Fully Differential Operational Amplifiers

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February 25, 2003

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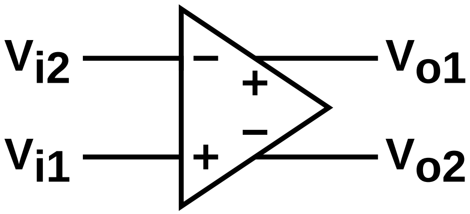
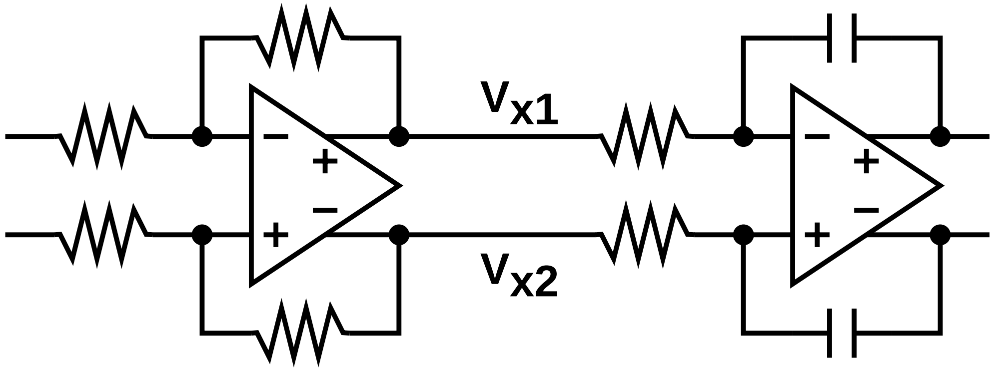
Fully Balanced Circuit Topology





$$\begin{bmatrix} v_{od} \\ v_{oc} \end{bmatrix} = \begin{bmatrix} A_{dm} & A_{cdm} \\ A_{dcm} & A_{cm} \end{bmatrix} \begin{bmatrix} v_{id} \\ v_{ic} \end{bmatrix}$$

$$V_{id} = V_{i1} - V_{i2} \qquad V_{ic} = (V_{i1} + V_{i2})/2$$

$$V_{od} = V_{o1} - V_{o2} \qquad V_{oc} = (V_{o1} + V_{o2})/2$$

- In practice, want

$$A_{dm} \gg 1 \qquad A_{cm} \ll 1$$

- If the circuit is fully symmetrical,

$$A_{cdm} = 0 \qquad A_{dcm} = 0$$

Fully Balanced Circuit Topology

- Signal is carried in $V_{xd} = V_{x1} - V_{x2}$. Let

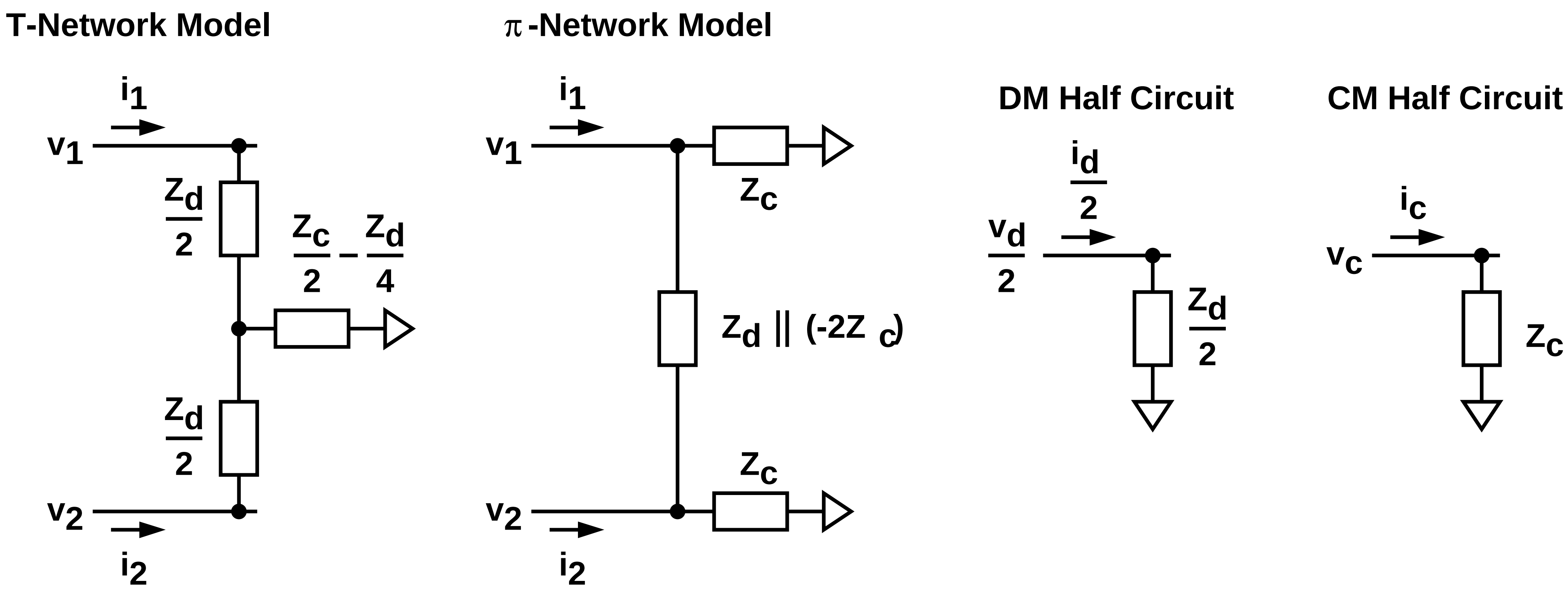
$$V_{x1} = A \sin \omega t + n_1 \qquad V_{x2} = -A \sin \omega t + n_2 \qquad V_{xd} = 2A \sin \omega t + n_1 - n_2$$

Assuming n_1 and n_2 are uncorrelated, then

$$\mathrm{SNR}_{x1} = \mathrm{SNR}_{x2} = \frac{A^2/2}{\overline{n^2}} \quad \Rightarrow \quad \mathrm{SNR}_{xd} = \frac{2A^2}{\overline{n_1^2} + \overline{n_2^2}} = \frac{2A^2}{2\overline{n^2}} = 2\mathrm{SNR}_{x1}$$

- Immune to common-mode noise, such as noises from power supplies and substrate.
- No even-order harmonic distortion in V_{xd}.
- Require additional common-mode feedback circuitry to set $V_{xc} = (V_{x1} + V_{x2})/2$.

Small-Signal Models for Differential Loading

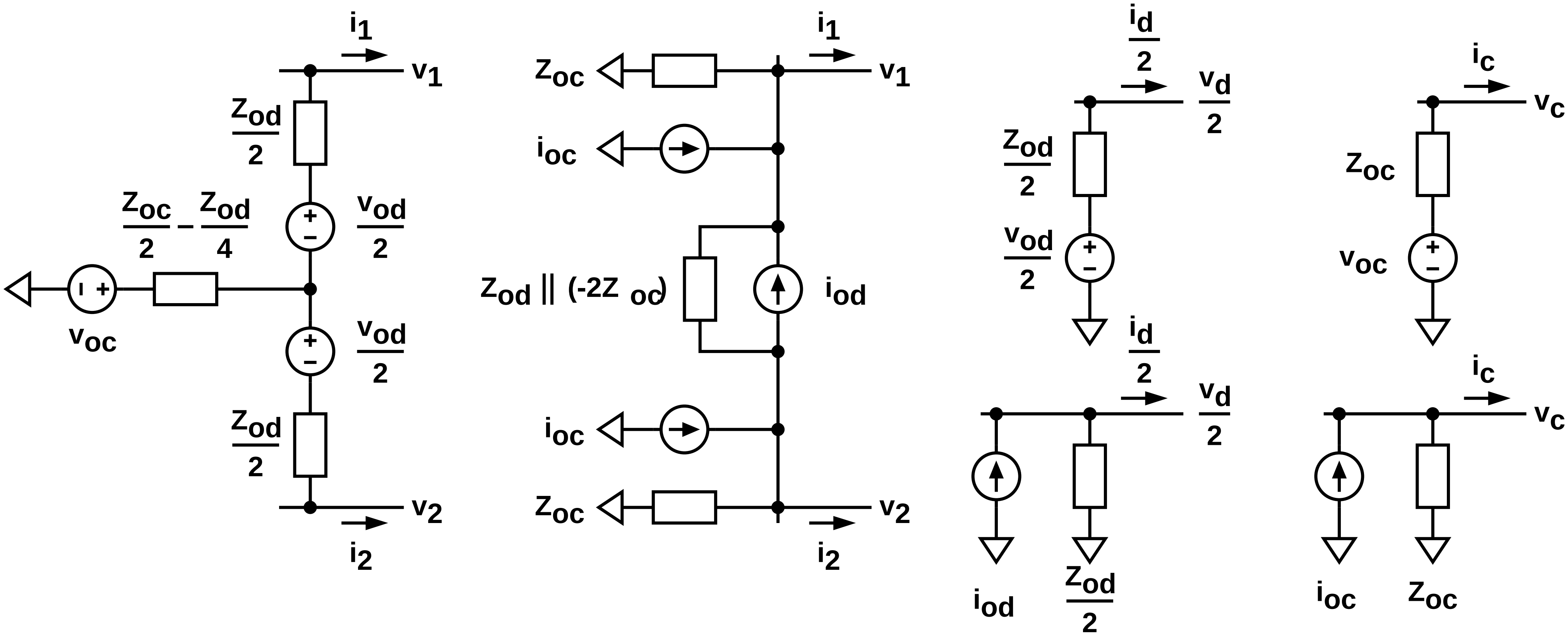


$$v_d = v_1 - v_2 \qquad v_c = (v_1 + v_2)/2 \qquad i_d = (i_1 - i_2)/2 \qquad i_c = (i_1 + i_2)/2$$

$$Z_d = \left.\frac{v_d}{i_d}\right|_{v_c=0} \qquad Z_c = \left.\frac{v_c}{i_c}\right|_{v_d=0}$$

Small-Signal Models for Differential Signal Sources

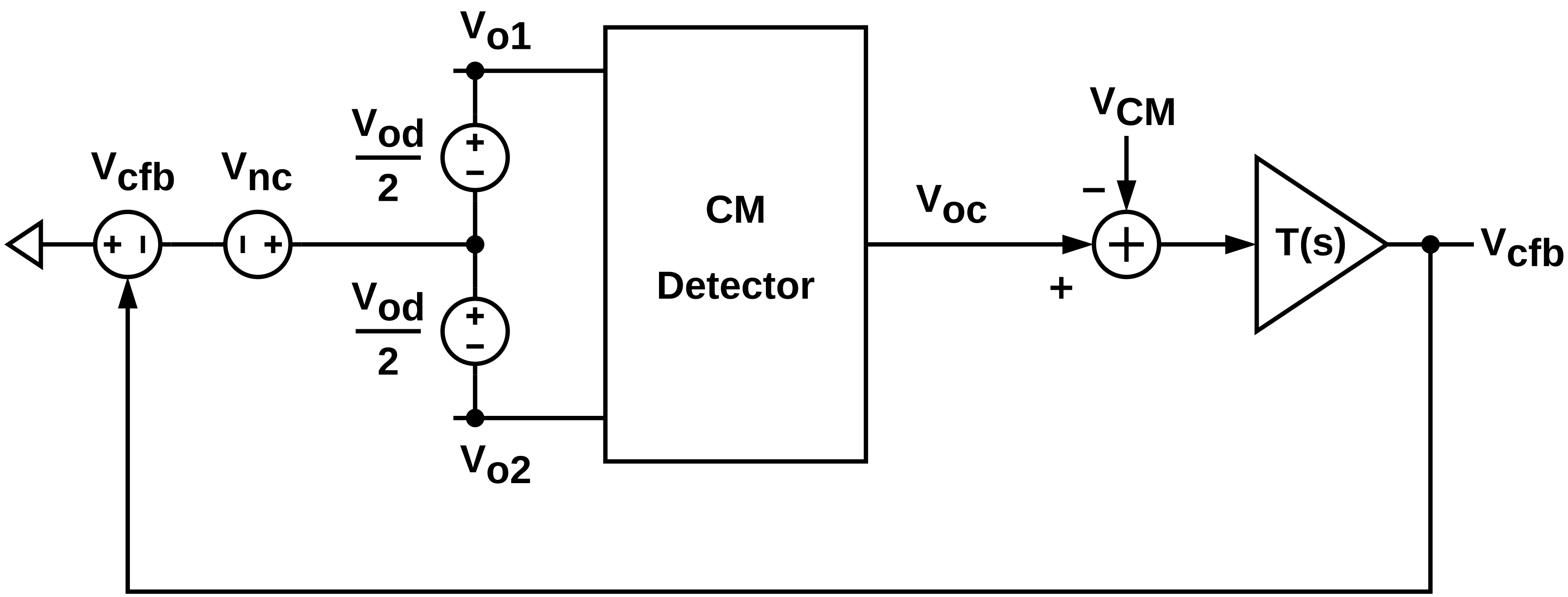
Thevenin-Network Model **Norton-Network Model** **DM Half Circuit** **CM Half Circuit**



$$v_{od} = A_{dm} v_{id} \qquad v_{oc} = A_{cm} v_{ic} \qquad i_{od} = G_{md} v_{id} \qquad i_{oc} = G_{mc} v_{ic}$$

$$A_{dm} = \left.\frac{v_{od}}{v_{id}}\right|_{i_{od}=0} \qquad A_{cm} = \left.\frac{v_{oc}}{v_{ic}}\right|_{i_{oc}=0} \qquad G_{md} = \left.\frac{i_{od}}{v_{id}}\right|_{v_{od}=0} \qquad G_{mc} = \left.\frac{i_{oc}}{v_{ic}}\right|_{v_{oc}=0}$$

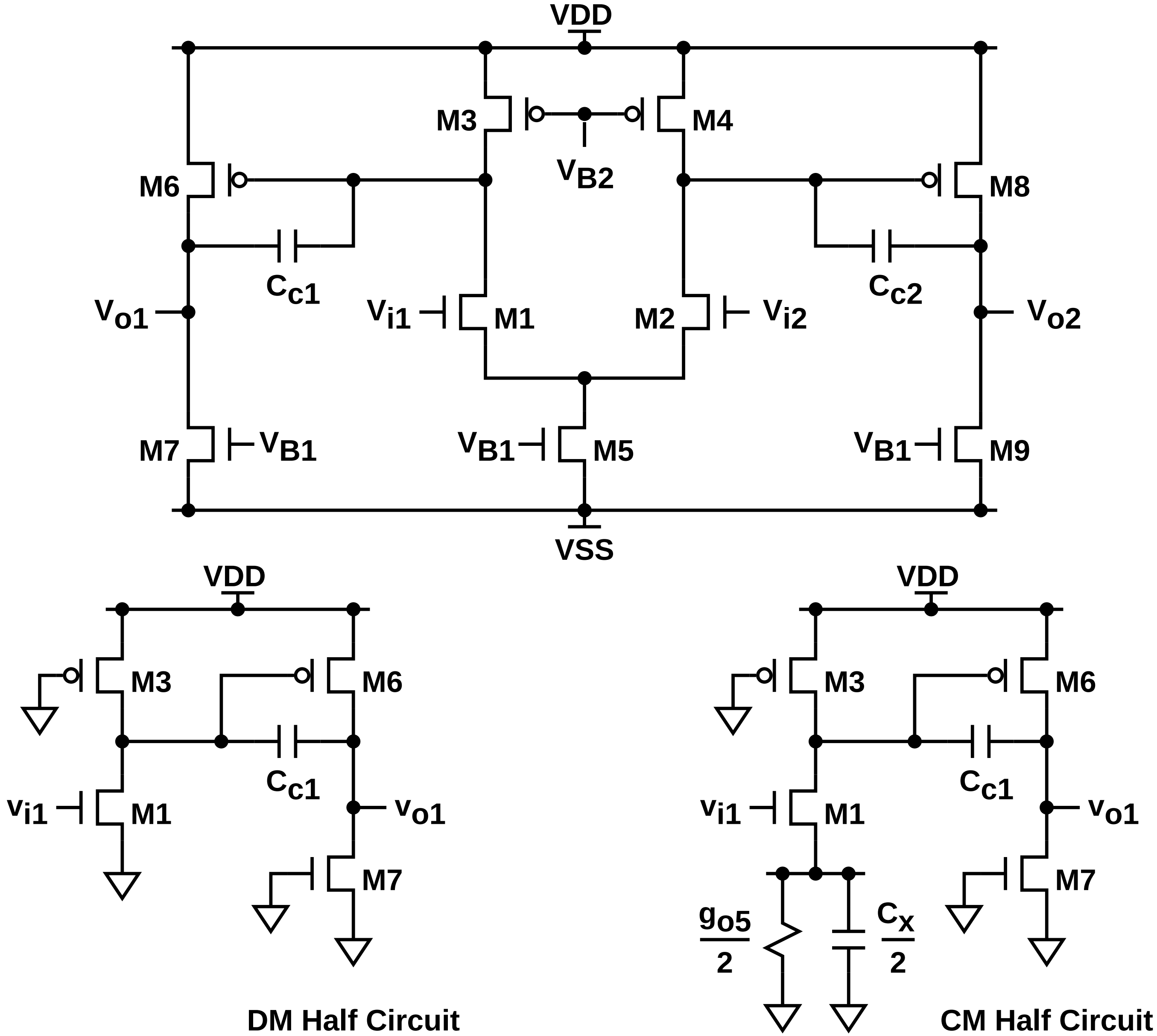
Common-Mode Feedback (CMFB)



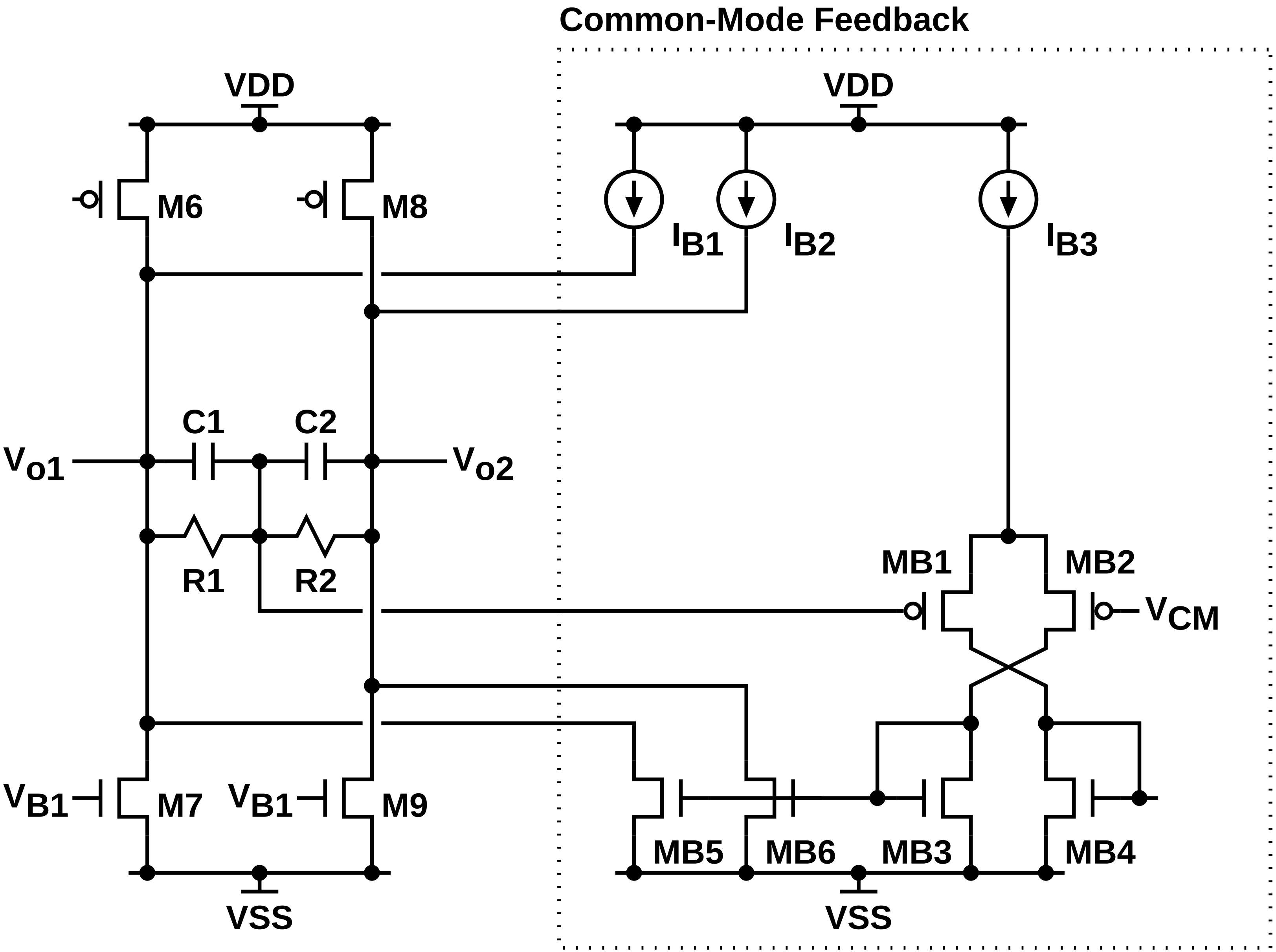
$$V_{cfb} = (V_{oc} - V_{CM}) \cdot T(s) \quad V_{oc} = V_{nc} - V_{cfb} \quad \Rightarrow \quad V_{oc} = \frac{T}{1+T} \times V_{CM} + \frac{1}{1+T} \times V_{nc}$$

- Want large CMFB loop gain, T, to stabilize V_{oc}.
- May want large ω_t of T to suppress high-frequency components in V_{nc}.
- Must check the frequency stability of $1/[1 + T(s)]$.

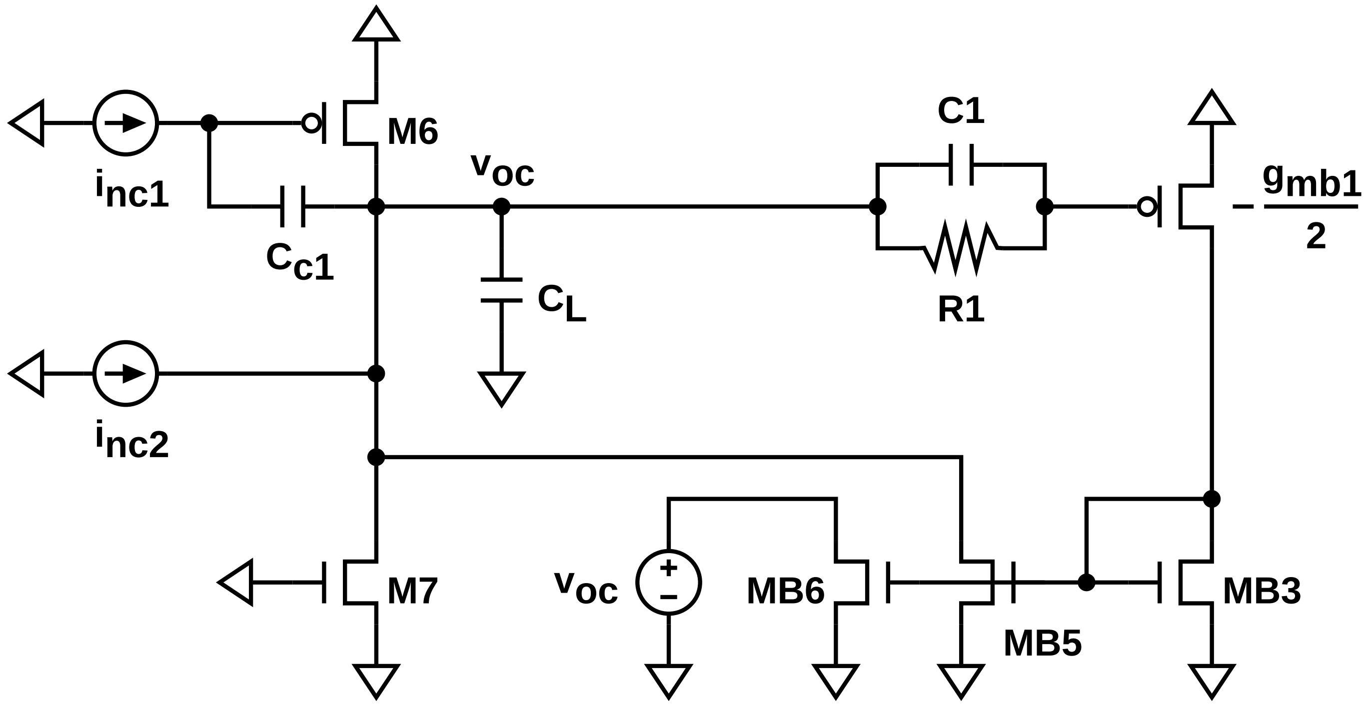
A Fully Differential Two-Stage Operational Amplifier



CMFB Using Resistive Divider and Error Amplifier

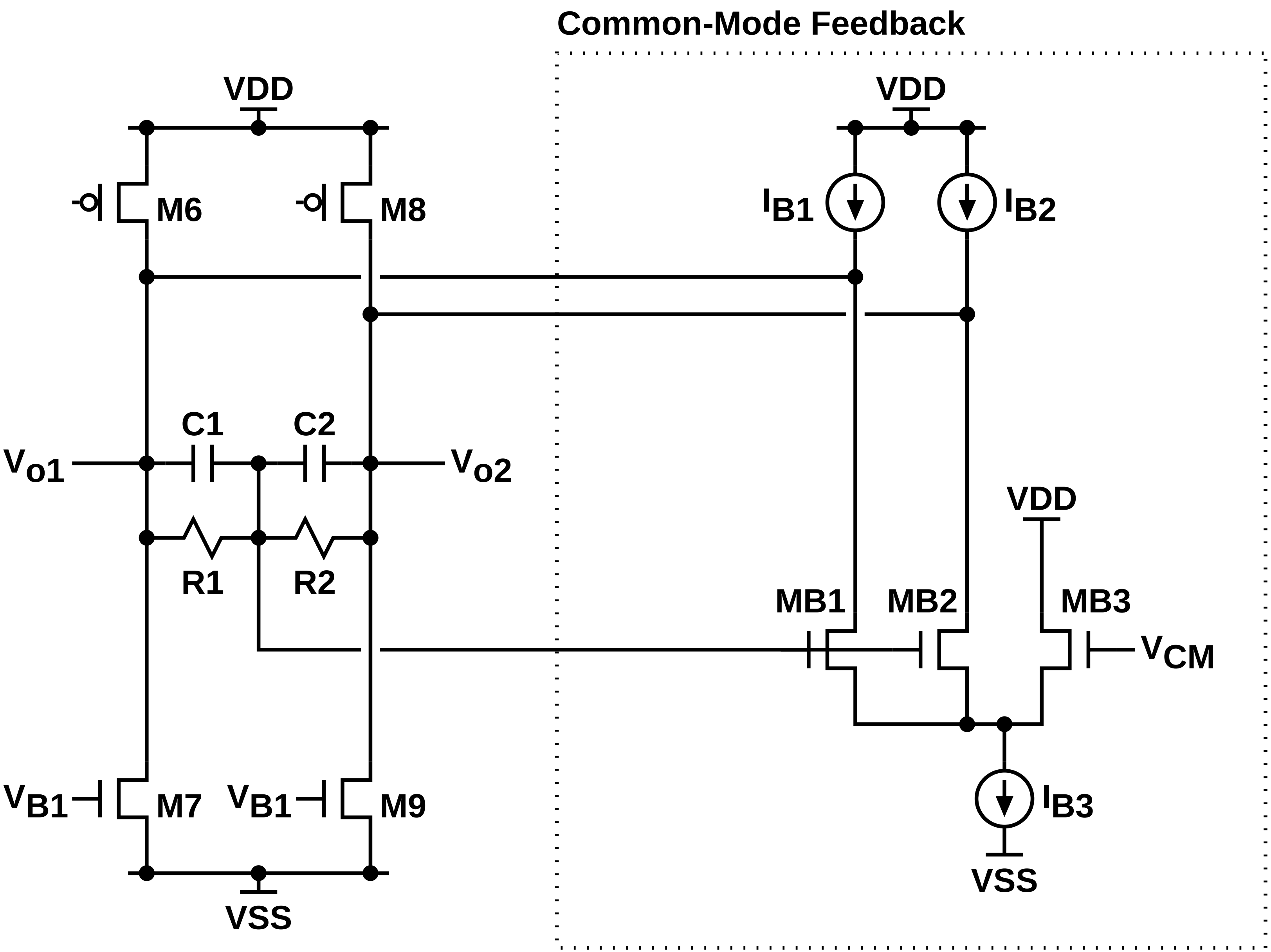


CMFB Using Resistive Divider and Error Amplifier

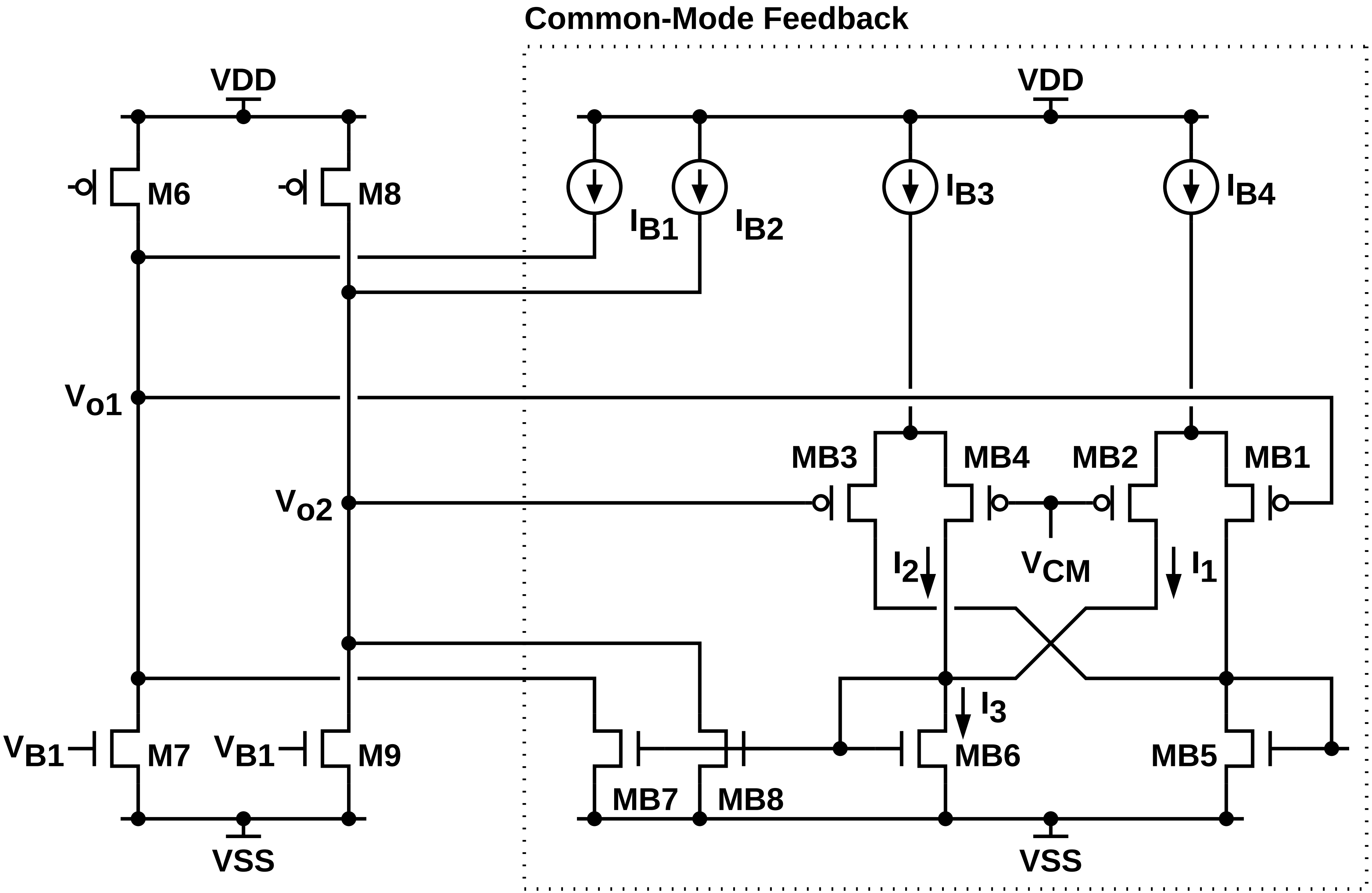


- The loop gain $|T| \approx g_{mb5}(r_{o6} \parallel r_{o7}) \cdot g_{mb1}/(2g_{mb3})$.
- C_1 and C_2 are used to improve high-frequency response.
- The resistive loading of R_1 and R_2 can degrade A_{dm}. Voltage buffers can be added between the opamp's outputs and the resistive divider.

CMFB Using Resistive Divider and Direct Current Injection



CMFB Using Dual Differential Pairs



CMFB Using Dual Differential Pairs

For the MB1-MB2 and MB3-MB4 source-coupled pairs,

$$I_{BB} = I_{B3} = I_{B4} = 2 \times \frac{k}{2} \cdot V_{ov}^2 \qquad k = k' \left(\frac{W}{L}\right)$$

$$I_{dd} = \frac{k}{2} V_{id} \sqrt{4\frac{I_{BB}}{k} - V_{id}^2} \qquad I_{d1} = \frac{I_{BB}}{2} + \frac{I_{dd}}{2} \qquad I_{d2} = \frac{I_{BB}}{2} - \frac{I_{dd}}{2}$$

$$\begin{aligned}
I_1 &= \frac{I_{BB}}{2} + \frac{k}{4}(V_{oc} + V_{od}/2 - V_{CM})\sqrt{4\frac{I_{BB}}{k} - (V_{oc} + V_{od}/2 - V_{CM})^2} \\
&\approx \frac{I_{BB}}{2} + \frac{k}{4}(V_{oc} - V_{CM} + V_{od}/2)\sqrt{4V_{ov}^2 - (V_{od}/2)^2 - (V_{oc} - V_{CM})V_{od}} \\
&\approx \frac{I_{BB}}{2} + \frac{k}{4}(V_{oc} - V_{CM} + V_{od}/2)\sqrt{4V_{ov}^2 - (V_{od}/2)^2} \\
&\quad \times \left\{1 - \frac{1}{2}\left[\frac{(V_{oc} - V_{CM})V_{od}}{4V_{ov}^2 - (V_{od}/2)^2}\right] - \frac{1}{8}\left[\frac{(V_{oc} - V_{CM})V_{od}}{4V_{ov}^2 - (V_{od}/2)^2}\right]^2 + \cdots\right\}
\end{aligned}$$

CMFB Using Dual Differential Pairs

$$I_2 \approx \frac{I_{BB}}{2} + \frac{k}{4}(V_{oc} - V_{CM} - V_{od}/2)\sqrt{4V_{ov}^2 - (V_{od}/2)^2}$$

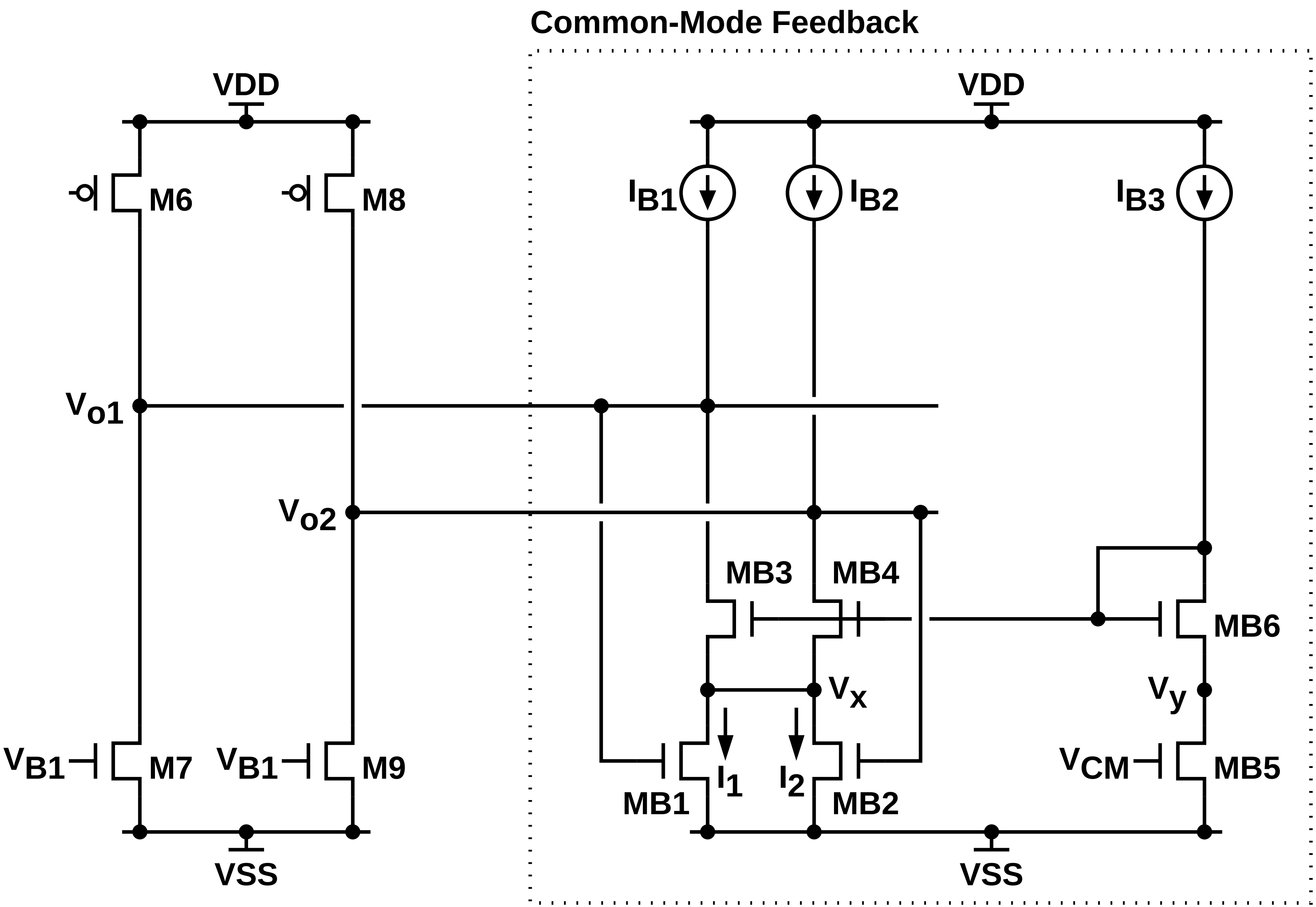
$$\times \left\{ 1 + \frac{1}{2}\left[\frac{(V_{oc} - V_{CM})V_{od}}{4V_{ov}^2 - (V_{od}/2)^2}\right] - \frac{1}{8}\left[\frac{(V_{oc} - V_{CM})V_{od}}{4V_{ov}^2 - (V_{od}/2)^2}\right]^2 + \cdots \right\}$$

$$I_3 = I_1 + I_2 \approx I_{BB} + \frac{k}{2}(V_{oc} - V_{CM})\sqrt{4V_{ov}^2 - (V_{od}/2)^2}$$

$$\times \left\{ 1 - \frac{1}{4}\left[\frac{V_{od}^2}{4V_{ov}^2 - (V_{od}/2)^2}\right] - \frac{1}{8}\left[\frac{(V_{oc} - V_{CM})V_{od}}{4V_{ov}^2 - (V_{od}/2)^2}\right]^2 + \cdots \right\}$$

- The input devices, MB1–MB4, must remain in the forward-active region over the voltage range of V_{o1} and V_{o2}.
- The variation in V_{od} can produce an ac component in I_3 as well as V_{oc}.
- If $V_{oc} = V_{CM}$, I_1 and I_2 are nonlinear functions of V_{od}, but $I_3 = I_1 + I_2$ is a constant.

CMFB Using Transistors in the Triode Region



CMFB Using Transistors in the Triode Region

MB1, MB2, and MB5 are in the triode region. Let $k_{B1} = k_{B2} = k_{B5} = k$,

$$I_1 = k\left(V_{o1} - V_{tn} - \frac{1}{2}V_x\right)V_x \quad I_2 = k\left(V_{o2} - V_{tn} - \frac{1}{2}V_x\right)V_x \quad I_{B3} = k\left(V_{CM} - V_{tn} - \frac{1}{2}V_y\right)V_y$$

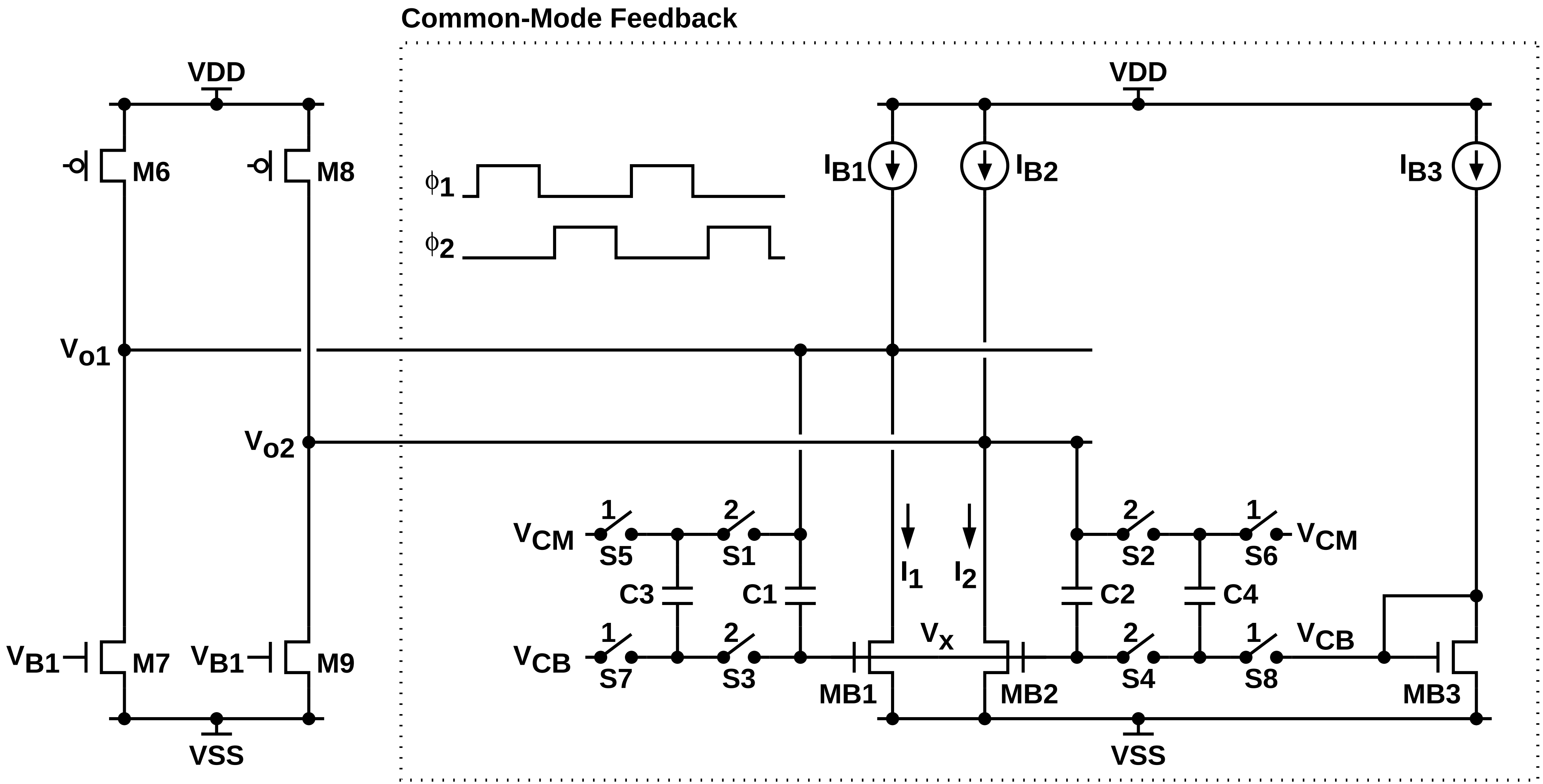
$$\Rightarrow \quad I_1 + I_2 = 2k\left(V_{oc} - V_{tn} - \frac{1}{2}V_x\right)V_x \qquad V_x \approx V_y = \frac{I_{B3}}{k\left(V_{CM} - V_{tn} - \frac{1}{2}V_y\right)}$$

$$I_1 + I_2 = 2I_{B3} \cdot \frac{V_{oc} - V_{tn} - \frac{1}{2}V_x}{V_{CM} - V_{tn} - \frac{1}{2}V_y} = 2I_{B3}\left(1 + \frac{V_{oc} - V_{CM}}{V_{CM} - V_{tn} - \frac{1}{2}V_y}\right)$$

- Output swing is reduced, since it is required that $V_{o1,o2} > V_{tn} + V_x$.

- MB1 and MB2 are in the triode region, their effective g_m can be small, thus degrading loop gain and bandwidth of the CMFB.

Switched-Capacitor CMFB



$$V_{oc} - V_x = V_{CM} - V_{CB} \quad \Rightarrow \quad V_{oc} \approx V_{CM}$$

Switched-Capacitor CMFB

- The opamp operates in two different modes. It is in the *normal* mode when ϕ_2 is low.
- Assuming ΔQ charges are injected into C_3 and C_4 when ϕ_1 switches are turned off,

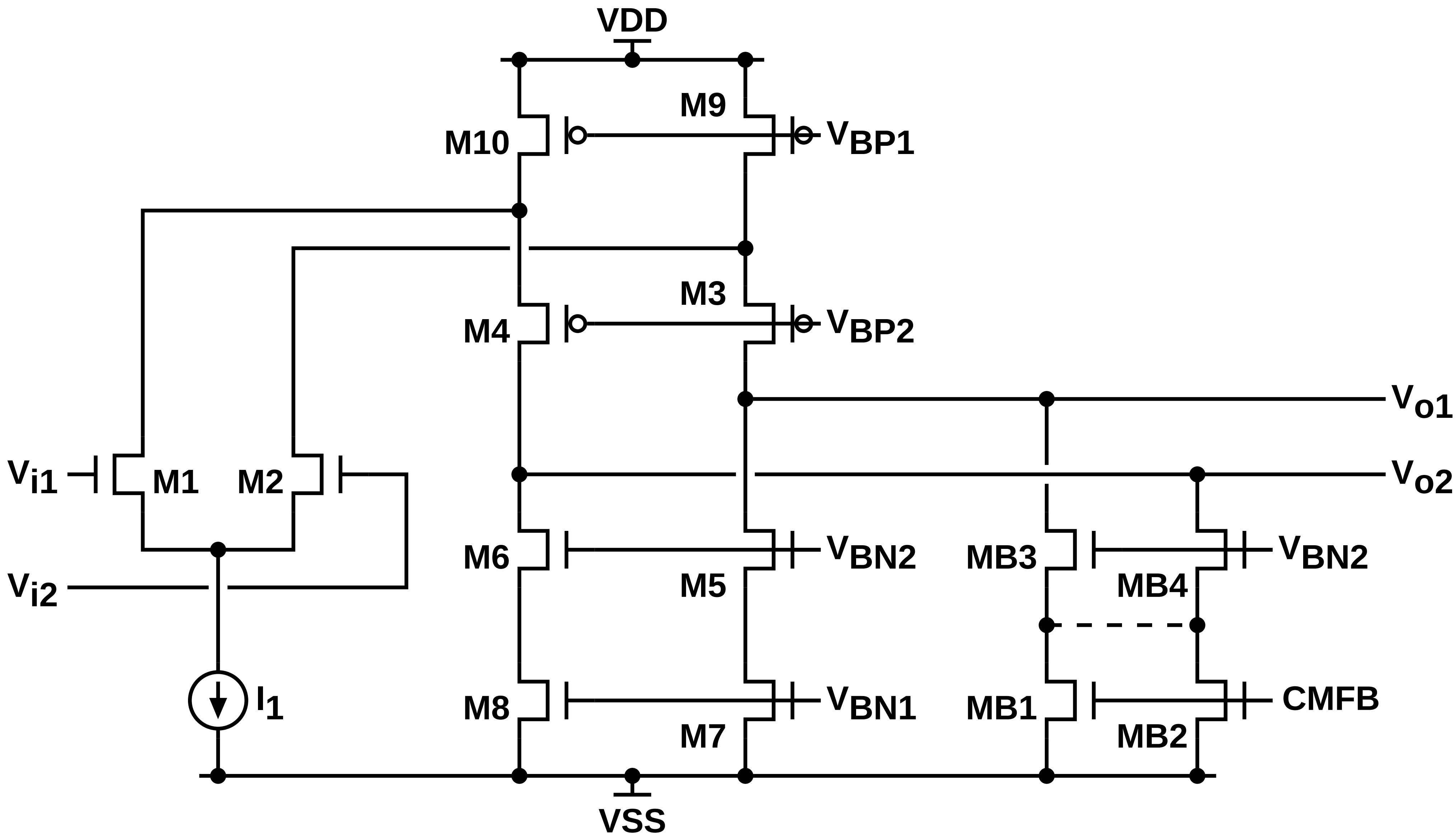
$$V_{oc} - V_x = V_{CM} - V_{CB} + \frac{\Delta Q}{C_3} \qquad \Rightarrow \qquad V_{oc} \approx V_{CM} + \frac{\Delta Q}{C_3}$$

- The loop gain of the CMFB is approximately

$$|T| \approx \frac{C_1}{C_1 + C_{gs,B1}} \times g_{m,B1} \cdot R_{o1}$$

- C_1 and C_2 add differential-mode capacitive loading to the outputs.
- The additional common-mode capacitive loading is $(C_1 + C_2) \parallel (C_{gs,B1} + C_{gs,B2})$.
- The value of $C_{3,4}$ may be between 1/4–1/10 of $C_{1,2}$ for low-pass filter function.

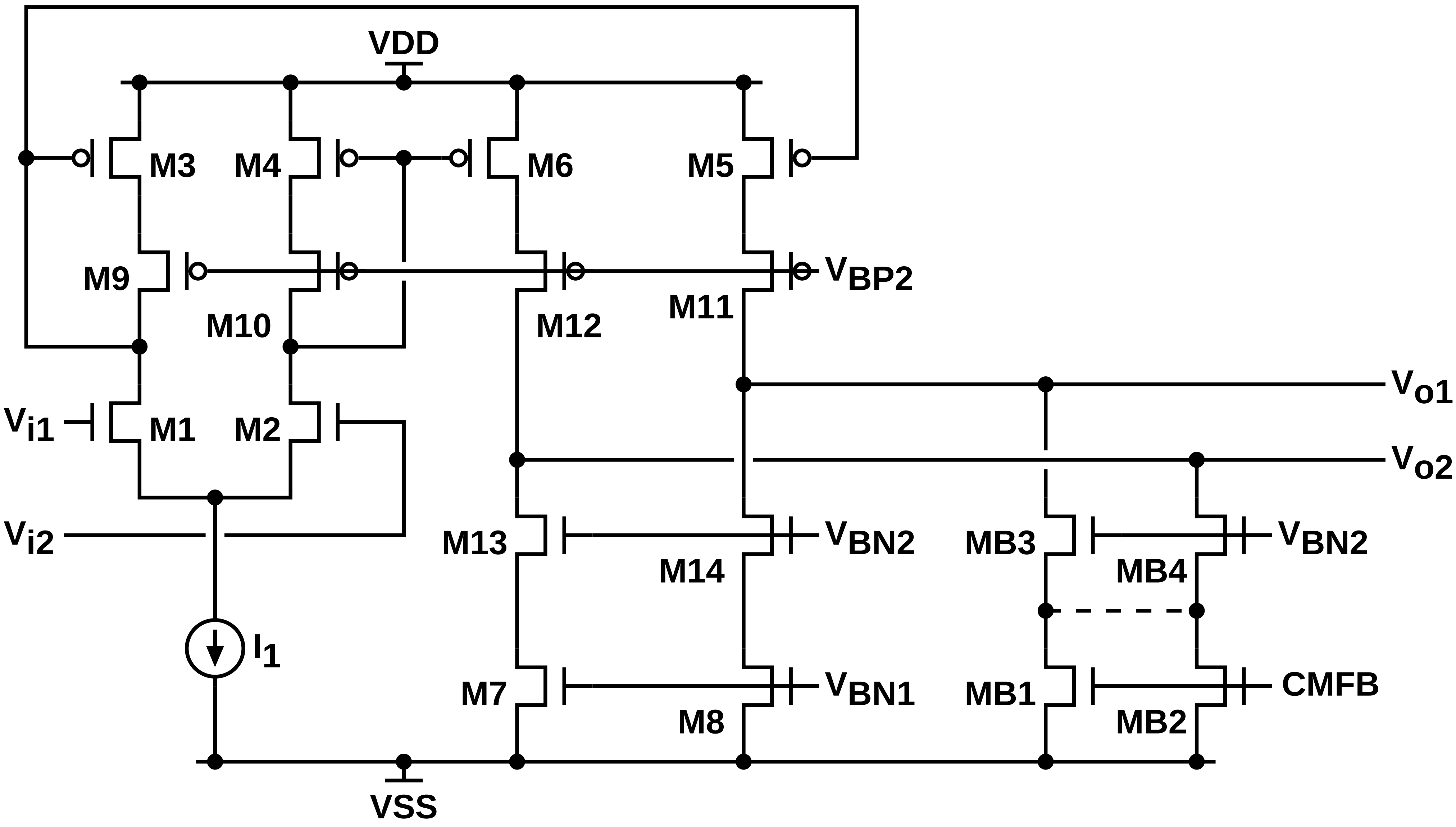
Folded-Cascode Operational Amplifier



Folded-Cascode Operational Amplifier

- Frequency compensation is provided by the capacitive loads at the outputs.
- Non-dominant poles are determined by M3 and M4, and $\approx \omega_{t3}$ (ω_{t4}).
- It is not uncommon that $I_{D1,D2} \simeq I_{D3,D4}$.
- For high-speed designs, use pMOST input stage. The resulting opamps has higher non-dominant poles.
- Active cascode configuration can be applied to M3, M4, M5, and M6.

Current-Mirror Operational Amplifier



Current-Mirror Operational Amplifier

The M3-M5 and M4-M6 current mirrors have a current gain of K.

$$\left(\frac{W}{L}\right)_3 = \left(\frac{W}{L}\right)_4 = \frac{1}{K}\left(\frac{W}{L}\right)_5 = \frac{1}{K}\left(\frac{W}{L}\right)_6$$

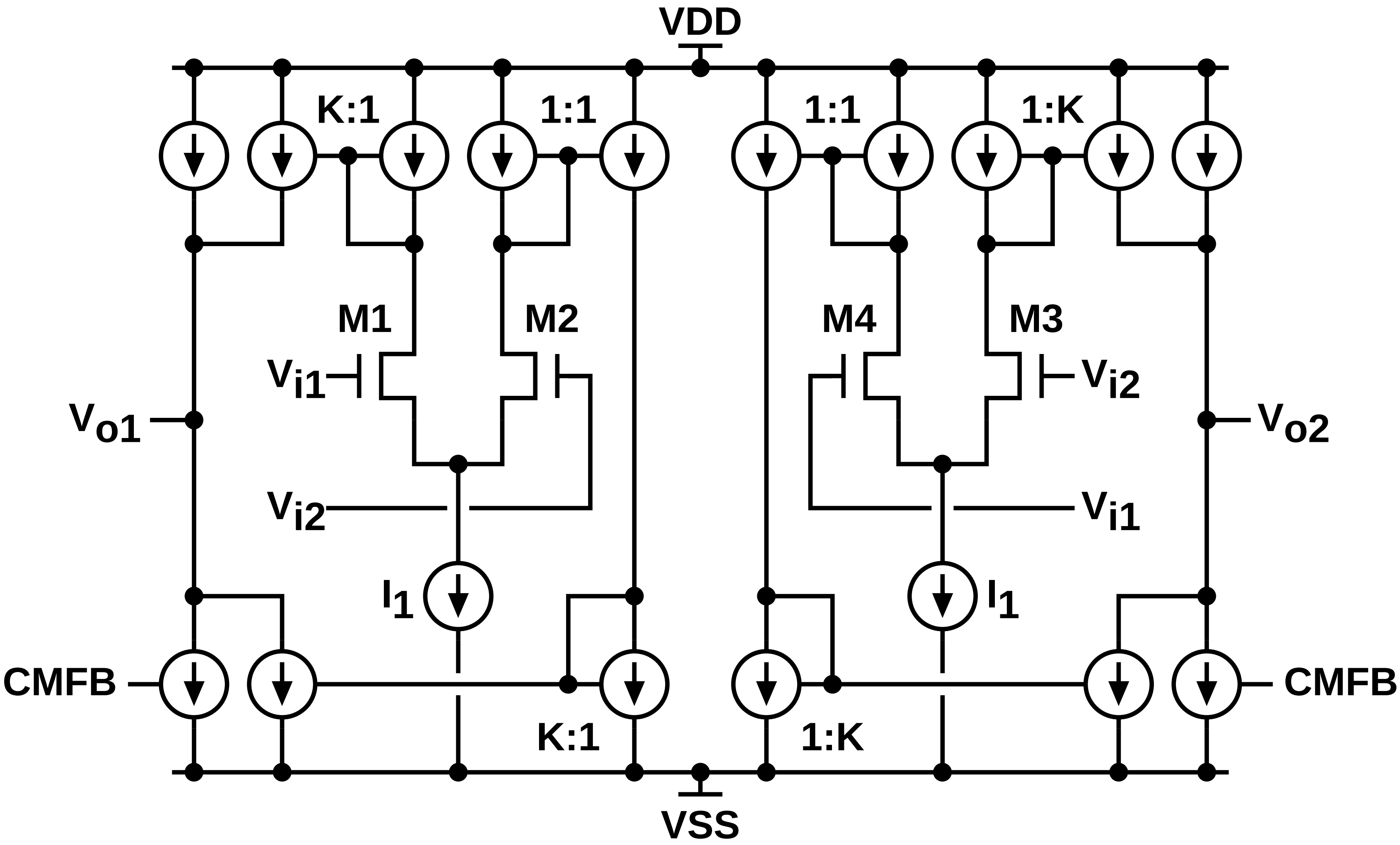
$$I_{D1} = I_{D2} = I_{D3} = I_{D4} = \frac{1}{K} I_{D5} = \frac{1}{K} I_{D6} = \frac{1}{2} I_1$$

- The single-ended maximum output current for slewing is

$$I_{o(max)} = \frac{K}{2} I_1$$

- For a general-purpose fully differential opamp, may use large pMOST input stage, K=2, and wide-swing enhanced output-impedance cascode current mirrors.

Current-Mirror Push-Pull Operational Amplifier



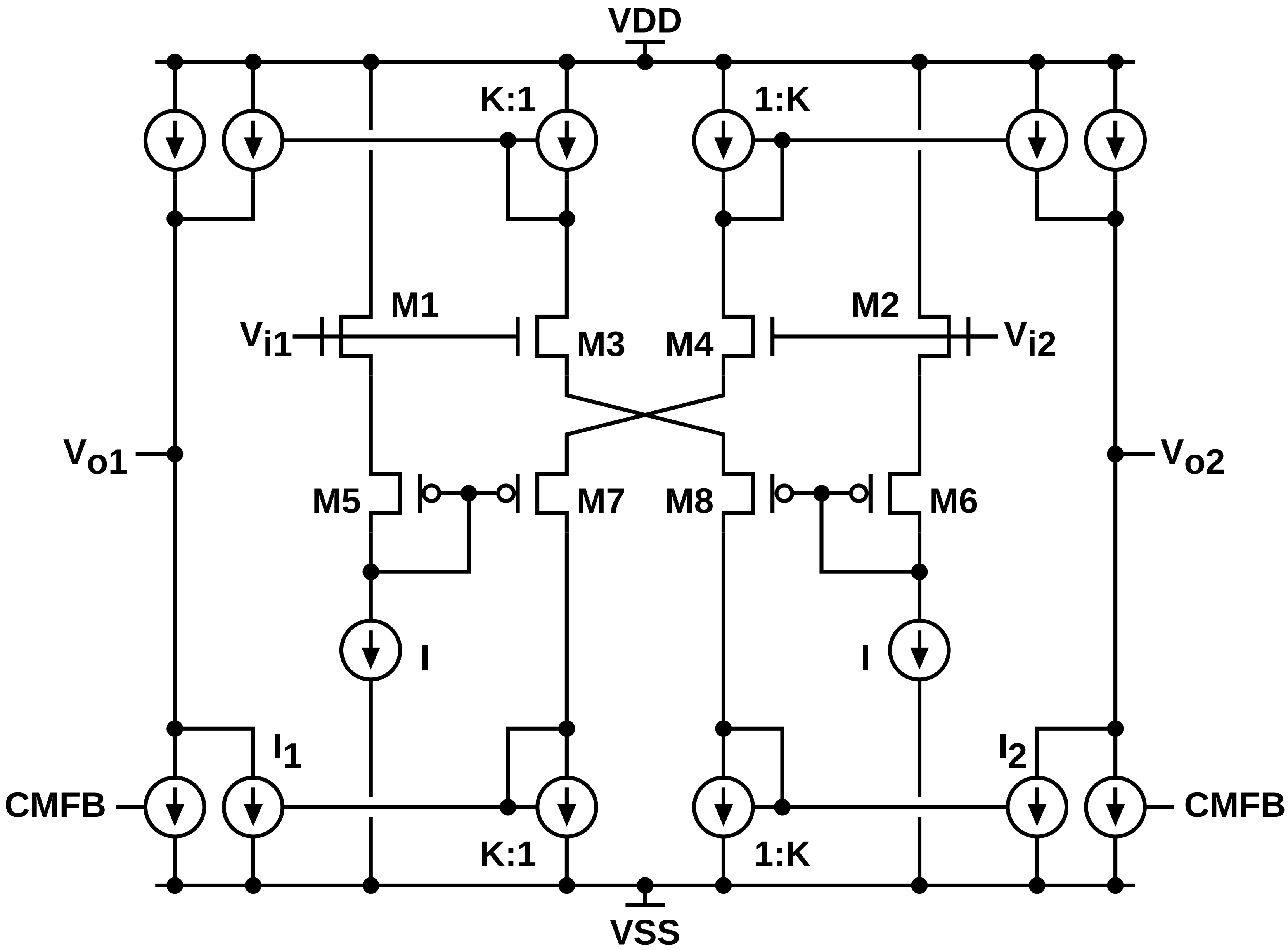
Current-Mirror Push-Pull Operational Amplifier

- The single-ended maximum output current for slewing is

$$I_{o(max)} = K I_1$$

- The small-signal response is slower due to additional signal paths.

Class-AB Operational Amplifier



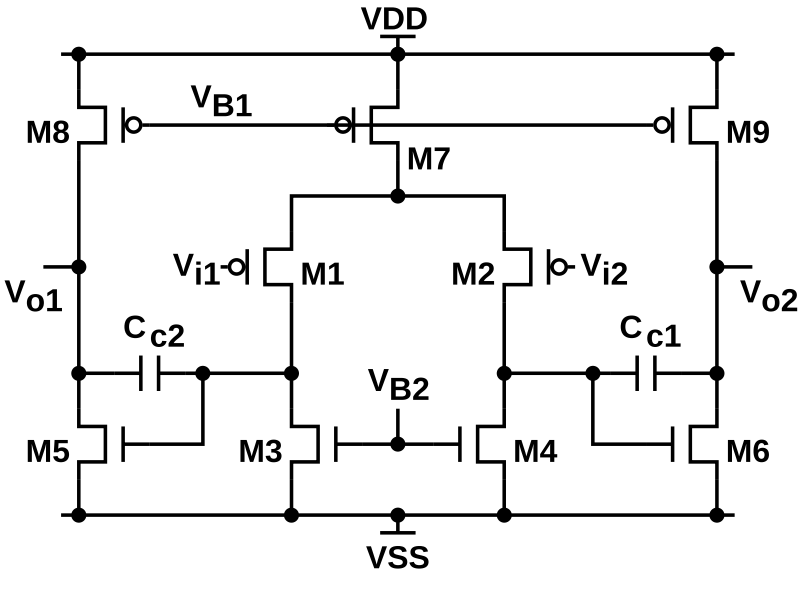
Class-AB Operational Amplifier

If nMOSTs M1–M4 are identical, and pMOSTs M5–M8 are identical, and all current mirrors have a current gain of K, then the bias currents are

$$I_{D1} = I_{D2} = I_{D3} = I_{D4} = \frac{1}{K} I_1 = \frac{1}{K} I_2 = I$$

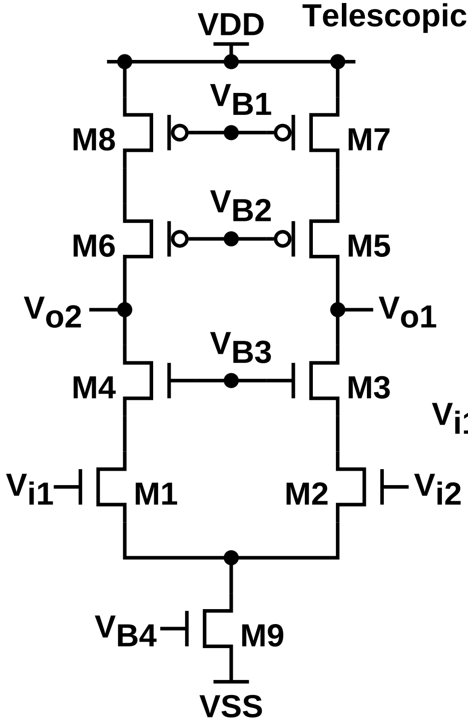
- Low quiescent power and large slew rate.
- The input level shifter increases the noise and offset, and adds additional poles.
- Not suitable for low-voltage operation.

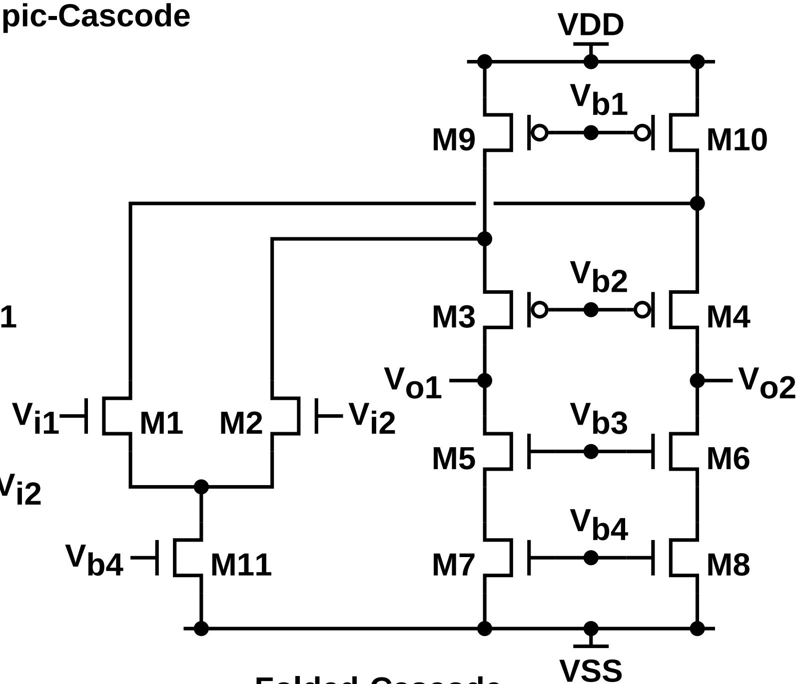
Fully Differential Operational Amplifiers



Two-Stage

Telescopic-Cascode





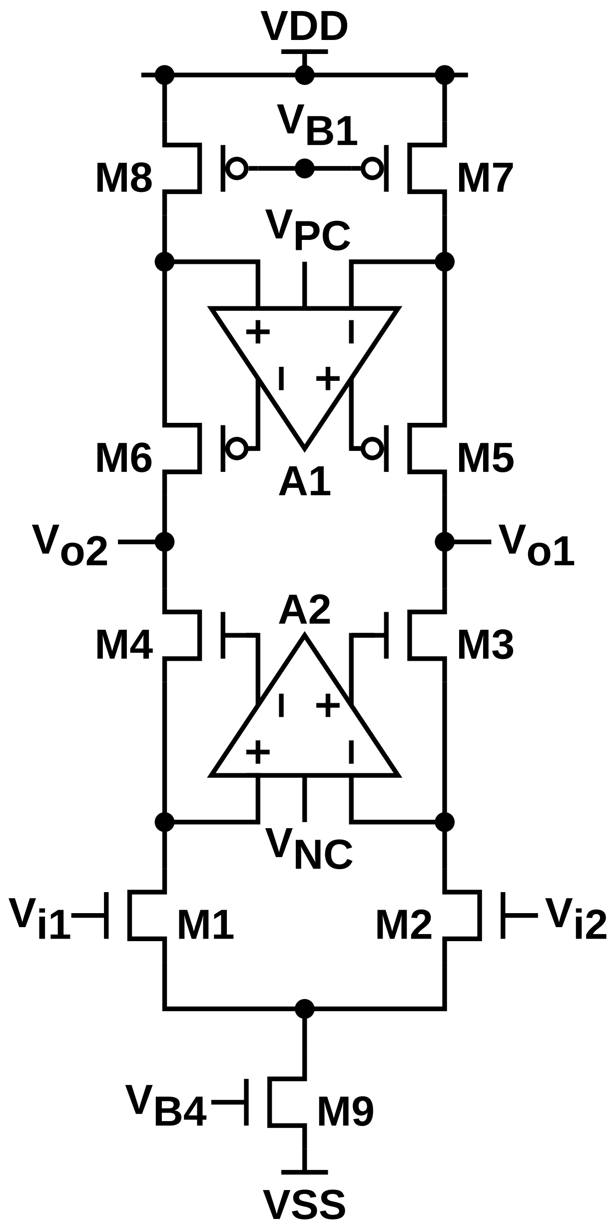
Folded-Cascode

$$\Delta V_o(\text{Two Stage}) = V_{DD} - 2V_{DSAT} \qquad \Delta V_o(\text{Telescopic}) = V_{DD} - 5V_{DSAT} - 3V_{margin}$$

$$\Delta V_o(\text{Folded-Cascode}) = V_{DD} - 4V_{DSAT} - 2V_{margin}$$

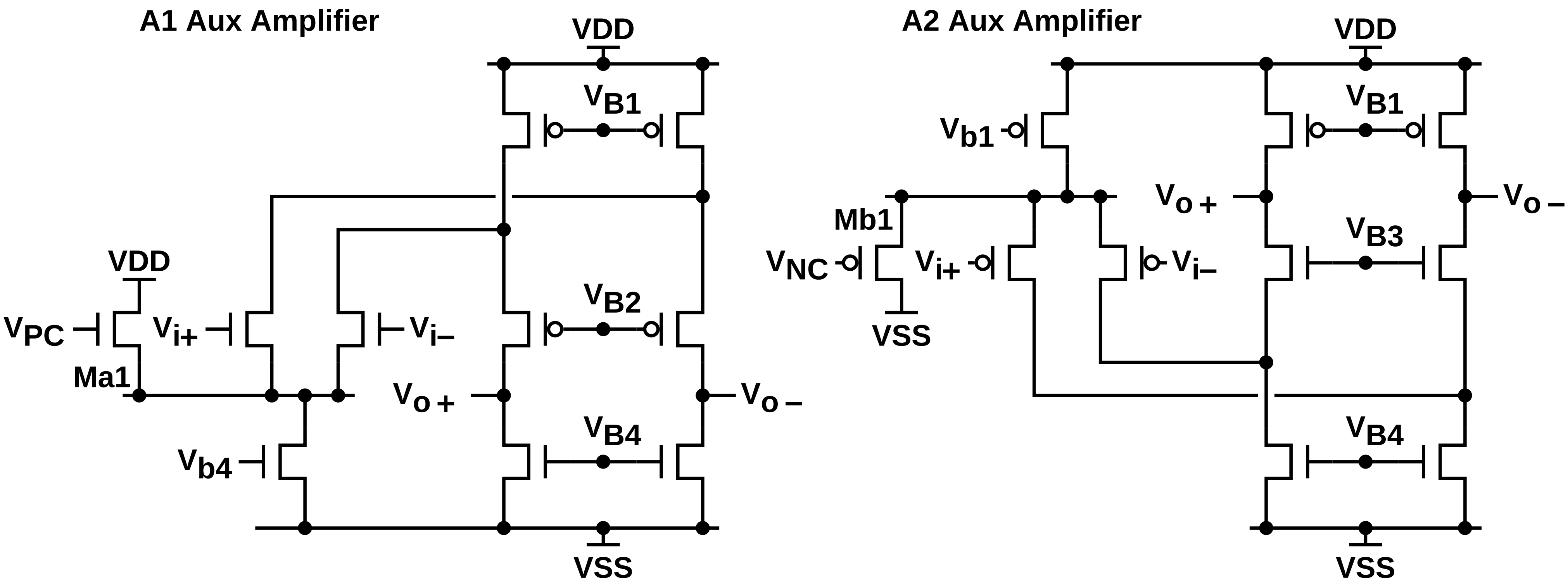
$$\text{SNR} \cdot \frac{\text{Speed}}{\text{Power}} \propto \frac{\Delta V_o^2}{kT/C} \cdot \frac{g_m/C}{V_{DD} \cdot I} \propto \frac{\Delta V_o^2}{V_{DD}}$$

Active-Cascode Telescopic Operational Amplifier



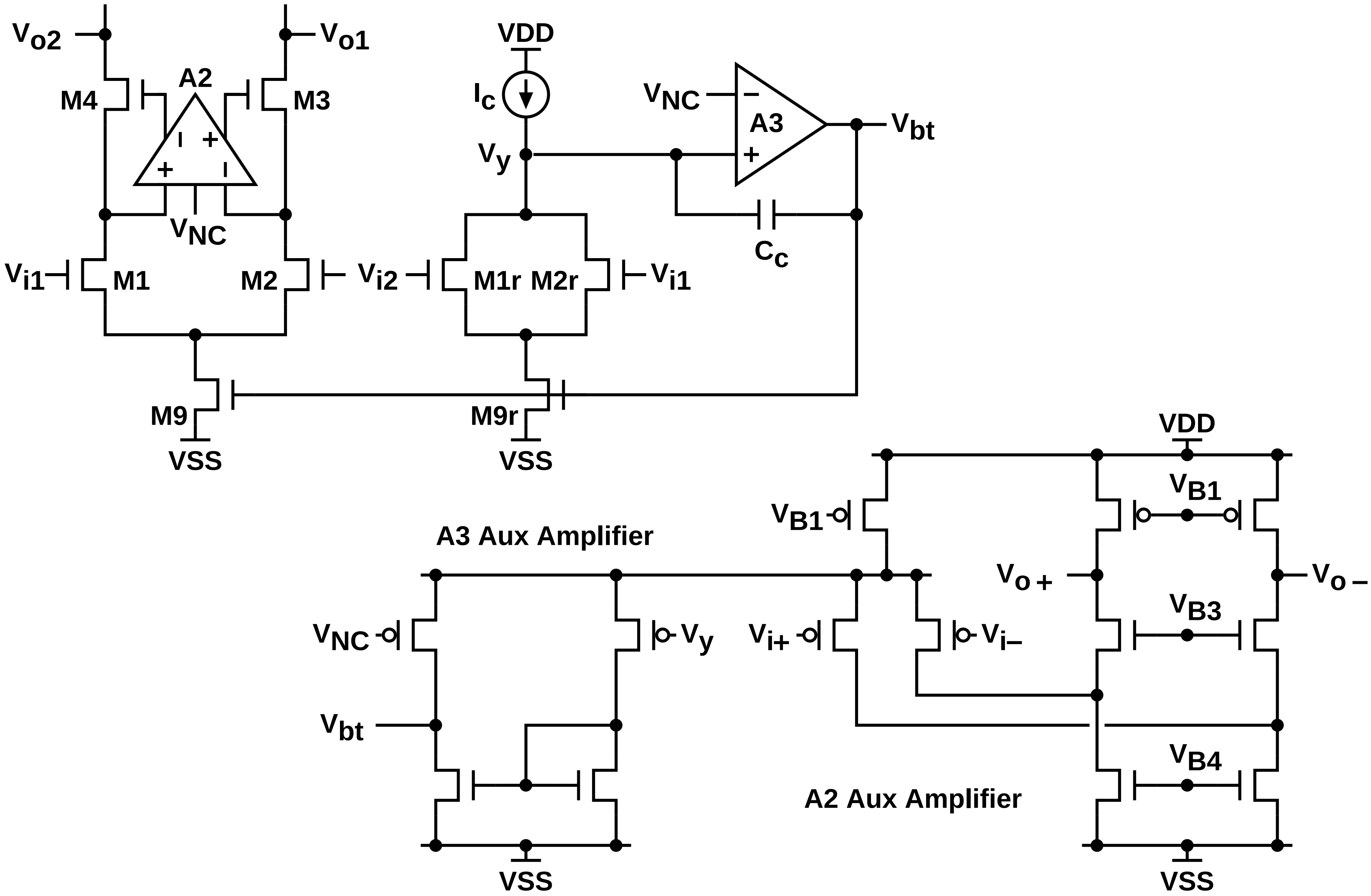
- Have the best speed/power ratio.
- A1 and A2 auxiliary amplifiers are used to increase output impedance and the dc voltage gain, $A_v(0)$.
- Explicit compensation capacitors may be required at the outputs of A1 and A2.
- To increase ΔV_o, M7, M8, and M9, can be biased in the triode region. However, $A_v(0)$ is reduced due to the reduced r_o of M7 and M8. Also, CMRR and PSRR are degraded due to the reduced r_o of M9.
- Reference: Gulati and Lee, JSSC 12/98, pp. 2010–2019.

Fully Differential Gain-Enhancement Auxiliary Amplifiers



- $V_{S3} \approx V_{S4} \approx V_{NC}$, due to the CMFB of M3, M4, and A2.
- $V_{S5} \approx V_{S6} \approx V_{PC}$, due to the CMFB of M5, M6, and A1.

Replica-Tail Feedback



Replica-Tail Feedback

- The feedback loop increase M9's output resistance, R_{o9}, i.e.,

$$R_{o9} = r_{o9}\left[1 + A_3 \cdot (g_{m9r} r_{o9r})(g_{m1r} r_{o1r})\right] = r_{o9}\left[1 + A_{loop}\right]$$

- It can be shown the effective common-mode transconductance of M1-M2-M9 is

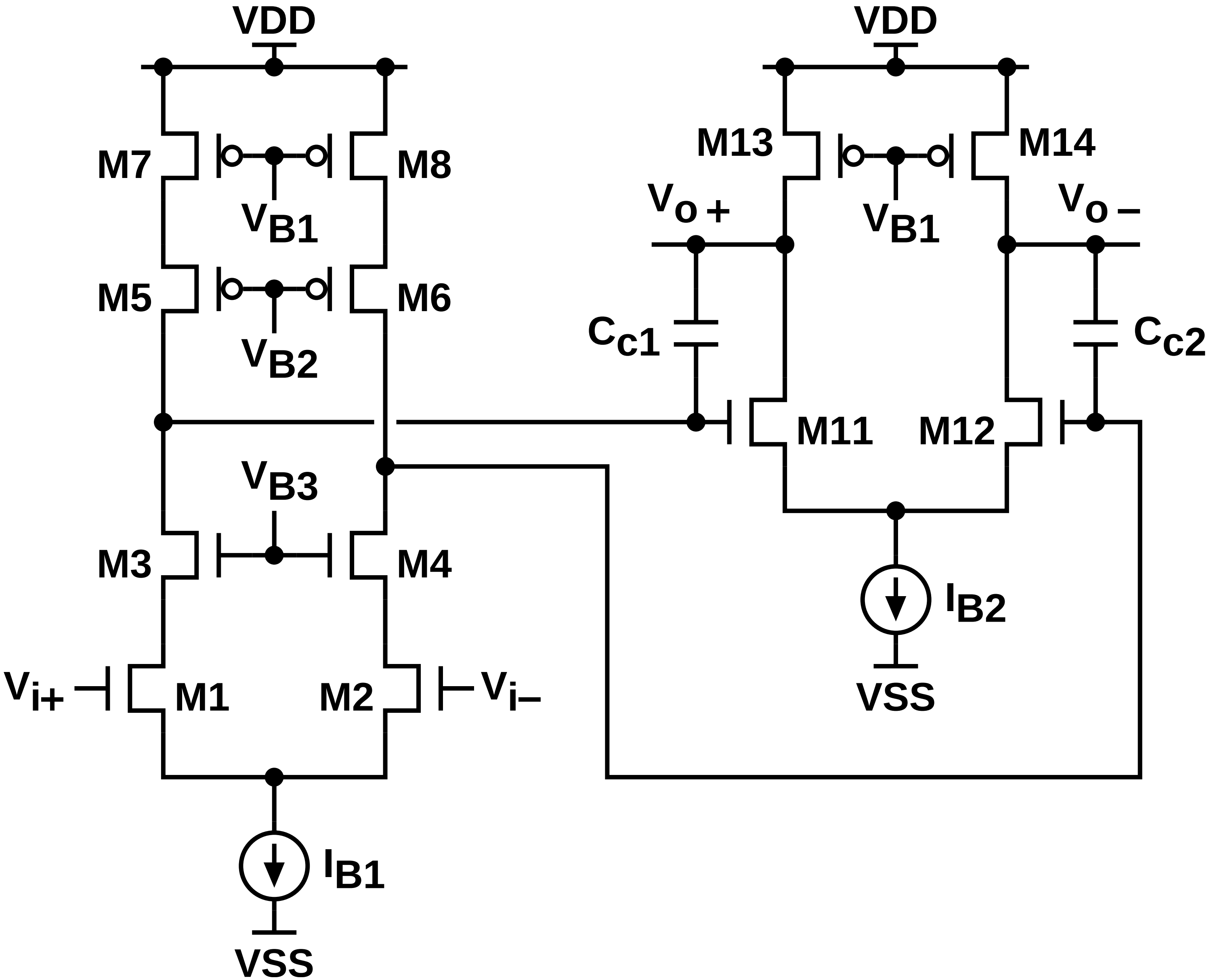
$$G_e = G_m \times \frac{1 + A_{loop} \cdot M}{1 + A_{loop}} \qquad M = 1 - \frac{g_{m9}}{g_{m9r}} \cdot \frac{G_{mr}}{G_m}$$

$$G_m = \frac{g_m}{1 + g_m r_{o9}} \qquad g_m = g_{m1} + g_{m2}$$

$$G_{mr} = \frac{g_{mr}}{1 + g_{mr} r_{o9r}} \qquad g_{mr} = g_{m1r} = g_{m2r}$$

- The mismatch M and the bandwidth of the feedback loop limit the enhancement effect.

High-Gain Two-Stage Operational Amplifier



High-Gain Two-Stage Operational Amplifier

- DC gain $A_0 \approx (g_{m1} r_{o1})^2 (g_{m11} r_{o11})/4$.
- Maximum single-rail output swing is $(V_{DD} - V_{SS}) - 3V_{DSAT}$.
- Standard two-stage frequency compensation.

$$\omega_u = A_0 \times \omega_1 = \frac{g_{m1}}{C_c} \qquad \omega_2 = \frac{g_{m11}}{C_L}$$

  Need to avoid the feedforward zero using additional circuitry.

- Both stages need common-mode feedback.
- May have large input Miller capacitance due to high dc gain of the first stage.