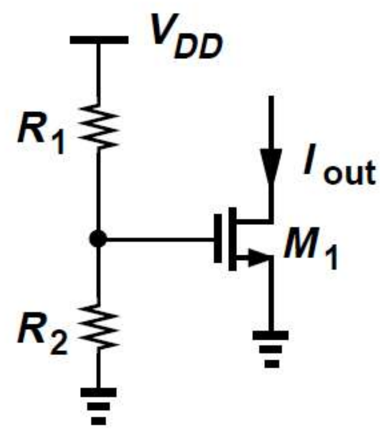
Basic Current Mirrors

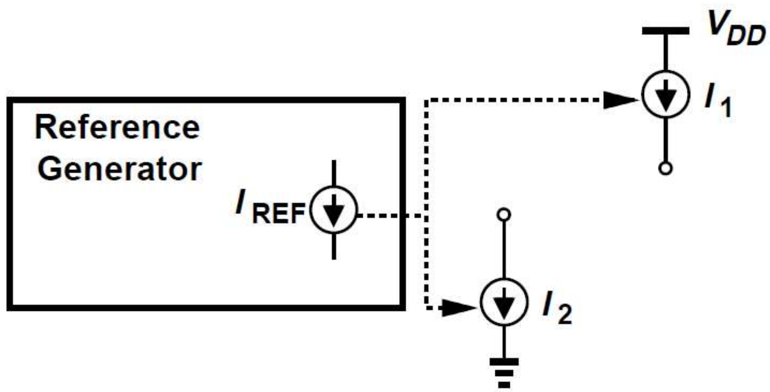


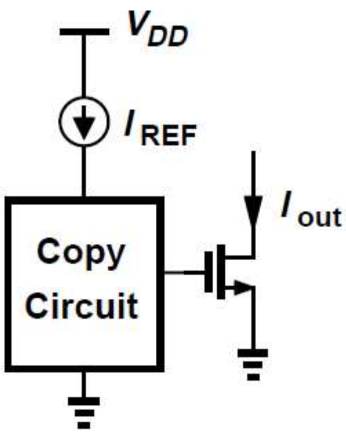
- AssumingM1 is in saturation, we can write

$$I_{out} \approx \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(\frac{R_2}{R_1+R_2}V_{DD}-V_{TH}\right)^2.$$

- The threshold voltage may vary by 50 to 100 mV from wafer to wafer
- Both μn and V_{TH} exhibit temperature dependence
- We must seek other methods of biasingMOS current sources.

Conceptual means of copying currents



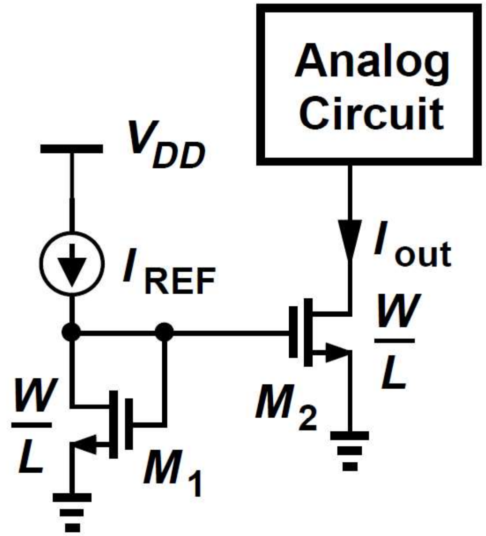


- **Use of a reference to generate various currents.**

$$I_{out} = f[f^{-1}(I_{REF})] = I_{REF}$$

- **Two identical MOS devices that have equal gate-source voltages and operate in saturation carry equal currents**

Effect of Channel-Length Modulation



- Neglecting channel-length modulation, we can write

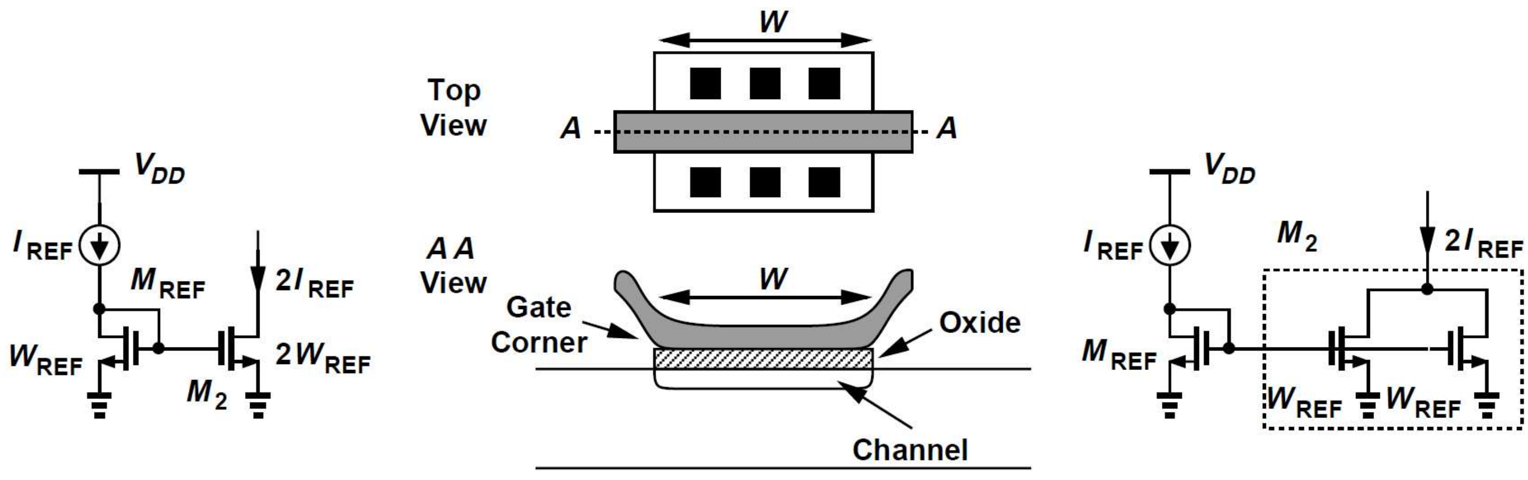
$$I_{REF} = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_1 (V_{GS} - V_{TH})^2$$

$$I_{out} = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_2 (V_{GS} - V_{TH})^2,$$

$$I_{out} = \frac{(W/L)_2}{(W/L)_1} I_{REF}.$$

- Allows precise copying of the current with no dependence on process and temperature

Sizing issues

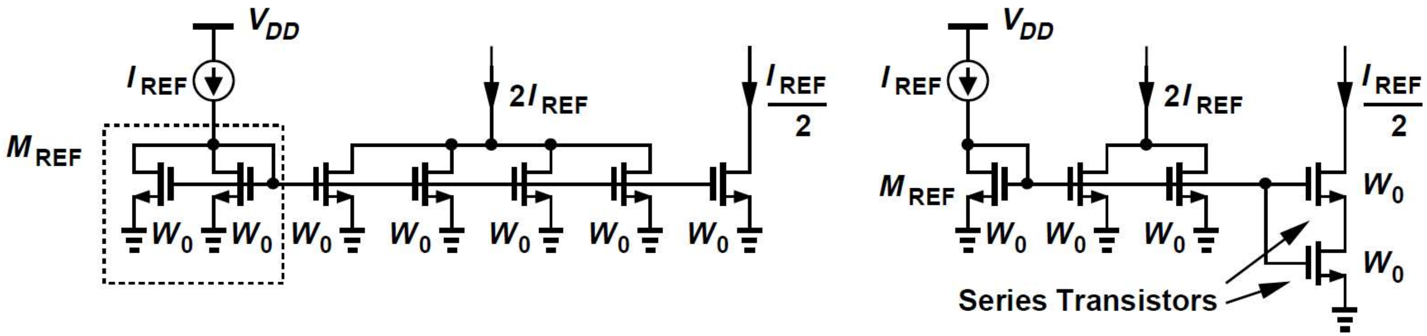


(a) (b) (c)

- Current mirrors usually employ the same length for all of the transistors.
- Current ratioing is achieved by only scaling the width of transistors.
- Direct scaling of the width also faces difficulties.
- We thus prefer to employ a “unit” transistor and create copies by repeating such a device.

Sizing Issues

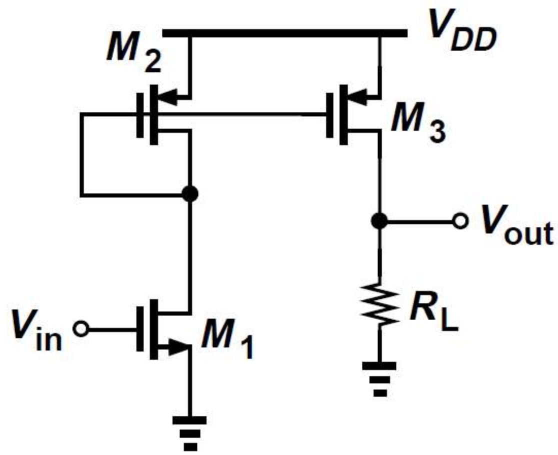
- How do we generate a current equal to I_{REF} /2=2 from I_{REF} ?



(a) (b)

- (a) half-width device, and (b) series transistors
- Approach (b) preserves an effective length of (L_{drawn}-2L_D) for each unit, yielding an equivalent length of 2(L_{drawn} - 2L_D)
- Current mirrors can process signals as well, example next slide.

Example



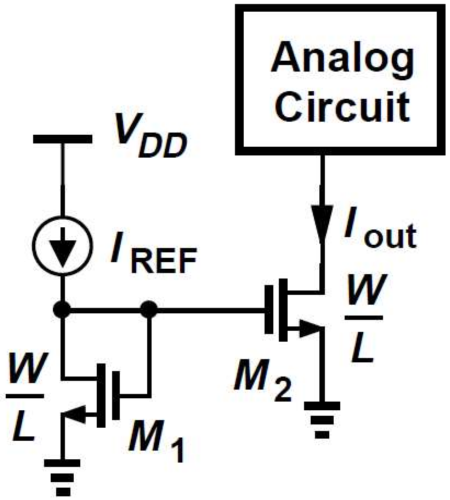
- **Calculate the small-signal voltage gain of the circuit shown in Figure.**

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

$$I_{D3} = I_{D2}(W/L)_3/(W/L)_2$$

- **Gain=** $g_{m1} R_L (W/L)_3/(W/L)_2$

Cascode Current Mirrors



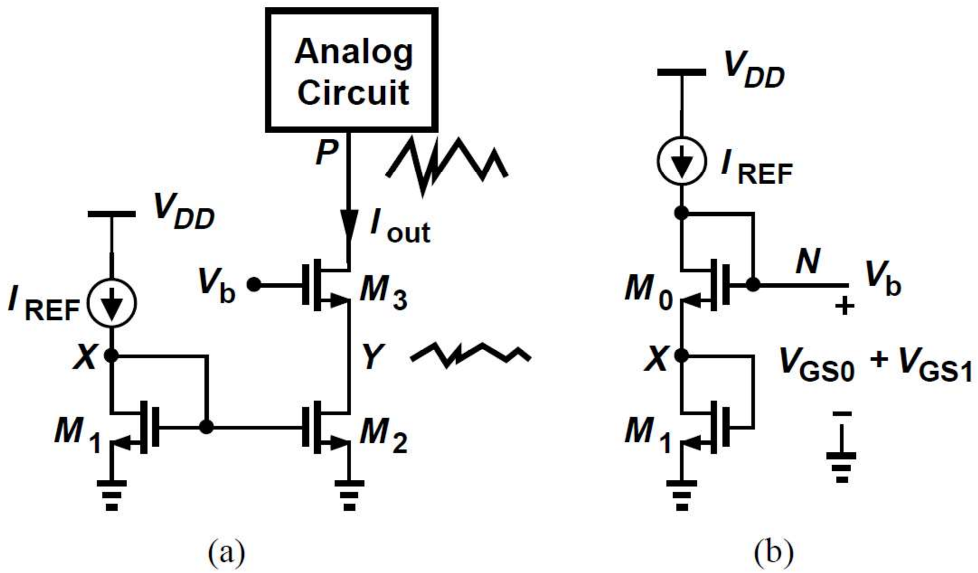
$$I_{D1} = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_1 (V_{GS} - V_{TH})^2(1 + \lambda V_{DS1})$$

$$I_{D2} = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_2 (V_{GS} - V_{TH})^2(1 + \lambda V_{DS2})$$

$$\frac{I_{D2}}{I_{D1}} = \frac{(W/L)_2}{(W/L)_1} \cdot \frac{1 + \lambda V_{DS2}}{1 + \lambda V_{DS1}}.$$

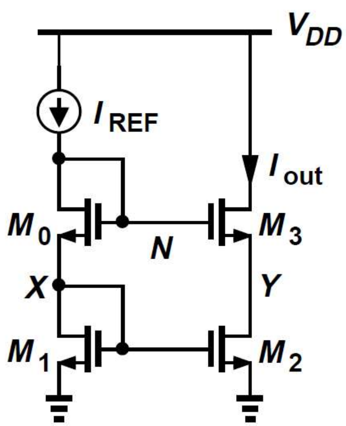
- **While** $V_{DS1} = V_{GS1} = V_{GS2}$**,** V_{DS2} **may not equal** V_{GS2}
- **We can (a) force V_{DS2} to be equal to V_{DS1}, or (b) force VDS1 to be equal to V_{DS2}.**

First Approach

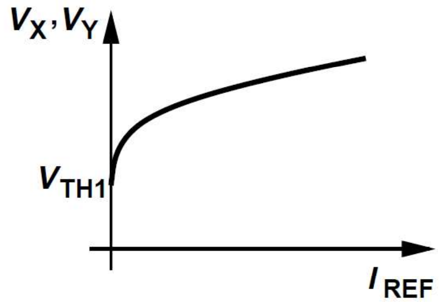


- **A cascode device can shield a current source, thereby reducing the voltage variations across it.**
- **But, how do we ensure that $V_{DS2} = V_{DS1}$?**
- **We must generate V_b such that $V_b - V_{GS3} = V_{DS1} (= V_{GS1})$**

Example



(a)



(b)

- **sketch VX and VY as a function of IREF. If IREF requires 0.5 V to operate as a current source, what is its maximum value?**

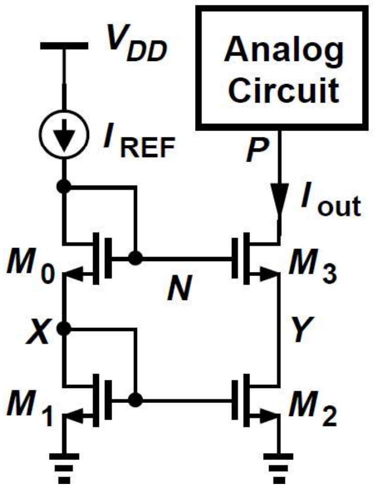
$$V_Y = V_X \approx \sqrt{2I_{REF}/[\mu_n C_{ox}(W/L)_1]} + V_{TH1}$$

$$V_N = V_{GS0} + V_{GS1}$$
$$= \sqrt{\frac{2I_{REF}}{\mu_n C_{ox}}}\left[\sqrt{\left(\frac{L}{W}\right)_0} + \sqrt{\left(\frac{L}{W}\right)_1}\right] + V_{TH0} + V_{TH1}$$

$$V_{DD} - \sqrt{\frac{2I_{REF}}{\mu_n C_{ox}}}\left[\sqrt{\left(\frac{L}{W}\right)_0} + \sqrt{\left(\frac{L}{W}\right)_1}\right] - V_{TH0} - V_{TH1} = 0.5\text{ V}$$

$$I_{REF,max} = \frac{\mu_n C_{ox}}{2}\frac{(V_{DD} - 0.5\text{ V} - V_{TH0} - V_{TH1})^2}{(\sqrt{(L/W)_0} + \sqrt{(L/W)_1})^2}.$$

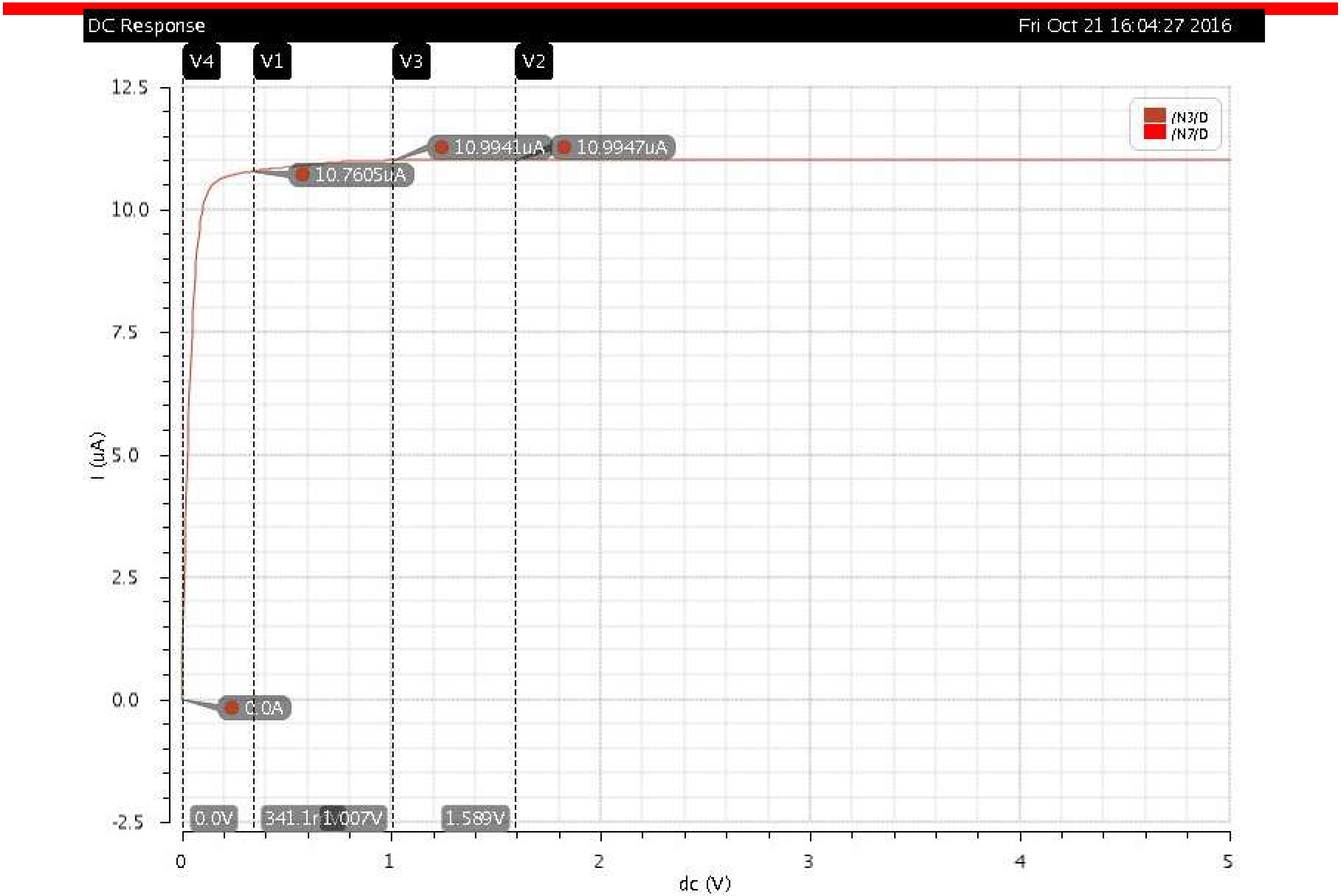
Example



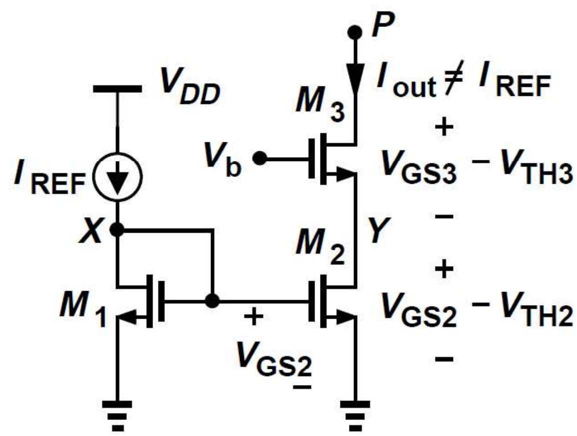
- **the minimum allowable voltage at node P is equal to**

$$\begin{aligned} V_N - V_{TH} &= V_{GS0} + V_{GS1} - V_{TH} \\ &= (V_{GS0} - V_{TH}) + (V_{GS1} - V_{TH}) + V_{TH} \end{aligned}$$

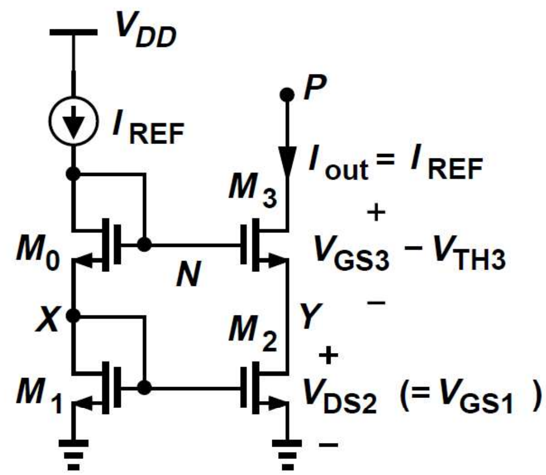
- **The cascode mirror "wastes" one threshold voltage in the headroom.**
- **Because $V_{DS2} = V_{GS2}$, whereas V_{DS2} could be as low as $V_{GS2} - V_{TH}$ while maintaining M2 in saturation.**


DC Response
Fri Oct 21 16:04:27 2016
V4
V1
V3
V2
/N3/D
/N7/D
10.9941uA
10.9947uA
10.7605uA
0.0A
0.0V
341.1r
1.007V
1.589V
I (uA)
dc (V)
12.5
10.0
7.5
5.0
2.5
0.0
-2.5
0
1
2
3
4
5

Approach summary



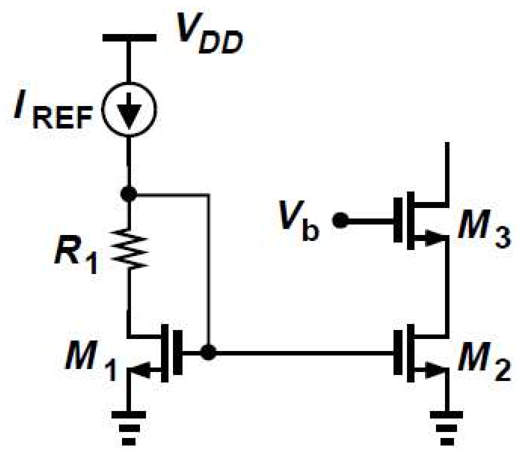
(a)



(b)

- In Fig(a), V_b is chosen to allow the lowest possible value of V_P but the output current does not accurately track I_{REF}.
- In Fig(b), a higher accuracy is achieved, but the minimum level at P is higher by one threshold voltage.

Second Approach

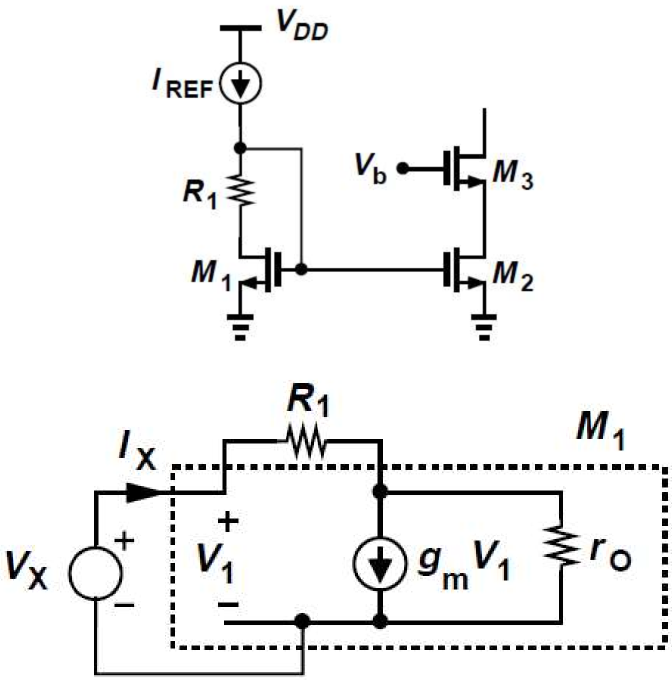


- Consider the branch shown in Fig. 5.16(b)
- As a candidate and write $V_b = V_{GS5} + R_6 I_6$.

$$R_1 I_{REF} \approx V_{TH1}$$

$$V_b = V_{GS3} + (V_{GS1} - V_{TH1})$$

Small signal Model



$$\frac{V_X - I_X R_1}{r_O} + g_m V_X = I_X.$$

$$\frac{V_X}{I_X} = \frac{R_1 + r_O}{1 + g_m r_O},$$

- Reduces to 1/gm in the absence of channel-length modulation.
- Thus, from a small-signal point of view, the combination is close to a diode-connected device.
- But

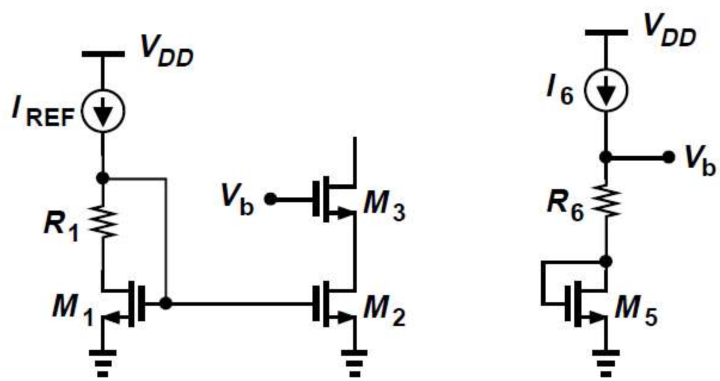
(1) It may be difficult to guarantee that $R_1 I_{REF} \approx V_{TH1}$

(2) The generation of

$$V_b = V_{GS3} + (V_{GS1} - V_{TH1})$$

is not straightforward.

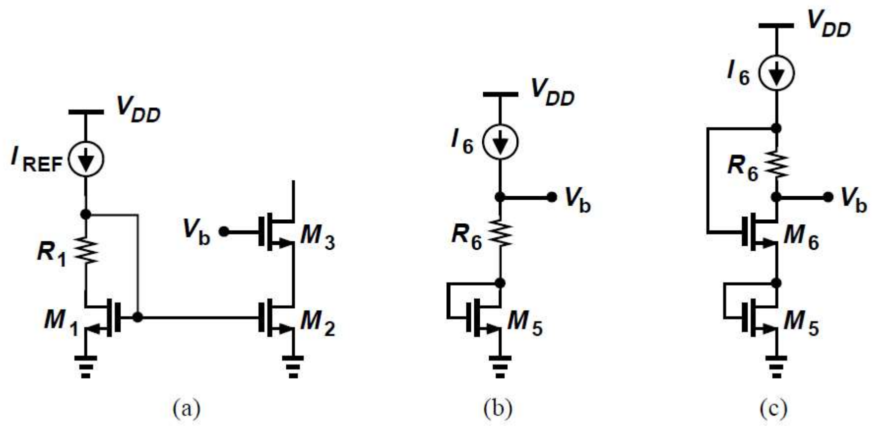
Generate Vb



(a) (b)

- **Consider the branch shown in Fig(b) as a candidate and write $V_b = V_{GS5} + R_6 I_6$.**
- **$V_{GS5} = V_{GS3}$**
- **However, the condition $R_6 I_6 = V_{GS1} - V_{TH1} = V_{GS1} - R_1 I_{REF}$ is hard to meet.**

Generate Vb

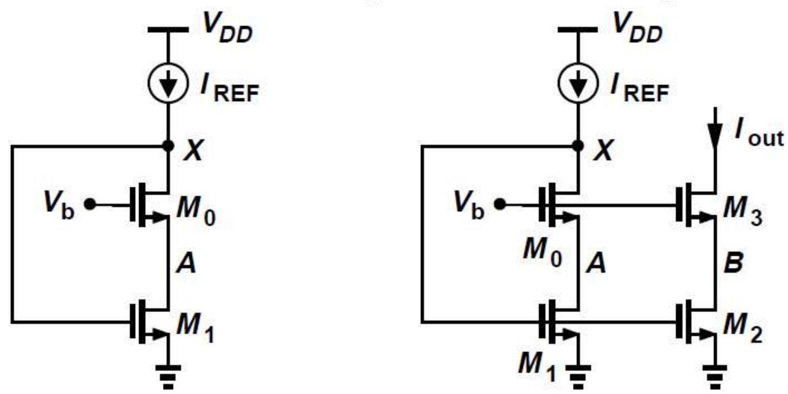


$$V_{GS5} = V_{GS3}$$

$$\begin{aligned} V_{GS6} - R_6 I_6 &= V_{GS1} - V_{TH1} \\ &= V_{GS1} - R_1 I_{REF} \end{aligned}$$

- It is now possible to ensure that V_{GS6} and V_{GS1} track each other.
- For example, we may simply choose $I_6 = I_{REF}$, $R_6 = R_1$, and (W/L)6 = (W=/L)1

Another circuit topology



(a) (b)

- **In this case**

$$V_{DS1} = V_b - V_{GS0}$$

- **Must have** $V_b - V_{TH0} \leq V_X\ (= V_{GS1})$ **for M0 to be saturated and** $V_{GS1} - V_{TH1} \leq V_A\ (= V_b - V_{GS0})$ **for M1 to be saturated.**
- **A solution exists if** $V_{GS0} + (V_{GS1} - V_{TH1}) < V_{GS1} + V_{TH0}$
- **We must therefore sizeM0 to ensure its overdrive is well below VTH1.**