類比電路設計(3349) - 2004

Passive and Active Current Mirrors

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Overview

- Reading

  B. Razavi Chapter 5.

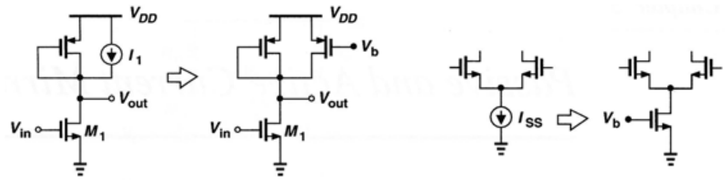
- Introduction

  In analog circuits current sources act as a large resistor without consuming excessive voltage headroom. This lecture deals with the design of current mirrors as both bias elements and signal processing components. Following a review of basic current mirrors, we study cascode mirror operation.

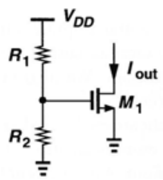
  Next, we analyze active current mirrors and describe the properties of differential pairs using such circuits as loads.

Basic current sources

- Application



- Definition of current by resistive divider

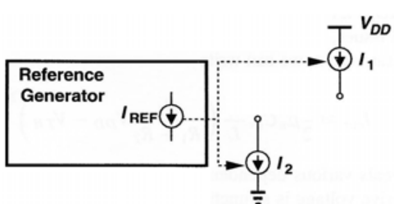


Assuming M_1 is in saturation, $I_{out} \approx \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(\frac{R_2}{R_1+R_2}V_{DD}-V_{TH}\right)^2$

- ☐ The expression reveals various dependencies of I_{out} upon the supply, process, and temperature.
- ☐ The overdrive voltage is a function of V_{DD} and V_{TH}; the threshold voltage may vary by 100mV from wafer to wafer. Furthermore, both μ_n and V_{TH} exhibit temperature dependence. Thus, I_{out} is poorly defined.

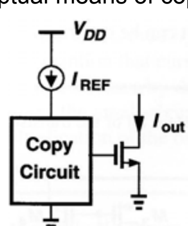
Basic current mirrors

- Use of a reference to generate various currents



A relatively complex circuit – sometimes requiring external adjustments – is used to generate a stable reference current, I_{REF}, which is then copied to many current sources in the system.

- Conceptual means of copying currents

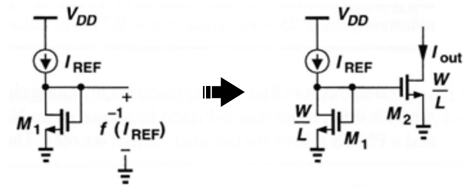


How do we guarantee $I_{out} = I_{REF}$?

Basic current mirrors (cont'd)

- Basic current mirror

Diode-connected device providing inverse function.



Neglecting channel-length modulation, we can write

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

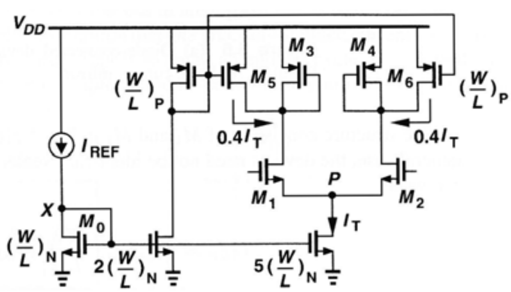
$$I_{out} = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_2 (V_{GS} - V_{TH})^2$$

obtaining

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

Basic current mirrors (cont'd)

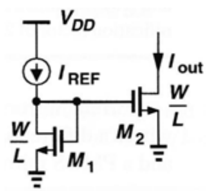
- Currents mirrors used to bias a differential amplifier



Current mirrors usually employ the same *length* for all of the transistors so as to minimize error due to the side-diffusion of the source and drain areas (L_D). Furthermore, the threshold voltage of short-channel devices exhibits some dependence on the channel length. Thus, current ratioing is achieving by only scaling the *width* of transistors.

Basic current mirrors (cont'd)

- Consider channel length modulation



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

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

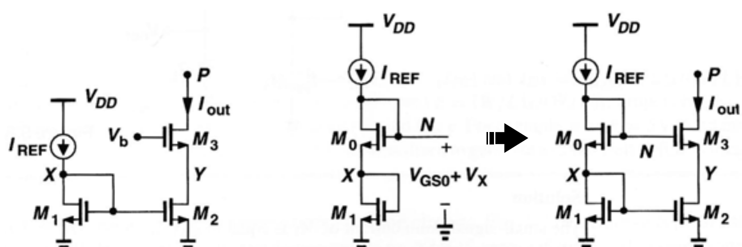
obtaining

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

While $V_{DS1} = V_{GS1} = V_{GS2}$, V_{DS2} may not equal V_{GS2} because of the circuitry fed by M_2.

Cascode current source

- Scheme – suppress the effect of channel-length modulation



How do we generate V_b in (a) to ensure $V_Y = V_X$?

Proper choice the dimensions of M_0 with respective to those of M_3 yields $V_{GS0} = V_{GS3}$.

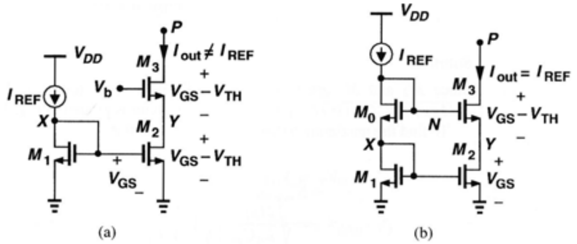
Thus, if

$$\frac{(W/L)_3}{(W/L)_0} = \frac{(W/L)_2}{(W/L)_1}$$

then $V_{GS3} = V_{GS0}$ and $V_X = V_Y$.

Cascode current source (cont'd)

- Voltage headroom consumed by a cascode mirror



$V_X \neq V_Y$ $\qquad\qquad$ $V_X = V_Y$

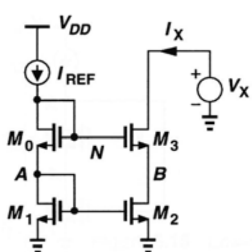
In (b), the minimum allowable voltage at node *P* is equal to

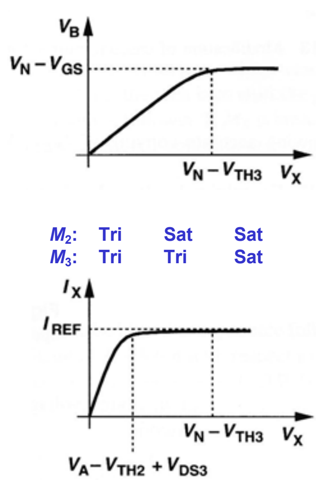
$$V_{P,min} = V_N - V_{TH} = V_{GS0} + V_{GS1} - V_{TH}$$
$$= (V_{GS0} - V_{TH}) + (V_{GS1} - V_{TH}) + V_{TH}$$

In (a), V_b is chosen to allow the lowest possible value of V_P but the output current does not accurately track I_{REF} because of $V_{DS1} \neq V_{DS2}$. In (b), higher accuracy is achieved but the minimum level at *P* is higher by one threshold voltage.

Cascode current source (cont'd)

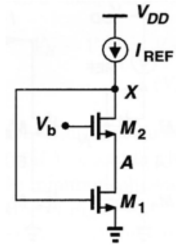
- Operation of cascode current mirror





Low-voltage cascode mirror

- Modification of cascode mirror for low-voltage operation



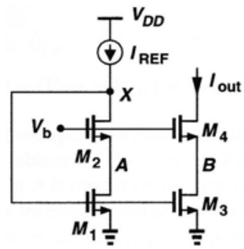
M_1, M_2 are in saturation:

M_2: $V_b - V_{TH2} \le V_X$ $(= V_{GS1})$

M_1: $V_{GS1} - V_{TH1} \le V_A$ $(= V_b - V_{GS2})$

$\Rightarrow V_{GS2} + (V_{GS1} - V_{TH1}) \le V_b \le V_{GS1} + V_{TH2}$

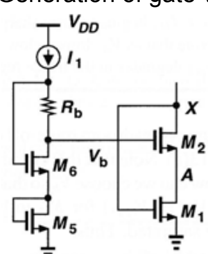
$\Rightarrow V_{GS2} - V_{TH2} \le V_{TH1}$



If $V_b = V_{GS2} + (V_{GS1} - V_{TH1}) = V_{GS4} + (V_{GS3} - V_{TH3})$, then the cascode current source M_3-M_4 consumes minimum headroom while $M1$ and $M3$ sustain equal drain-source voltages, allowing accurate copying of I_{REF}.

Low-voltage cascode mirror (cont'd)

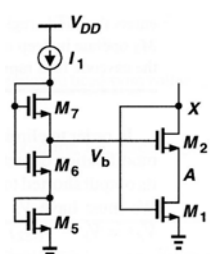
- Generation of gate voltage Vb for cascode mirror



M_1, M_2 are in saturation:

$V_{b,min} = V_{GS2} + (V_{GS1} - V_{TH1})$

Select: $V_{GS5} \approx V_{GS2}$, $V_{DS6} = V_{GS5} - RbI_1 \approx V_{GS1} - V_{TH1}$.



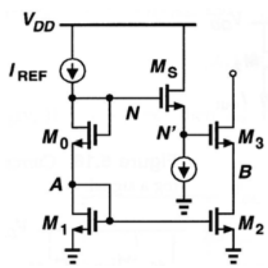
M_7 : large $(W/L)_7$, so that $V_{GS7} \approx V_{TH7}$

$\therefore V_{DS6} \approx V_{GS6} - V_{TH7}$

$V_b = V_{GS5} + V_{GS6} - V_{TH7}$

Low-voltage cascode mirror (cont'd)

- Low-voltage cascode using a source follower level shifter



If M_S is biased at a very low current density, $I_D/(W/L)$,

then $V_{GSS} \approx V_{THS} \approx V_{TH3}$, i.e., $V_N' \approx V_N - V_{TH3}$, and

$$V_B = V_{GS1} + V_{GS0} - V_{TH3} - V_{GS3}$$
$$= V_{GS1} - V_{TH3}$$

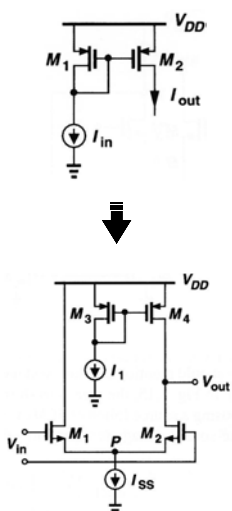
implying that M_2 is at the edge of the triode region.

In this topology, however,

- ☐ $V_{DS2} \neq V_{DS1}$
- ☐ If the body effect is considered for M_0 , M_S and M_3, it is different to guarantee that M_2 operates in saturation.

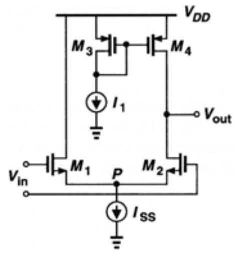
Active current mirrors

- Current mirror processing a signal

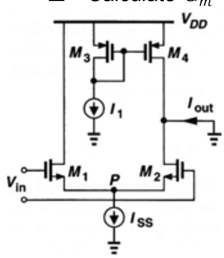


M_1 and M_2 are identical: $I_{out} = I_{in}$ (for $\lambda = 0$)

- Differential pair with current source load



☐ Calculate G_m

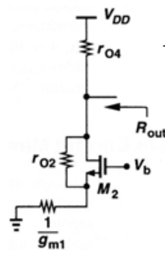


Assuming $\gamma = 0$

$$G_m = \frac{I_{out}}{V_{in}} = \frac{g_{m1}V_{in}/2}{V_{in}} = \frac{g_{m1}}{2}$$

$$|A_v| = G_m R_{out}$$

☐ Calculate R_{out}



$$R_{out1} = (1 + g_{m2}r_{o2})(1/g_{m1}) + r_{o2} = 2r_{o2} + 1/g_{m1} \approx 2r_{o2}$$

Thus, $R_{out} \approx (2r_{o2}) \| r_{o4}$

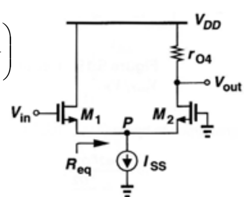
$$|A_v| \approx \frac{g_{m1}}{2}\left[(2r_{o2}) \| r_{o4}\right]$$

- Differential pair with current source load(cont'd)

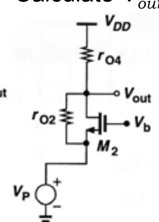
- Calculate V_p / V_{in}

$$R_{eq} \approx \frac{1}{g_{m2}} + \frac{r_{o4}}{g_{m2}r_{o2}} = \frac{1}{g_{m2}}\left(1 + \frac{r_{o4}}{r_{o2}}\right)$$

$$\frac{V_P}{V_{in}} = \frac{R_{eq}}{R_{eq} + \frac{1}{g_{m1}}} \approx \frac{1 + \frac{r_{o4}}{r_{o2}}}{2 + \frac{r_{o4}}{r_{o2}}}$$



- Calculate V_{out} / V_p



$$\frac{V_{out}}{V_P} = \frac{1 + g_{m2}r_{o2}}{1 + \frac{r_{o2}}{r_{o4}}} \approx \frac{g_{m2}r_{o2}}{1 + \frac{r_{o2}}{r_{o4}}}$$

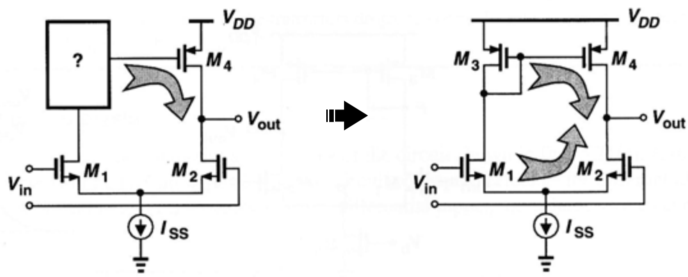
Note: if $r_{o4} \to 0$, $V_p / V_{in} \to 1/2$, and if $r_{o4} \to \infty$, $V_p / V_{in} \to 1$.

- Calculate V_{out} / V_{in}

$$\frac{V_{out}}{V_{in}} = \frac{V_P}{V_{in}} \cdot \frac{V_{out}}{V_P} = \frac{1 + \frac{r_{o4}}{r_{o2}}}{2 + \frac{r_{o4}}{r_{o2}}} \cdot \frac{g_{m2}r_{o2}}{1 + \frac{r_{o2}}{r_{o4}}} = \frac{g_{m2}r_{o2}r_{o4}}{2r_{o2} + r_{o4}} = \frac{g_{m2}}{2}\left[(2r_{o2}) \| r_{o4}\right]$$

Differential pair with active current mirrors

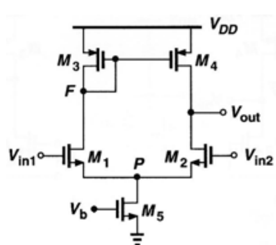
- Concept of combining the drain currents of M_1 and M_2



M_3 and M_4 are identical.

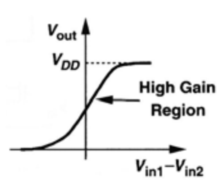
Differential pair with active current mirrors Large-signal analysis

- Large-signal analysis



Operation:

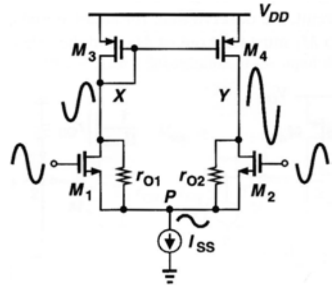
- If $V_{in1} << V_{in2}$, M_1 is off and so are M_3 and M_4. M_2 and M_5 operate in triode region, carrying zero current. Thus, $V_{out} = 0$.
- As V_{in1} approaches V_{in2} for a small difference, M_2 and M_4 are saturated, providing a high gain.
- As V_{in1} becomes more positive than V_{in2}, I_{D1}, $|I_{D3}|$, and $|I_{D4}|$ increase and I_{D2} decreases, eventually driving M_4 into the triode region.
- If $V_{in1} >> V_{in2}$, M_2 turns off, M_4 operates in deep triode region with zero current, and $V_{out} = V_{DD}$.



The choice of the input common-mode voltage:

For M_2 to be saturated, $V_{out} \geq V_{in,CM} - V_{TH}$. Thus, to allow maximum output swings, the input CM level must be as low as possible, with $V_{in,CM,\ min} = V_{GS1,2} + V_{DS5,\ min}$.

Differential pair with active current mirrors Small-signal analysis



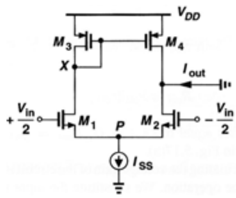
- Small-signal analysis
  - Asymmetric swings in a differential pair with active current mirror
  - Calculate G_m

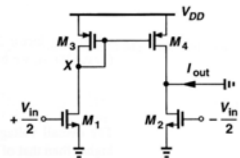
    Node *P* can be viewed as a virtual ground.

$$\begin{cases} I_{D1} = |I_{D3}| = |I_{D4}| = \dfrac{g_{m1}V_{in}}{2} \\ I_{D2} = -\dfrac{g_{m2}V_{in}}{2} \end{cases}$$

$$\Rightarrow I_{out} = I_{D2} + I_{D4} = -g_{m1,2}V_{in}$$

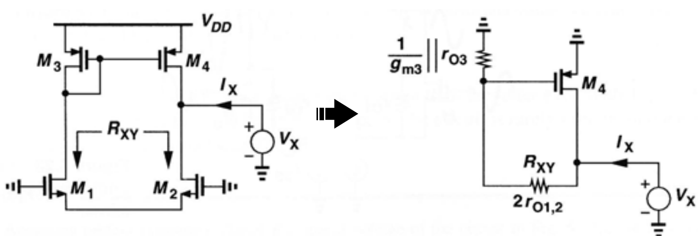
$$\therefore |G_m| = g_{m1,2}$$





Differential pair with active current mirrors Small-signal analysis (cont'd)

- Calculate R_{out}



$$I_X = 2\frac{V_X}{2r_{o1,2} + \frac{1}{g_{m3}} \| r_{o3}} + \frac{V_X}{r_{o4}}$$

where the factor 2 accounts for current copying action of M_3 and M_4.

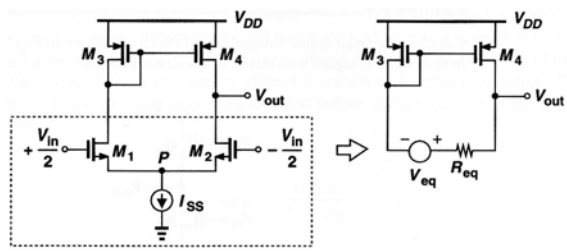
For $2r_{o1,2}$ >> $(1/g_{m3})\|r_{o3}$, we have $R_{out} \approx r_{o2} \parallel r_{o4}$

- Calculate A_v

  $|A_v| = G_m R_{out} = g_{m1,2}\,(r_{o2} \parallel r_{o4})$

Differential pair with active current mirrors Small-signal analysis (cont'd)

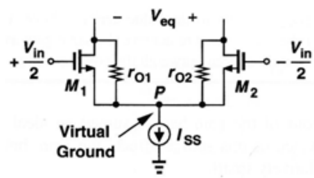
- Substitution of the input differential pair by a Thevenin equivalent



- ☐ Calculate V_{eq} and R_{eq}

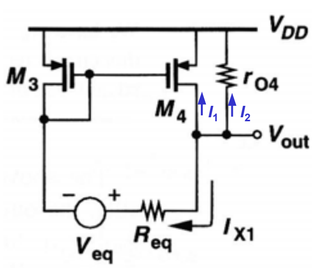
$$V_{eq} = g_{m1,2}\ r_{o1,2}\ V_{in}$$

$$R_{eq} = 2\ r_{o1,2}$$



Differential pair with active current mirrors Small-signal analysis (cont'd)

- ☐ Calculate $A_v = V_{out}\ /\ V_{in}$



The current through R_{eq} is

$$I_{X1} = \frac{V_{out} - g_{m1,2} r_{o1,2} V_{in}}{2r_{o1,2} + \frac{1}{g_{m3}} \| r_{o3}}$$

The fraction of this current that flows through $1/g_{m3}$ is mirrored into M_4 with unity gain. That is,

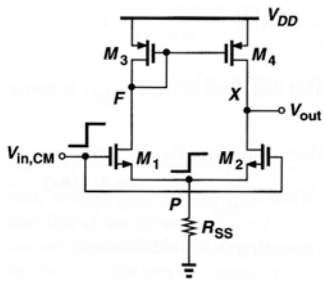
$$I_{X1} + \underbrace{\frac{V_{out} - g_{m1,2} r_{o1,2} V_{in}}{2r_{o1,2} + \frac{1}{g_{m3}} \| r_{o3}} \cdot \frac{r_{o3}}{r_{o3} + \frac{1}{g_{m3}}}}_{I_1} + \underbrace{\frac{V_{out}}{r_{o4}}}_{I_2} = 0$$

Assuming $2r_{o1,2} >> (1/g_{m3,4}) \| r_{o3,4}$, we obtain

$$\frac{V_{out}}{V_{in}} = \frac{g_{m1,2} r_{o3,4} r_{o1,2}}{r_{o1,2} + r_{o3,4}} = g_{m1,2}\left(r_{o1,2} \| r_{o3,4}\right)$$

Differential pair with active current mirrors Common-mode properties

- Differential pair with active current mirror sensing a common-mode change



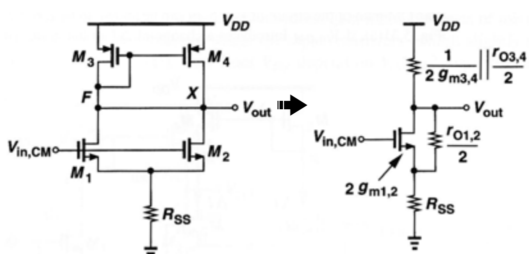
The CM gain is defined in terms of the single-ended output component produced by the input CM change:

$$A_{CM} = \frac{\Delta V_{out}}{\Delta V_{in,CM}}$$

Differential pair with active current mirrors Common-mode properties (cont'd)

- Simplified circuit of CM circuit

$$A_{CM} \approx -\frac{\dfrac{1}{2g_{m3,4}} \Big\| \dfrac{r_{o3,4}}{2}}{\dfrac{1}{2g_{m1,2}} + R_{SS}} = -\frac{1}{1+2g_{m1,2}R_{SS}} \cdot \frac{g_{m1,2}}{g_{m3,4}}$$



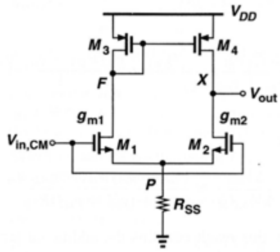
where we have assumed $1/(2g_{m3,4}) << ro_{3,4}$ and neglected the effect of $r_{o1,2}/2$. Even with perfect symmetry, the output signal is corrupted by input CM variations, a drawback that does not exist in the fully differential circuits.

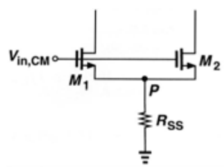
- CMRR

$$CMRR = \left|\frac{A_{DM}}{A_{CM}}\right| = g_{m1,2}(r_{o1,2} // r_{o3,4}) \frac{g_{m3,4}(1+2g_{m1,2}R_{SS})}{g_{m1,2}} = \underline{(1+2g_{m1,2}R_{SS})g_{m3,4}(r_{o1,2} // r_{o3,4})}$$

Differential pair with active current mirrors Mismatch

- Differential pair with g_m mismatch





Considering M_1 and M_2 as a single transistor with $g_m = g_{m1} + g_{m2}$,

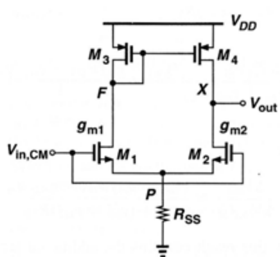
$$\Delta V_P = \Delta V_{in,CM} \frac{R_{SS}}{R_{SS} + \dfrac{1}{g_{m1} + g_{m2}}}$$

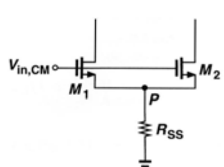
where body effect is neglected. The change of I_{D1} and I_{D2} are given by

$$\Delta I_{D1} = g_{m1}(\Delta V_{in,CM} - \Delta V_P) = \frac{\Delta V_{in,CM}}{R_{SS} + \dfrac{1}{g_{m1} + g_{m2}}} \frac{g_{m1}}{g_{m1} + g_{m2}}$$

$$\Delta I_{D2} = g_{m2}(\Delta V_{in,CM} - \Delta V_P) = \frac{\Delta V_{in,CM}}{R_{SS} + \dfrac{1}{g_{m1} + g_{m2}}} \frac{g_{m2}}{g_{m1} + g_{m2}}$$

Differential pair with active current mirrors Mismatch (cont'd)





And $|\Delta I_{D4}| = g_{m4}|\Delta V_{GS4}| = g_{m4}\left(\dfrac{1}{g_{m3}} // r_{o3}\right)\Delta I_{D1}$

Neglecting the effect of r_{o1} and r_{o2} :

$$\begin{aligned}\Delta V_{out} &= (\Delta I_{D4} - \Delta I_{D2}) r_{o4} \\ &= \left[\frac{g_{m1}\Delta V_{in,CM}}{1 + (g_{m1} + g_{m2})R_{SS}} \frac{r_{o3}}{r_{o3} + \dfrac{1}{g_{m3}}} - \frac{g_{m2}\Delta V_{in,CM}}{1 + (g_{m1} + g_{m2})R_{SS}}\right] r_{o4} \\ &= \frac{\Delta V_{in,CM}}{1 + (g_{m1} + g_{m2})R_{SS}} \frac{(g_{m1} - g_{m2}) r_{o3} - g_{m2}/g_{m3}}{r_{o3} + \dfrac{1}{g_{m3}}} r_{o4}\end{aligned}$$

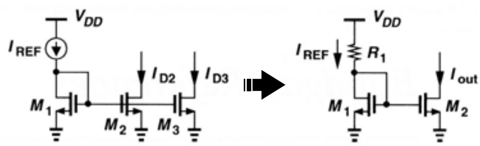
If $r_{o3} >> 1/g_{m3}$, we have

$$\frac{\Delta V_{out}}{\Delta V_{in,CM}} \approx \frac{(g_{m1} - g_{m2}) r_{o3} - g_{m2}/g_{m3}}{1 + (g_{m1} + g_{m2})R_{SS}}$$

$(g_{m1} - g_{m2}) r_{o3}$ reveals the effect of the transconductance Mismatch on the common-mode gain.

Supply-independent biasing

- Current-mirror biasing using (a) an ideal current source, (b) a resistor.

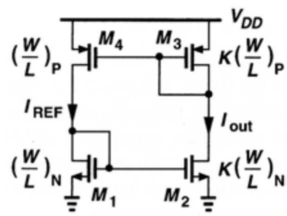


Te output current is quite sensitive to V_{DD}:

$$\Delta I_{out} = \frac{\Delta V_{DD}}{R_1 + 1/g_{m1}} \cdot \frac{(W/L)_2}{(W/L)_1}$$

How do we generate I_{REF} independent of the supply voltage?

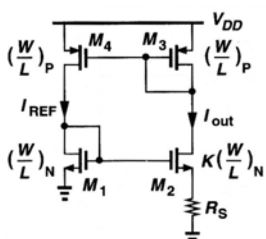
- Simple circuit to establish supply-independent currents.



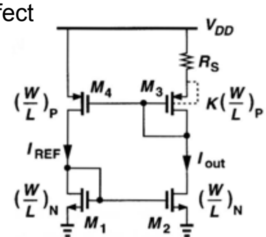
- ☐ In order to arrive at a less sensitive solution, we postulate that the circuit must *bias itself*, i.e. , I_{REF} must be somehow derived from I_{out}.
- ☐ If M_1-M_4 operate in saturation and λ = 0, then I_{out} = KI_{REF}, and hence can support *any* current level.

Supply-independent biasing

- Addition of R_S to define the currents



- Alternative implementation eliminating body effect



Assuming λ = 0, then I_{out} = I_{REF} and V_{GS1} = V_{GS2} + $I_{D2}R_S$

$$\sqrt{\frac{2I_{out}}{\mu_n C_{ox}(W/L)_N}} + V_{TH1} = \sqrt{\frac{2I_{out}}{\mu_n C_{ox} K(W/L)_N}} + V_{TH2} + I_{out}R_S$$

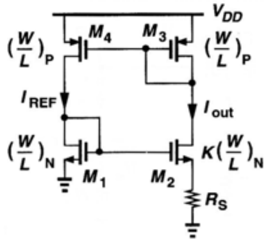
Neglecting body effect, we have $\sqrt{\frac{2I_{out}}{\mu_n C_{ox}(W/L)_N}}\left(1 - \frac{1}{\sqrt{K}}\right) = I_{out}R_S$

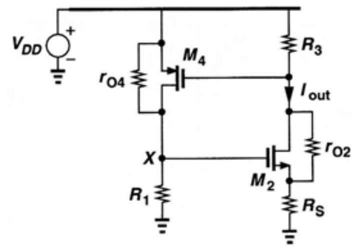
That is, $I_{out} = \frac{2}{\mu_n C_{ox}(W/L)_n} \cdot \frac{1}{R_S^2}\left(1 - \frac{1}{\sqrt{K}}\right)^2$

The current is independent of the supply voltage (but still a function of process and temperature).

Supply-independent biasing (cont'd)

- Addition of R_S to define the currents (assuming $\lambda \neq 0$). Determine $\Delta I_{out} / \Delta V_{DD}$.





$R_1 = r_{o1} \parallel (1/g_{m1})$, $R_3 = r_{o3} \parallel (1/g_{m3})$

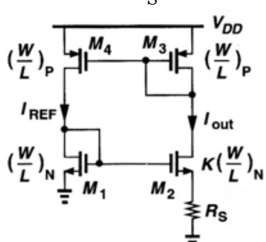
$$\frac{V_{DD} - V_X}{r_{o4}} + I_{out} R_3 g_{m4} = \frac{V_X}{R_1}$$

The equivalent transconductance of M_2 and R_S is $G_{m2} = \dfrac{I_{out}}{V_X} = \dfrac{g_{m2} r_{o2}}{R_S + r_{o2} + (g_{m2} + g_{mb2}) R_S r_{o2}}$

Thus, $\dfrac{I_{out}}{V_{DD}} = \dfrac{1}{r_{o4}} \left[\dfrac{1}{G_{m2}(r_{o4} \| R_1)} - g_{m4} R_3 \right]^{-1} \rightarrow 0$, if $r_{o4} = \infty$.

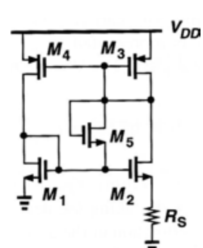
Supply-independent biasing (cont'd)

- Addition of R_S to define the currents



  - An important issue in supply-independent biasing is the existence of "degenerate" bias points. For example, if all the transistors carry zero current when the supply is turned on, they may remain off indefinitely because the loop can support a zero current in both branches.
  - In other words, the circuit can settle in one of *two* different operating condition

- Addition of start-up device



  - The diode-connected device M_5 provides a current path from V_{DD} through M_3 and M_1 to ground upon start-up.
  - This technique is practical on if $V_{TH1} + V_{TH5} + |V_{TH3}| < V_{DD}$ and $V_{GS1} + V_{TH5} + |V_{GS3}| > V_{DD}$, the latter to ensure M_5 remains off after start-up.