $\therefore V_{eff} = V_{GS} - V_T = 2.62\,V.$
$= 2.1\,mS$

Now, $R_{out} \approx r_{ds2}(1 + g_m R_S) = 2.91\,k\Omega\,(1 + 2.1E{-3} \cdot 400) = 5.35\,k\Omega$

(b) $\Delta I_D = \frac{\Delta V_{DS}}{R_{out}} = \frac{2}{5.35\,k} = 0.37\,mA \Rightarrow \frac{0.37}{2.75} \times 100\% = 13\%$ error (better!)

(c) In prior case, we needed to maintain $V_{DS} > V_{GS} - V_T = V_{eff} = 2.62\,V$

→ now, we add voltage across resistor to this



$V_R = I_{out} \cdot R_S = 2.75\,mA \cdot 400 = 1.1\,V.$

$\Rightarrow V_{out,min} = V_R + V_{DS,min} = 1.1 + 2.62 = 3.72\,V$

(reduction in headroom, tradeoff)

This gives some improvement but maybe we can do better.
- also resistors tend to be bulky and poorly controlled, we would like a solution using transistors only!