Low Voltage Power Supply Current Source

- Simple implementation of a current source in many applications including a tail current yields a low output impedance.
- Cascode implementations of current sources yields a larger output impedance, however the trade-off is the reduction on the headroom.
- One alternative implementation of a current source with higher output impedance without sacrificing significant overhead is presented next.

TABLE I

COMPARISON BETWEEN CURRENT SOURCE TOPOLOGIES

Current Source Type	R_{out}	$V_{compliance}$
Simple	r_o	$V_{DS_{sat}}$
Cascode	$r_o^2 g_m$	$2V_{DS_{sat}} + V_T$
Regulated Cascode [1]	$r_o^3 g_m^2$	$2V_{DS_{sat}} + V_T$
Generalized Cascode [2]	$r_o^2 g_m$	$2V_{DS_{sat}}$
Active Regulated Cascode [3]	$A r_o^2 g_m$	$2V_{DS_{sat}}$
This Work	$r_o^2 g_m$	$V_{DS_{sat}}$

r_o is the output resistance of a MOS transistor and $V_{DS_{sat}}$ is the drain-source saturation voltage.

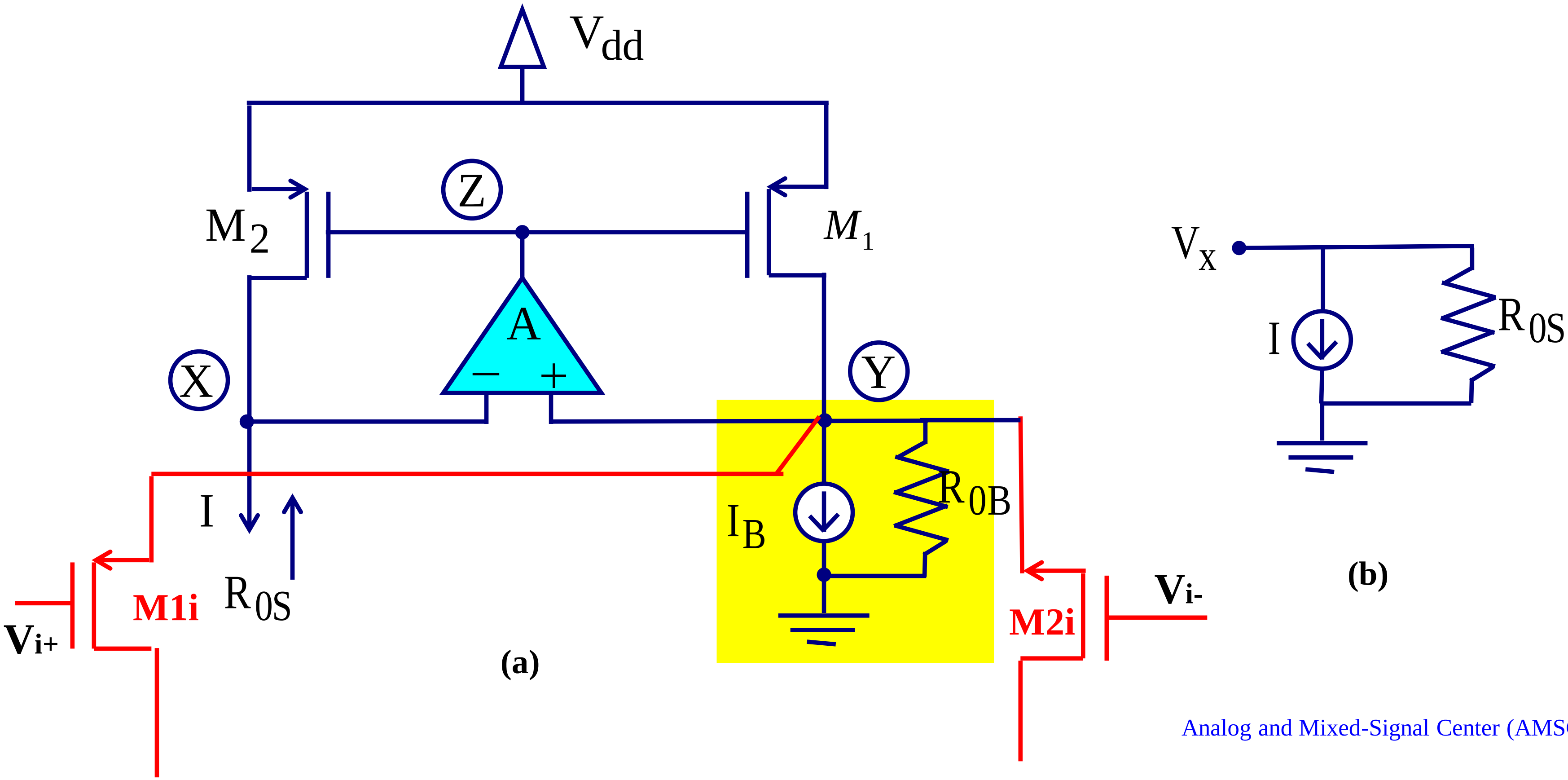
- The current source discussed here has an output resistance about 25 times larger than that of a single transistor current source.

- This current source improves the common-mode input range and the CMRR of differential pairs.

- Another use of this LV CS is for the voltage follower topologies.

Low Voltage CURRENT-SOURCE

This current source is applied at the PMOS differential pair (M2i and M1i) at node Y



A conceptual Schematic of the low voltage current source. (a) Current source representation (b) Architecture

You, F., Embabi, H.K., Duque-Carrillo, J.F., Sanchez-Sinencio, E., "An improved tail current source for low voltage applications ,"

IEEE Journal of Solid-State Circuits , Volume: 32, Issue: 8 , Aug 1997, Page(s): 1173 –1180

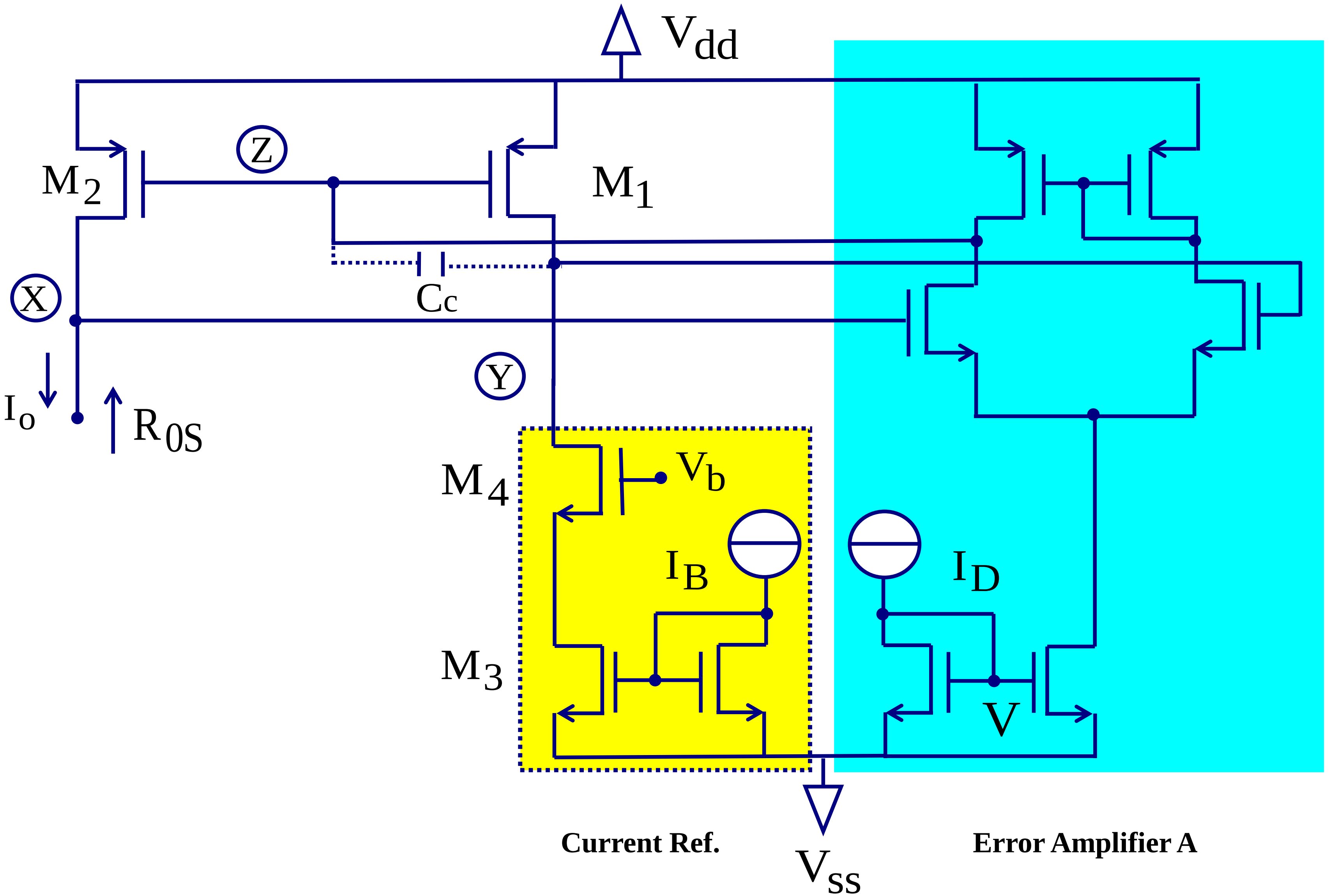
$$R_{os} = \frac{1 + g_{m1}A_o/(g_{o1} + g_{oB})}{g_{o2}(1 + A_o g_{m1}/(g_{o1} + g_{oB}) - A_o g_{m2}/g_{o2})} \quad (1)$$

where $g_{m1}(g_{m2}), g_{o1}(g_{o2})$ **are the transconductance and output conductance of** $M_1(M_2)$**, respectively.** A_o **is the DC gain of the error amplifier "A" and** $g_{oB}(R_{oB})$ **is the output conductance (resistance) of the reference current source** I_B**. Assuming that** $g_{m1} = g_{m2}$ **and** $g_{o1} = g_{o2}$**, equation (1) can be simplified as:**

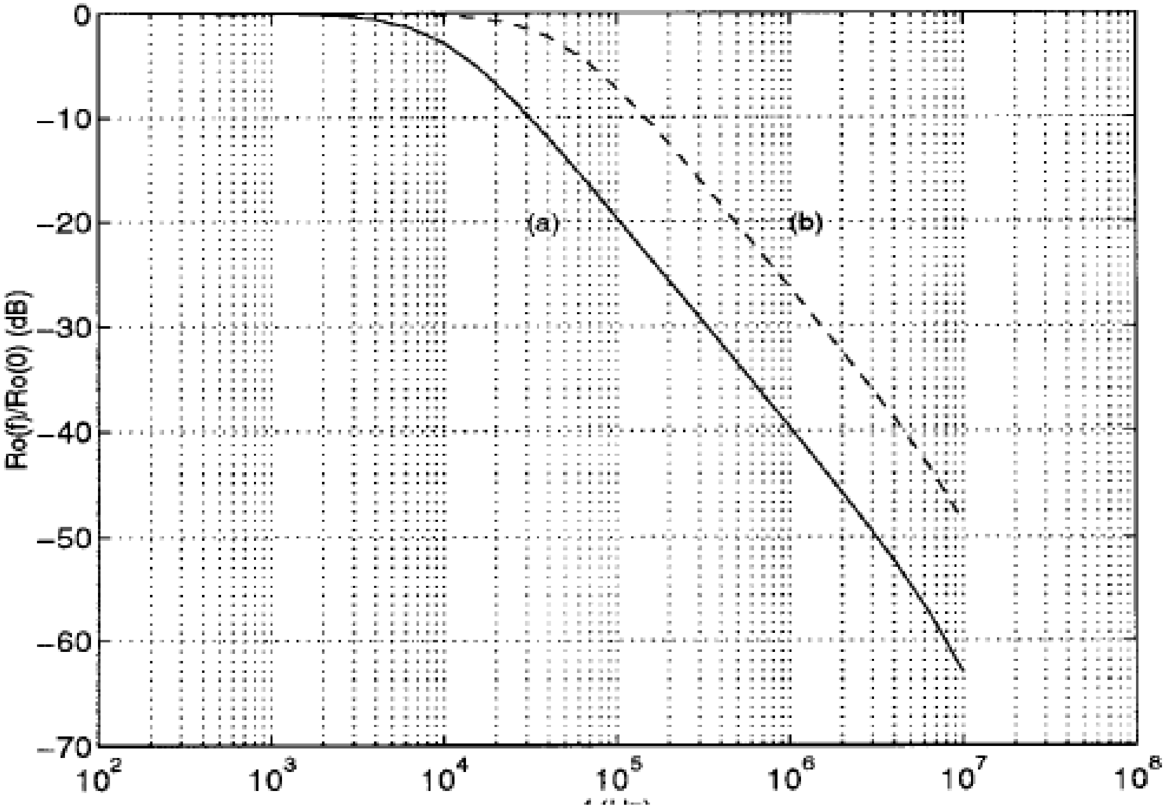
$$R_{os} \approx -R_{oB} \quad (2)$$

Note that the resistance is negative and is equal to the resistance of the reference source I_B**.**

$$R_{os} \approx -\frac{g_{m4}}{g_{o3}g_{o4}} \quad (3)$$



Full implementation of the LV current source.



FREQUENCY RESPONSE OF THE TAIL CURRENT OUTPUT IMPEDANCE (a) WITHOUT DIFFERENTIAL PAIR AND (b) WITH DIFFERENTIAL PAIR

Stability Considerations

Stability Considerations

- This LV Current Sources has one negative feedback loop and one positive feedback loop.
- The negative feedback (consisting of the error amplifier A and transistor M1)must be greater than the positive(which consists of A and transistor M2) to have stability.
- The analysis is based on the small signal equivalent circuit shown in the next slide.
- Cx and Cy are the parasitic capacitances at nodes X and Y.
- go_{B} is the conductance of the current source I_{B}
- The error amplifier is characterized by one dominant pole at $\mathbf{C_{Z}}$/goa

$$H_{OL} = \frac{v_o(s)}{v_i(s)},$$

It can be shown that

$$H_{OL}(s) = A_L^-(s) - A_L^+(s) \quad (5)$$

where A_L^- is the gain of the negative loop and is given by

$$A_L^- = \frac{A_o g_{m1}/(g_{o1} + g_{oB})}{(1 + sC_z/g_{oa})(1 + sC_y/(g_{o1} + g_{oB}))}$$

and A_L^+ is the gain of the positive loop and is given by

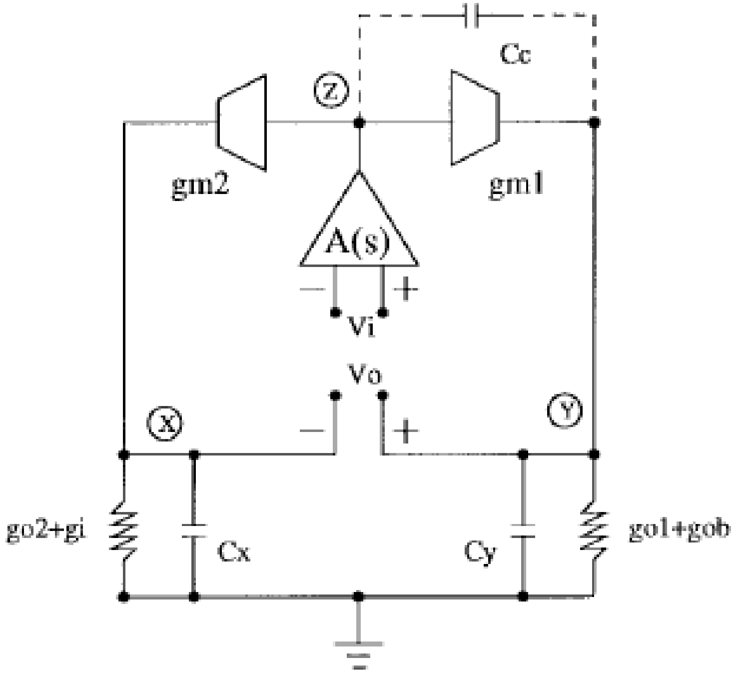
$$A_L^+ = \frac{A_o g_{m2}/(g_{o2} + g_i)}{(1 + sC_z/g_{oa})(1 + sC_x/(g_{o2} + g_i))}.$$

The $g_{m1}(g_{m2})$ and $g_{o1}(g_{o2})$ account for the transconductance and output conductance of $M_1(M_2)$. Equation (5) can be simplified as

$$H_{OL}(s) = \frac{A_o g_m(g_i - g_{oB})}{g_o(g_o + g_i)} \times \frac{1 + s(C_x - C_y)/(g_i - g_{oB})}{(1 + s/p_x)(1 + s/p_y)(1 + s/p_z)} \quad (6)$$

where

$$p_z = \frac{g_{oa}}{C_z}, \quad p_y = \frac{(g_o + g_{oB})}{C_y}, \quad p_x = \frac{(g_i + g_o)}{C_x},$$



Open loop transfer function with Cc

$$H_{OL} = \frac{A_o g_m(g_i - g_{oB})}{(g_o + g_{oB})(g_o + g_i)} \times \frac{(1 - s/z_1)(1 + s/z_2)}{(1 + s/p_x)(1 + s/p'_y)(1 + s/p'_z)} \quad (7)$$

$$z_1 = \frac{g_i - g_{oB}}{C_c} \times \frac{g_m}{g_m + g_i} \qquad p'_y = \frac{g_m}{C_y + C_x}$$

$$z_2 = \frac{g_m}{C_y} \qquad p'_z = \frac{g_{oa} g_o}{g_m C_c}.$$

- The frequency response is simulated for two cases; i) With a simple transistor tail current and ii) using the proposed LV current source.
- The effect of Rc is shown in curve (c)
- Fig 5 (a) does not have Cc, thus its poor phase margin. Adding Rc to CC in series yields Fig 5(c)

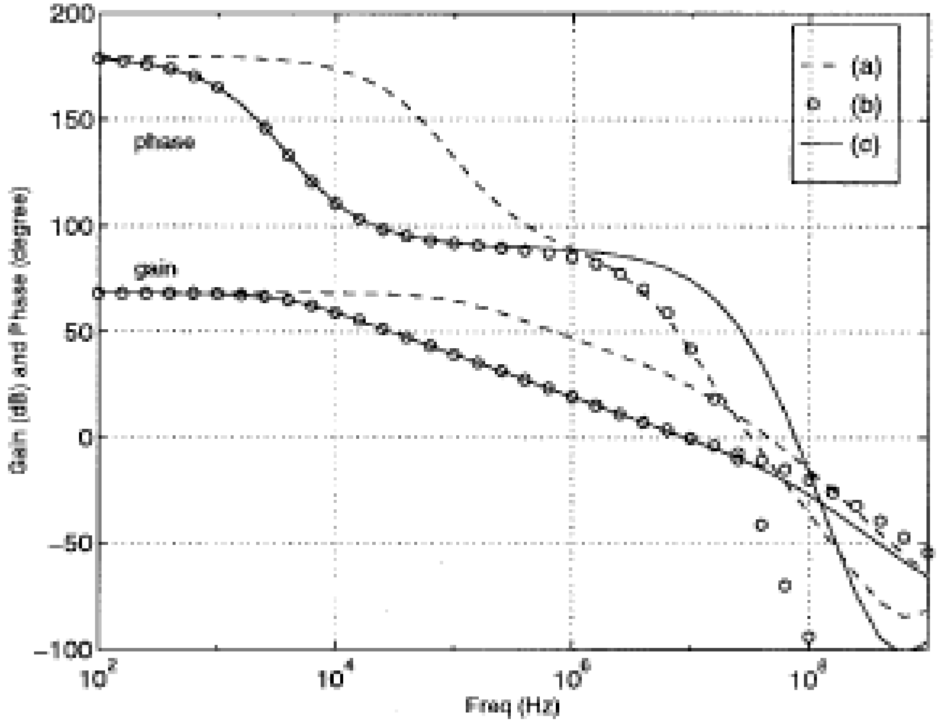
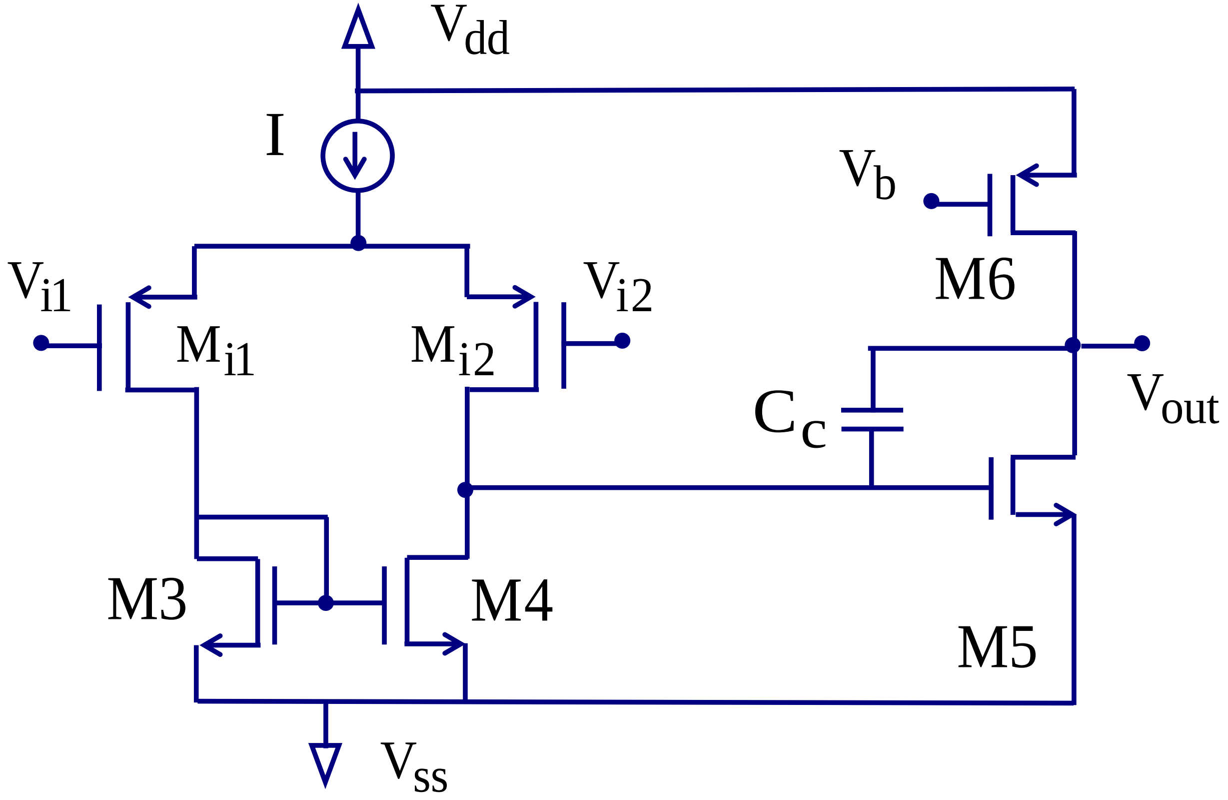


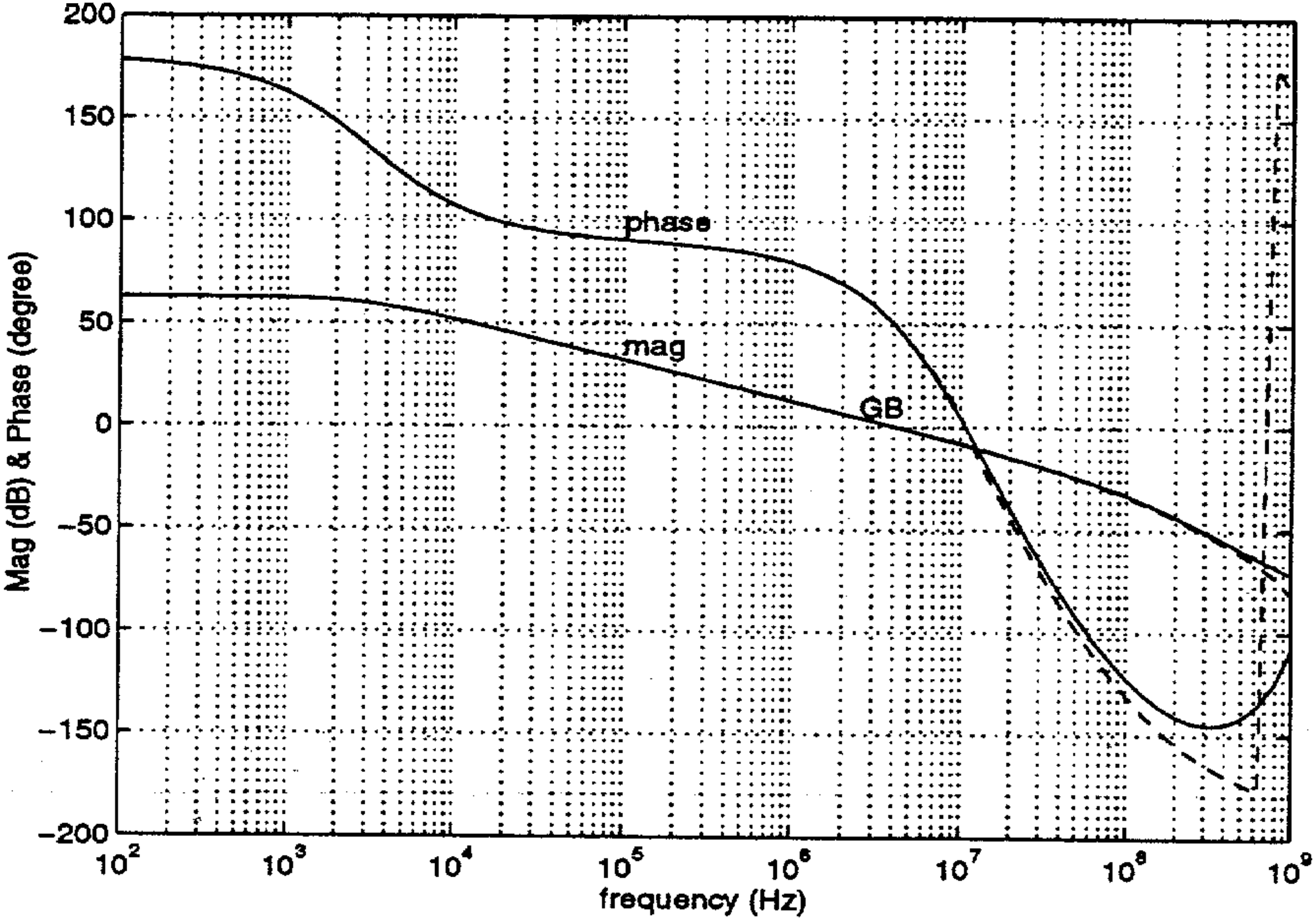
Fig. 5. Open-loop frequency response of the tail current source.

The effect of the Miller capacitance has been verified through the simulation. Curve (b) in Fig. 5 demonstrates that a 45° phase margin can be achieved by using a 1-pF compensation capacitance C_C. Since the LV current source will be mainly used to provide tail current for circuits with low input impedance, such as the differential pairs, g_k will be of the order of g_m. This implies that z_1, which is a right-half plane (RHP) zero, will be close to the unity gain frequency, thus causing significant reduction in the phase margin. To overcome this problem, a resistance R_C can be added in series with C_C to push z_1 to higher frequencies (as is the case for the conventional two-stage CMOS amplifier). It can be shown

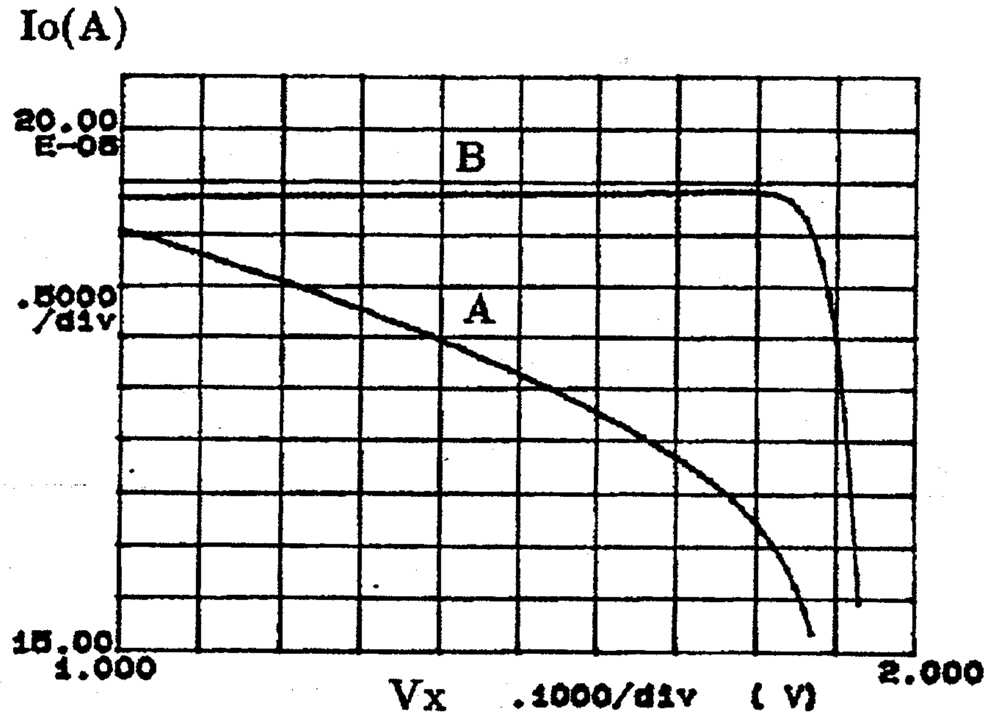


Use of the proposed current source in simple two stage amplifier.

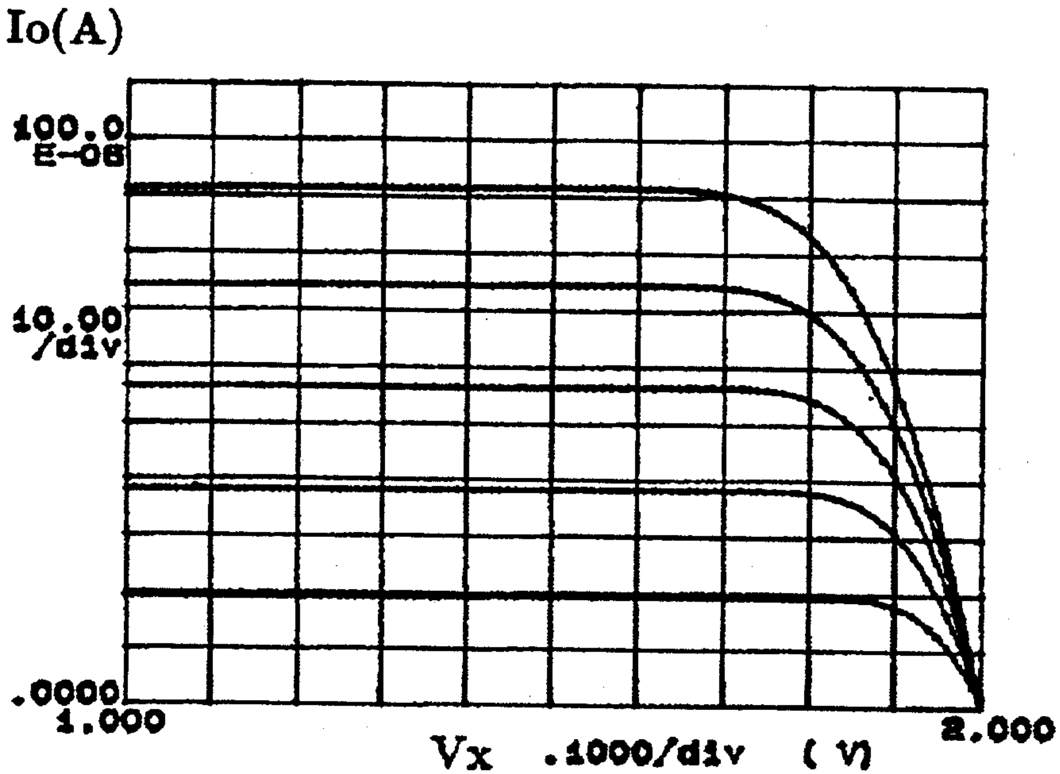
Cc has a resistor in series Rc to provide better stability.



Frequency response of two stage amplifiers using LV current source (dashed) and simple current source (solid).

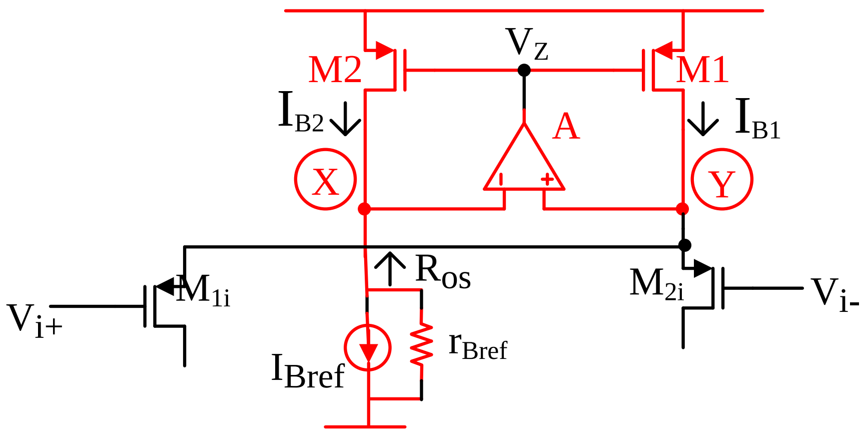


Measured output current of the simple (curve A) and LV (Curve B) current source.



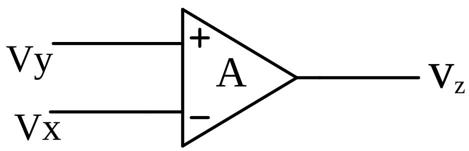
Measured I-V characteristics of the LV current source.

LOW VOLTAGE CURRENT SOURCES

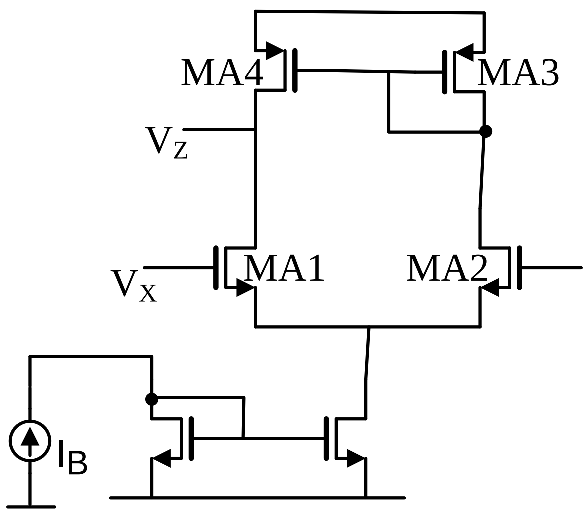


- Amplifier "A" forces the V_{DS} of M1 and M2 to be equal. Thus $I_{B1} = I_{B2}$.
- $R_{os} = -R_{Bref}$
- $V_{REF}^{MAX} = V_{y}^{MAX} = V_{DD} - V_{SG,AUX} - V_{SD,AUX}$
- Large variations of V_Z are caused when M1 and M2 enter in triode mode.

Using an auxiliary differential amplifier LV Current Source

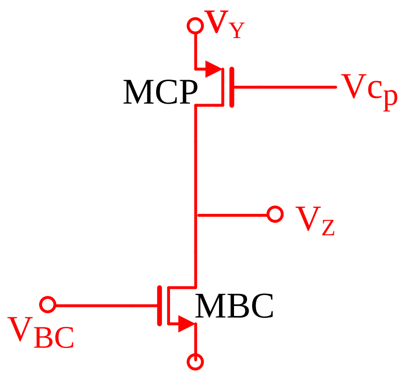


$$A(s) = \frac{A_o}{1 + sC_Z / g_{Oa}}$$



An Alternative Approach for a LV Current Source*

- The auxiliary amplifier is implemented by NMOS cascoding transistor MCP connected between node Z and node y.



Thus

$$A = \frac{V_Z}{V_Y} = g_{m_{MCP}} r_{o_{MCP}}$$

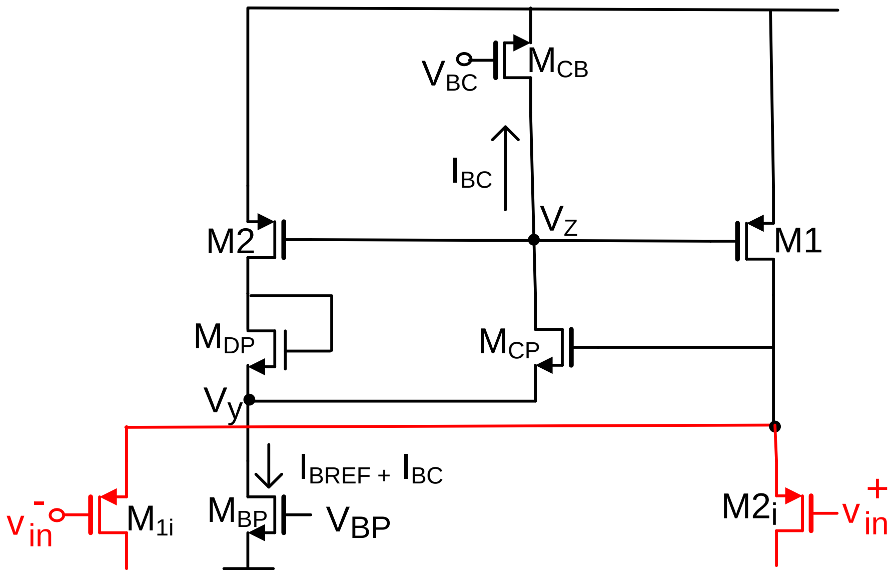
$$V_y = V_{CP} + V_{SG}$$

Then

$$V_y^{MAX} = V_{DD} - \left|V_{DS_{SAT}}\right|$$

V_{icm} can be extended to the upper rail $V_{icm}^{MAX} = V_{DD}$

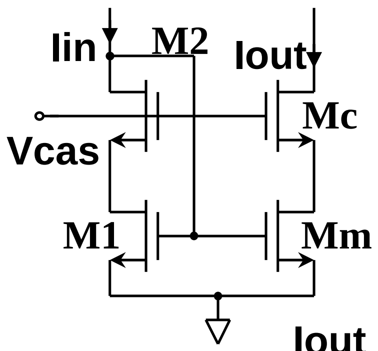
* J. Ramirez-Angulo, R. G. Carvajal and A. J. Lopez-Martin, “Single Transistor High Impedance Tail Current Source with Extended Common-Mode Input Range and Reduced Supply Requirements”, *IEEE Trans. on Circuits and Systems II*, Vol. 54, No. 7, pp. 581-585, July 2007.

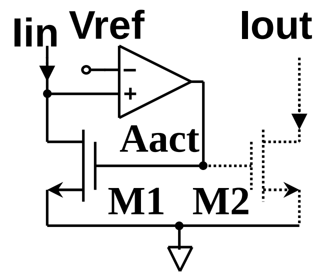


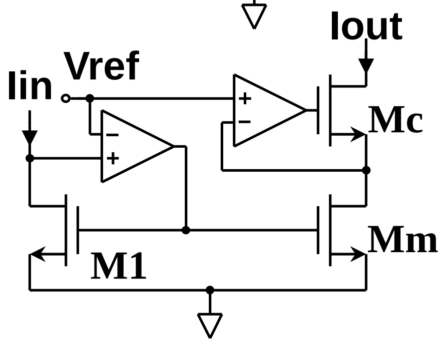
- To reduce mismatch betweeen V_{DS1} and V_{DS2} a diode connected NMOS MDP is inserted.

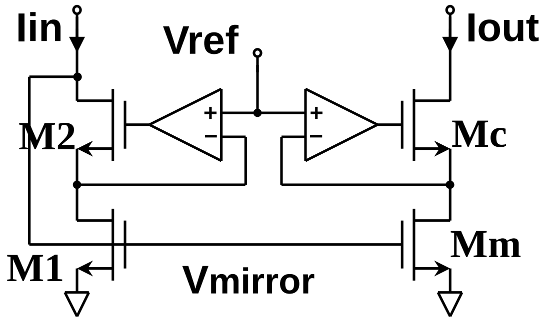
- Implementation of an enhanced LV current source where amplifier is implemented by a simple cascoding (two transistor MCB and MBP)
- $(W/L)_{1i,2i} = (W/L)$
- $(W/L)_{1,2} = 2(W/L)$
- $(W/L)_{MPB} = 4(W/L)$

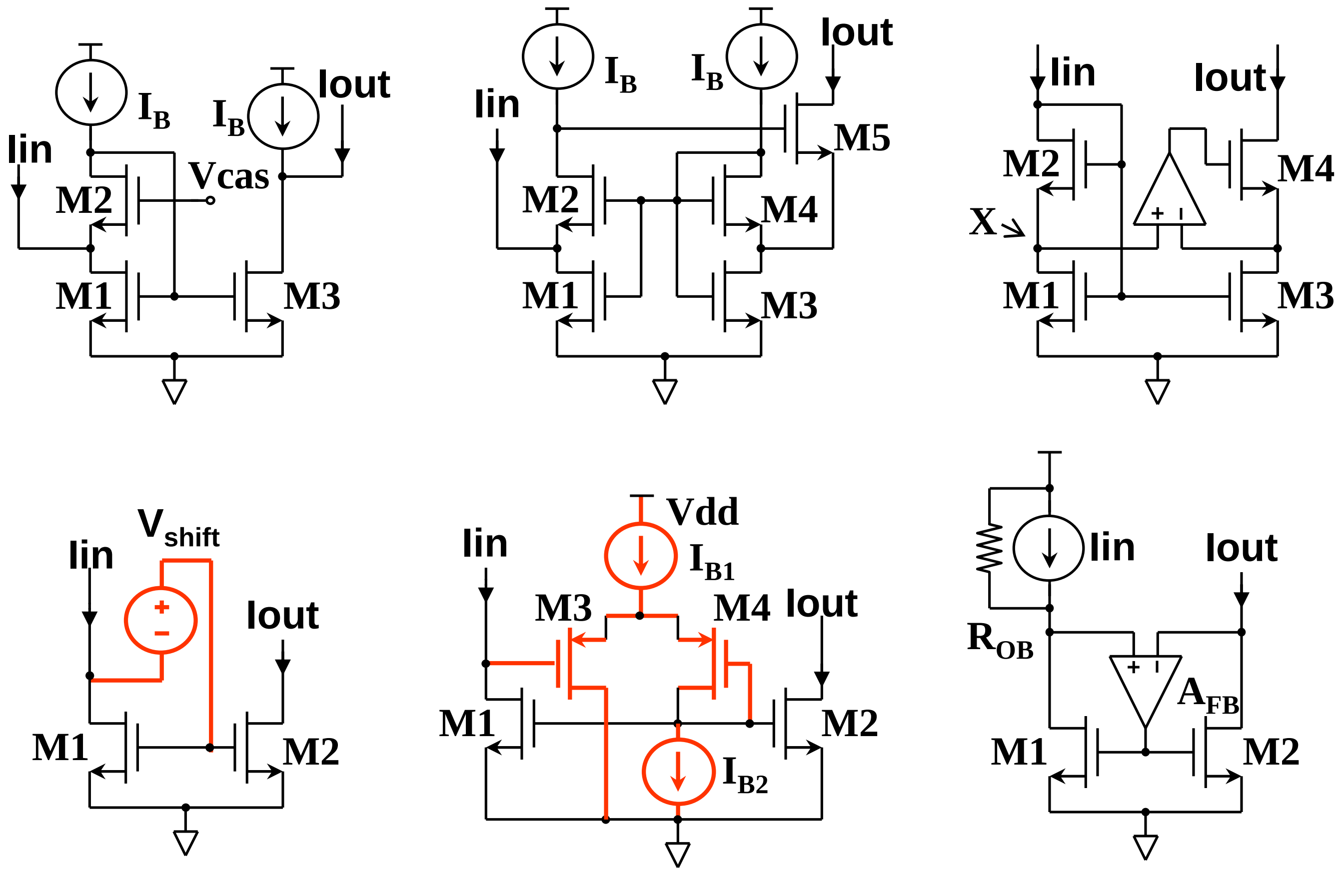
Goals: To reduce the input impedance and to increase the output impedance, while keeping the voltage operation











I_B
I_B
Iout
Iin
Vcas
M2
M1
M3
I_B
I_B
Iout
Iin
M5
M2
M4
M1
M3
Iin
Iout
M2
M4
X
+
M1
M3
V_{shift}
Iin
Iout
+
−
M1
M2
Vdd
Iin
I_{B1}
M3
M4
Iout
M1
M2
I_{B2}
Iin
Iout
R_{OB}
+
A_{FB}
M1
M2