Operational Amplifiers with Single-Ended Outputs

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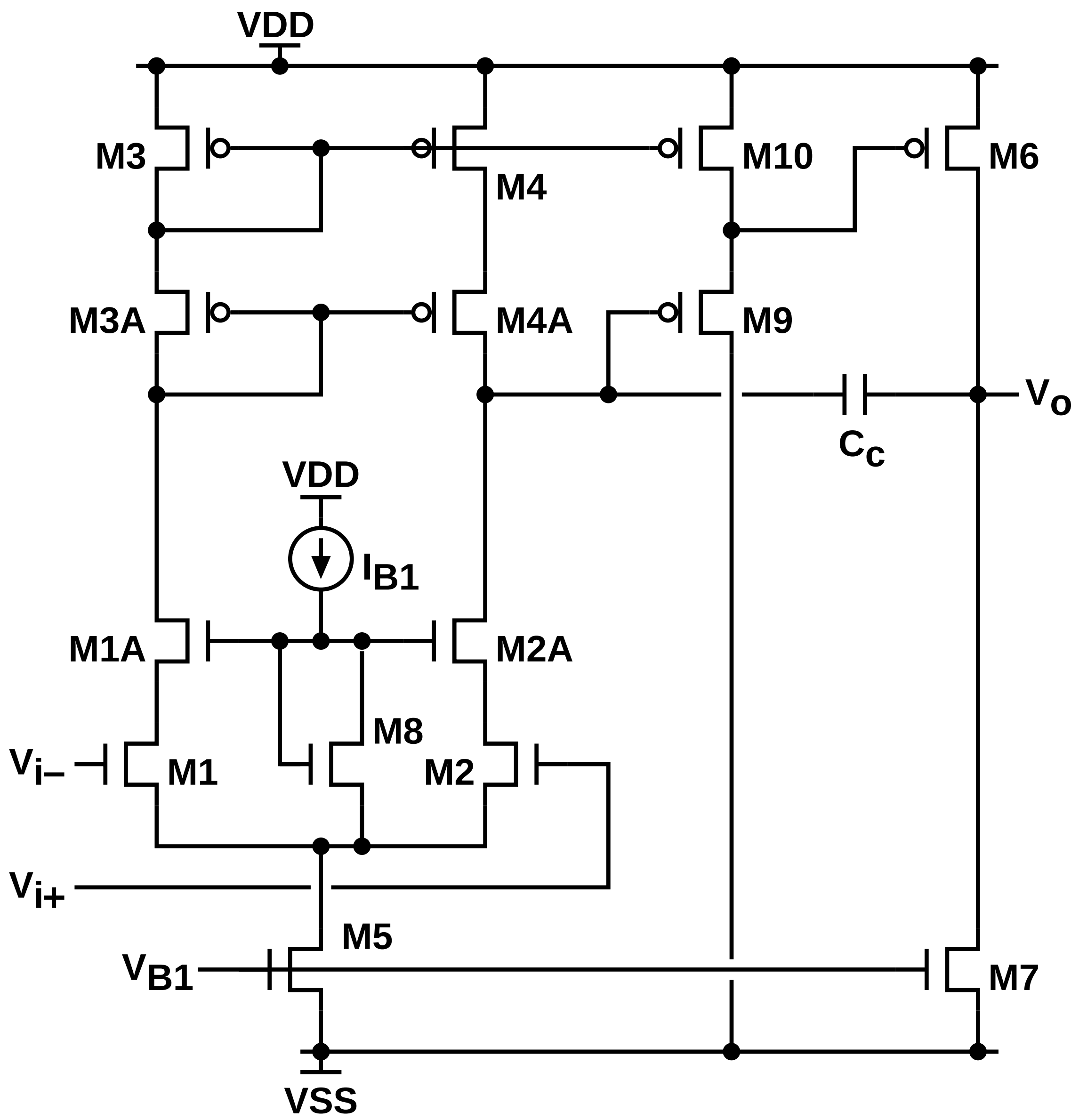
Two-Stage Operational Amplifier with Cascode

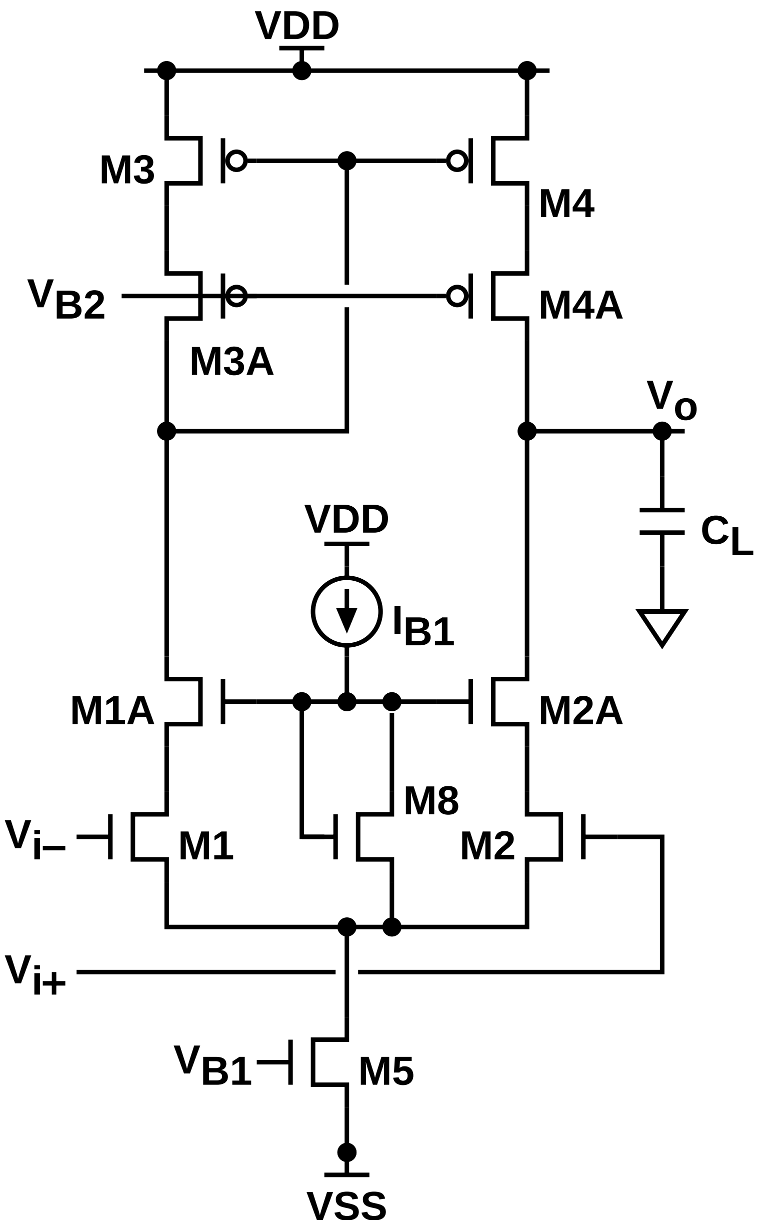


- The volage gain $A_v \propto (g_m r_o)^3$.
- Size M8 so that

$$V_{DS1} = V_{DS2} \simeq V_{DSAT}$$

- Input common-mode range is reduced by cascodes.
- The additional poles are non-dominant and located near ω_T.
- The 2nd stage can also use cascodes.

Telescopic-Cascode Operational Amplifier



- The volage gain $A_v \propto (g_m r_o)^2$.

- Consider the output current branch,

$$V_{DD} - (V_{IC} - V_t) > \Delta V_o + 3V_{ov}$$

$$\Rightarrow \quad V_{DD} - V_{IC} > \Delta V_o + 3V_{ov} - V_t$$

Since $V_{IC,min} = V_t + 2V_{ov} + V_{SS}$, we have

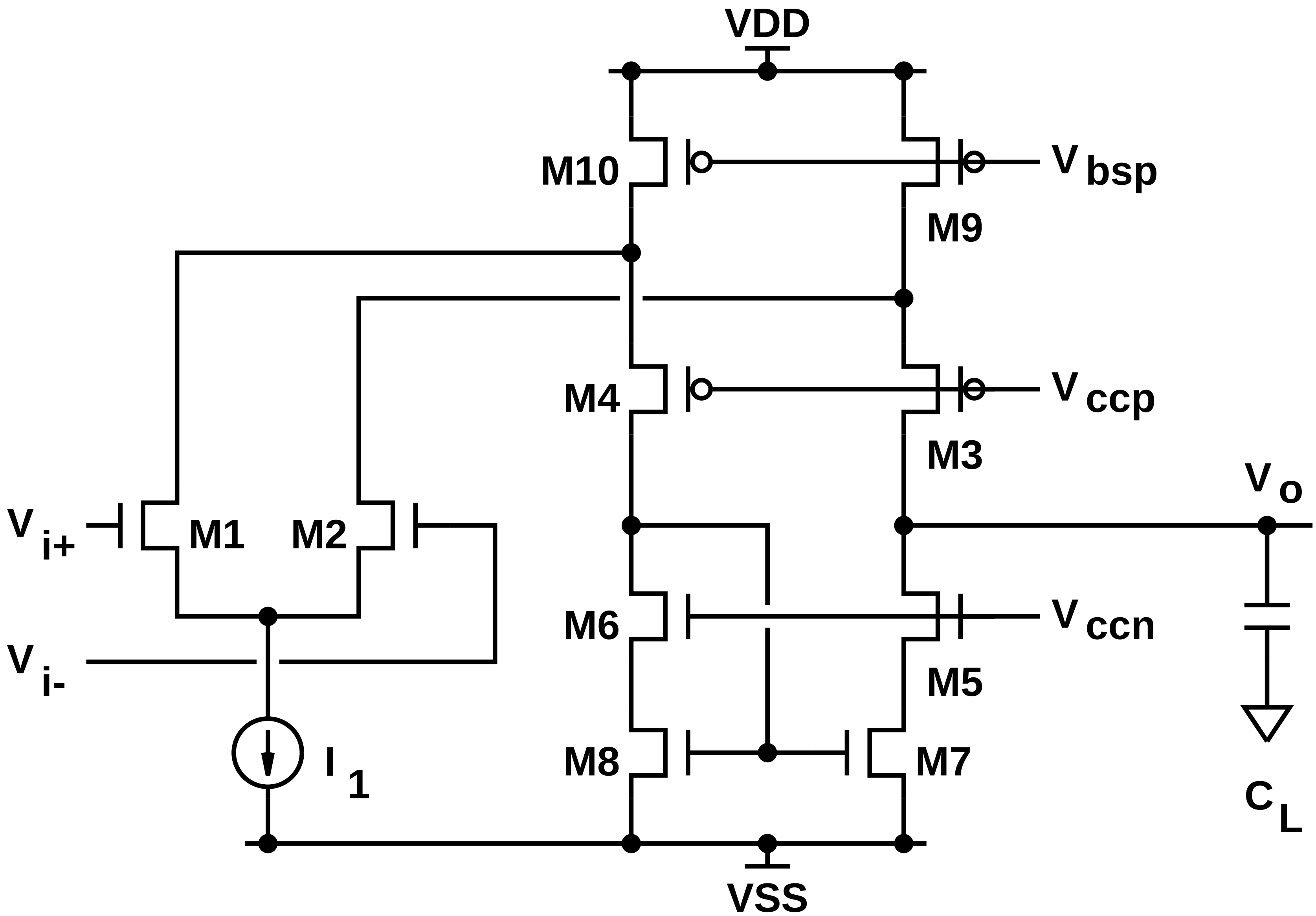
$$V_{DD} - V_{SS} > \Delta V_o + 5V_{ov}$$

- Consider the non-output branch,

$$V_{DD} - (V_{IC} - V_t) > V_t + 2V_{ov}$$

$$\Rightarrow \quad V_{DD} - V_{IC} > 2V_{ov} \quad \text{or} \quad V_{DD} - V_{SS} > V_t + 4V_{ov}$$

Folded-Cascode Operational Amplifier



Folded-Cascode Operational Amplifier

- Consider output stage

$$V_{DD} - V_{SS} > \Delta V_o + 4V_{ov} \quad \text{or} \quad V_{DD} - V_{SS} > V_t + 3V_{ov}$$

Consider input stage

$$V_{IC,max} = V_{DD} - V_{ov} + V_t \qquad V_{IC,min} = V_{SS} + V_t + V_{ov} + V_{o,min}(I_1)$$

- The differential-mode voltage gain is

$$A_v = \frac{A_v(0)}{1 - s/p_1} \qquad A_v(0) = g_{m1}R_o \qquad p_1 = -\frac{1}{R_oC_L} \qquad R_o = \frac{1}{\frac{g_{o2}+g_{o9}}{g_{m3}r_{o3}} + \frac{g_{o7}}{g_{m5}r_{o5}}}$$

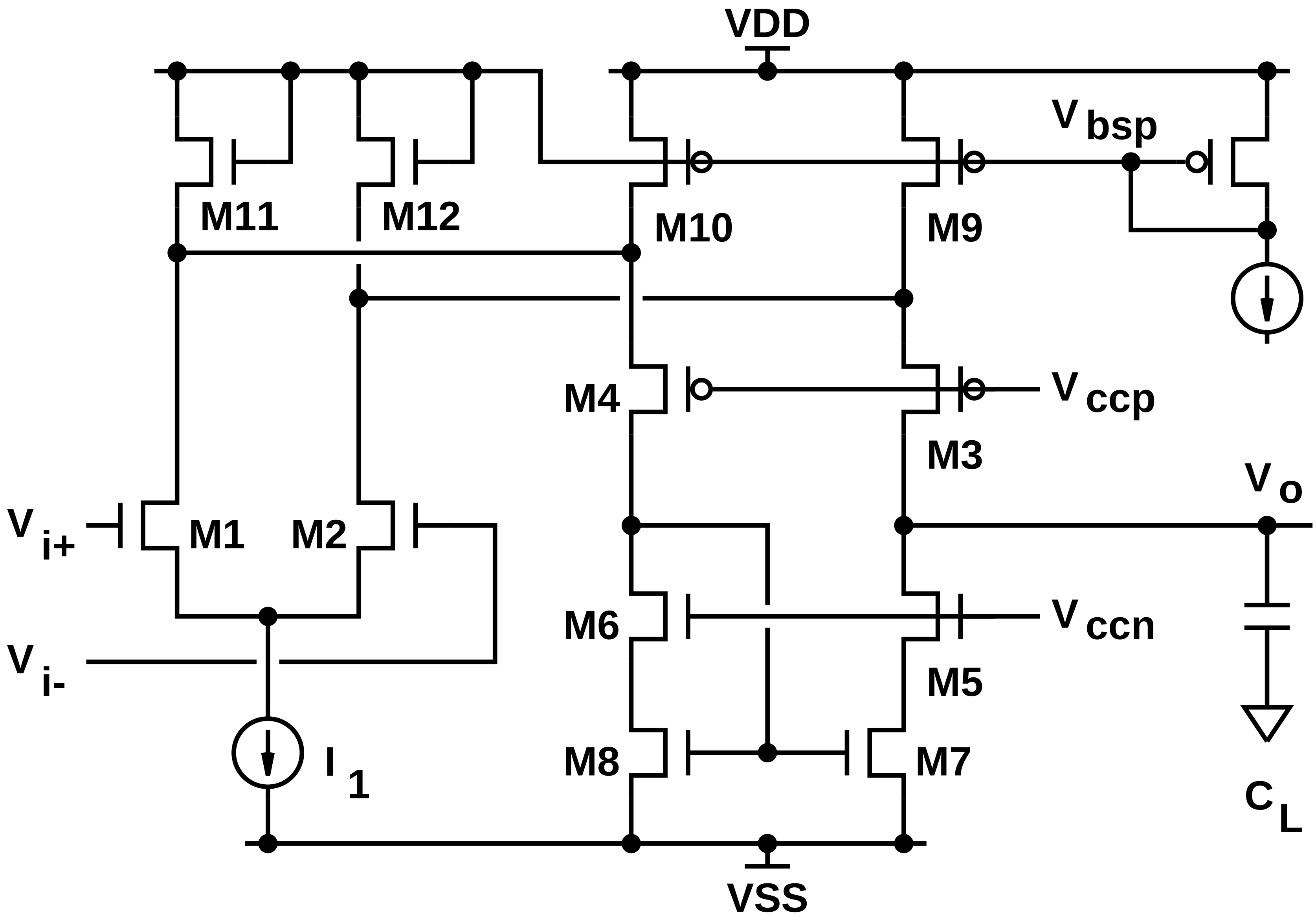
At midband frequencies where $\omega \gg |p_1|$

$$A_v \approx \frac{A_v(0)}{-s/p_1} = \frac{g_{m1}}{sC_L} = \frac{\omega_u}{s} \qquad \omega_u = \frac{g_{m1}}{C_L}$$

Folded-Cascode Operational Amplifier

- The dominant poile is associated with the only high-impedance node at V_o. All other poles are located near ω_T, and their magnitude are normally larger than $|p_2|$ of the two-stage opamps.
- C_L provides the dominant-pole frequency compensation. Increasing C_L *improves* the phase margin.
- If lead compensation is desired, a resistor can be placed in series with C_L.
- Use nMOST input stage for larger g_{m1} and better thermal noise performance.
- Good PSRR since no pole-splitting C_c.
- Slightly higher noise due to more devices.
- Suitable for low-voltage operation.
- Active cascodes can be used to increase voltage gain.

Folded-Cascode Operational Amplifier



Folded-Cascode Operational Amplifier

If bias currents $I_{D1,D2} > I_{D3,D4}$, i.e., $I_1 > I_{D9,D10}$,

- Without M11 and M12, the slew rate is

$$SR = \frac{I_{D9}}{C_L} = \frac{I_{D10}}{C_L}$$

- During slew condition, M11 and M12 can be used to clamp the drain volage of M1 and M2 to reduce bias recovery time, and increase I_{D9} and I_{D10} to improve SR.

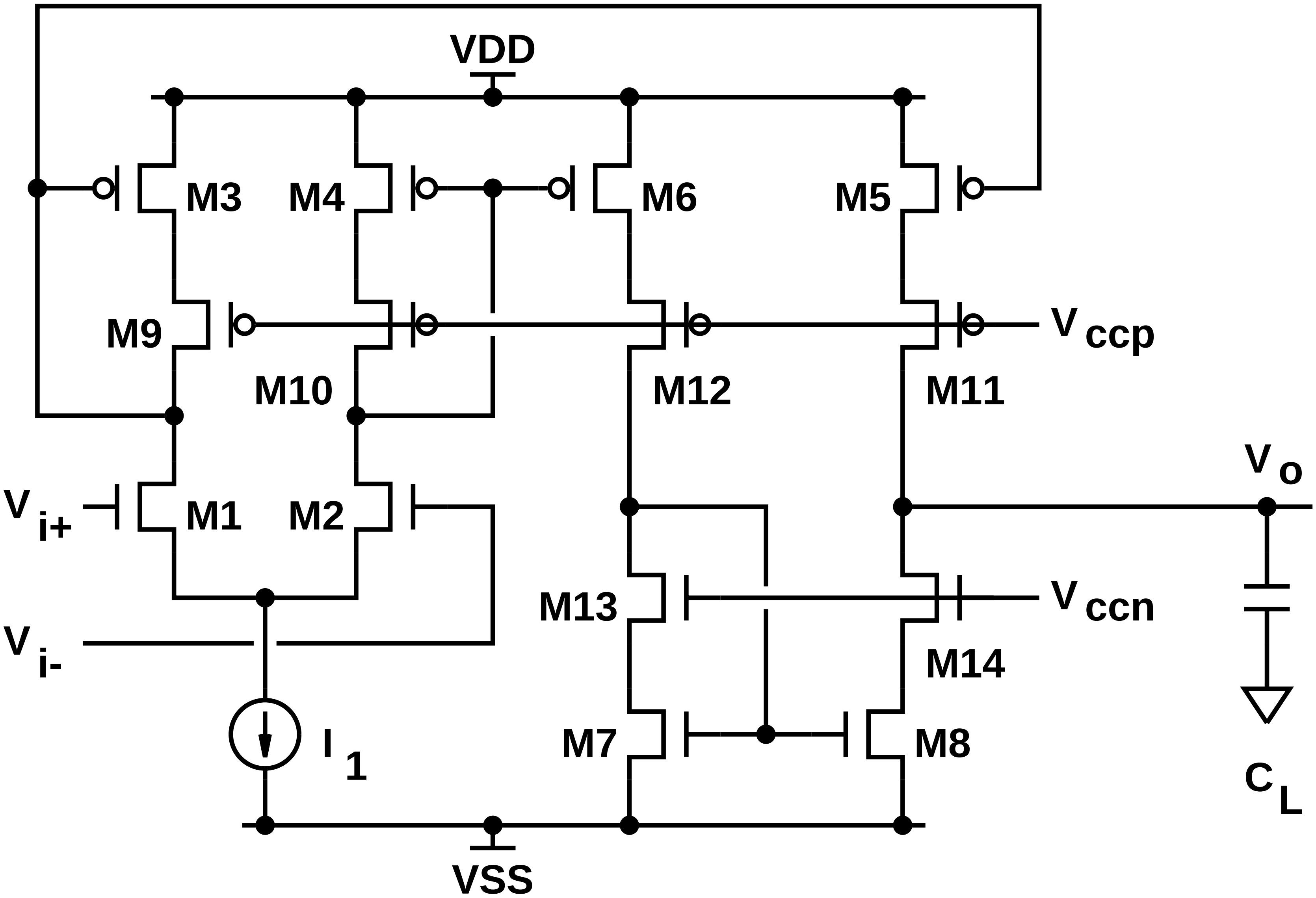
If bias currents $I_{D1,D2} < I_{D3,D4}$, i.e., $I_1 < I_{D9,D10}$,

- This slew rate is

$$SR = \frac{I_1}{C_L} \qquad I_1 = I_{D1} + I_{D2}$$

- M11 and M12 are not required.

Current-Mirror Operational Amplifier



Current-Mirror Operational Amplifier

$$\left(\frac{W}{L}\right)_3 = \left(\frac{W}{L}\right)_4 = \left(\frac{W}{L}\right)_6 = \frac{1}{K}\left(\frac{W}{L}\right)_5 \qquad \left(\frac{W}{L}\right)_7 = \frac{1}{K}\left(\frac{W}{L}\right)_8$$

$$I_{D1,D2} = I_{D3,D4} = I_{D6} = I_{D7} = \frac{1}{K}I_{D5} = \frac{1}{K}I_{D8} = \frac{1}{2}I_1 \qquad \text{SR} = \frac{KI_1}{C_L}$$

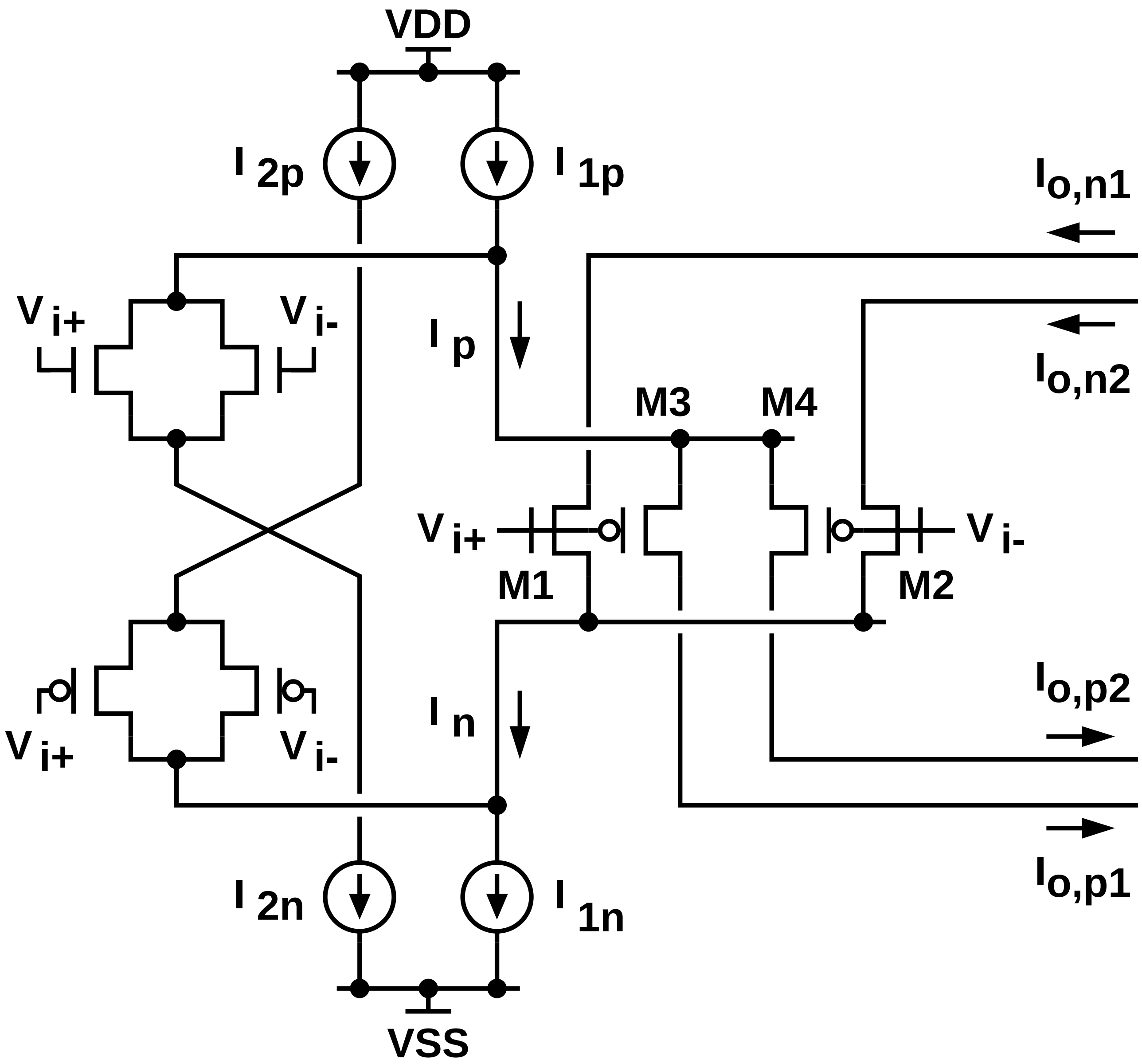
$$A_v(0) = Kg_{m1}R_o \qquad R_o = \frac{1}{\frac{g_{o5}}{g_{m11}r_{o11}} + \frac{g_{o8}}{g_{m14}r_{o14}}} \qquad p_1 = -\frac{1}{R_oC_L} \qquad \omega_u = \frac{Kg_{m1}}{C_L}$$

- For a given power dissipation, the current-mirror opamps have larger bandwidth and SR than the folded-cascode opamps. But they also suffer from larger thermal noise.

- For small C_L, K may have to be reduced to prevent the nondominant poles from degrading the phase margin.

- A practical upper limit on K is around 5. For a general-purpose opamp, $K \simeq 2$.

Rail-to-Rail Complementary Input Stage



Rail-to-Rail Complementary Input Stage

- Total input stage transconductance is

$$G_m = g_{m1} + g_{m3}$$

- G_m variation due to V_{ic} change can degrade CMRR. Want

$$g_{m1} + g_{m3} = \sqrt{\mu_n C_{ox}(W/L)_1 I_n} + \sqrt{\mu_p C_{ox}(W/L)_3 I_p} = \text{Constant}$$

If $\mu_n C_{ox}(W/L)_1 = \mu_p C_{ox}(W/L)_3$, want

$$\sqrt{I_n} + \sqrt{I_p} = \sqrt{I_{1n} - I_{2p}} + \sqrt{I_{1p} - I_{2n}} = \text{Constant}$$

Rail-to-Rail Complementary Input Stage

- Let

$$I_{1n} = I_{1p} = 4I \qquad I_{2n} = I_{2p} = 3I$$

At $V_{ic} \simeq (V_{DD} - V_{SS})/2$

$$\sqrt{I_n} + \sqrt{I_p} = \sqrt{1I} + \sqrt{1I} = 2\sqrt{I}$$

At $V_{ic} \simeq V_{SS}$, $I_n = 0$ and $I_{2n} = 0$,

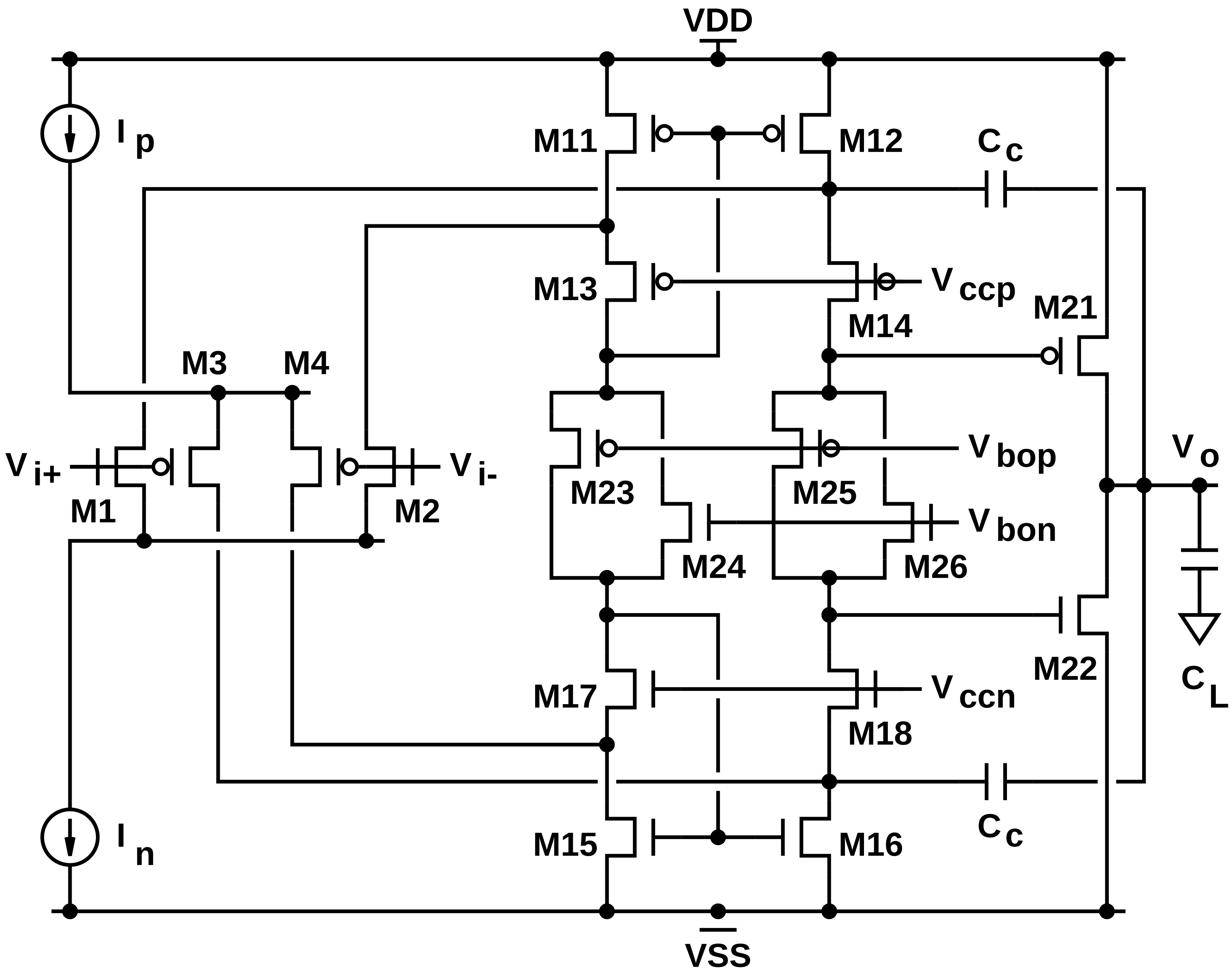
$$\sqrt{I_n} + \sqrt{I_p} = \sqrt{0I} + \sqrt{4I} = 2\sqrt{I}$$

At $V_{ic} \simeq V_{DD}$, $I_p = 0$ and $I_{2p} = 0$,

$$\sqrt{I_n} + \sqrt{I_p} = \sqrt{4I} + \sqrt{0I} = 2\sqrt{I}$$

- Less than 5% change in G_m is possible.
- The variation of the input-referred dc offset V_{OS} due to V_{ic} change also degrade CMRR.

A Rail-to-Rail Input/Output Opamp



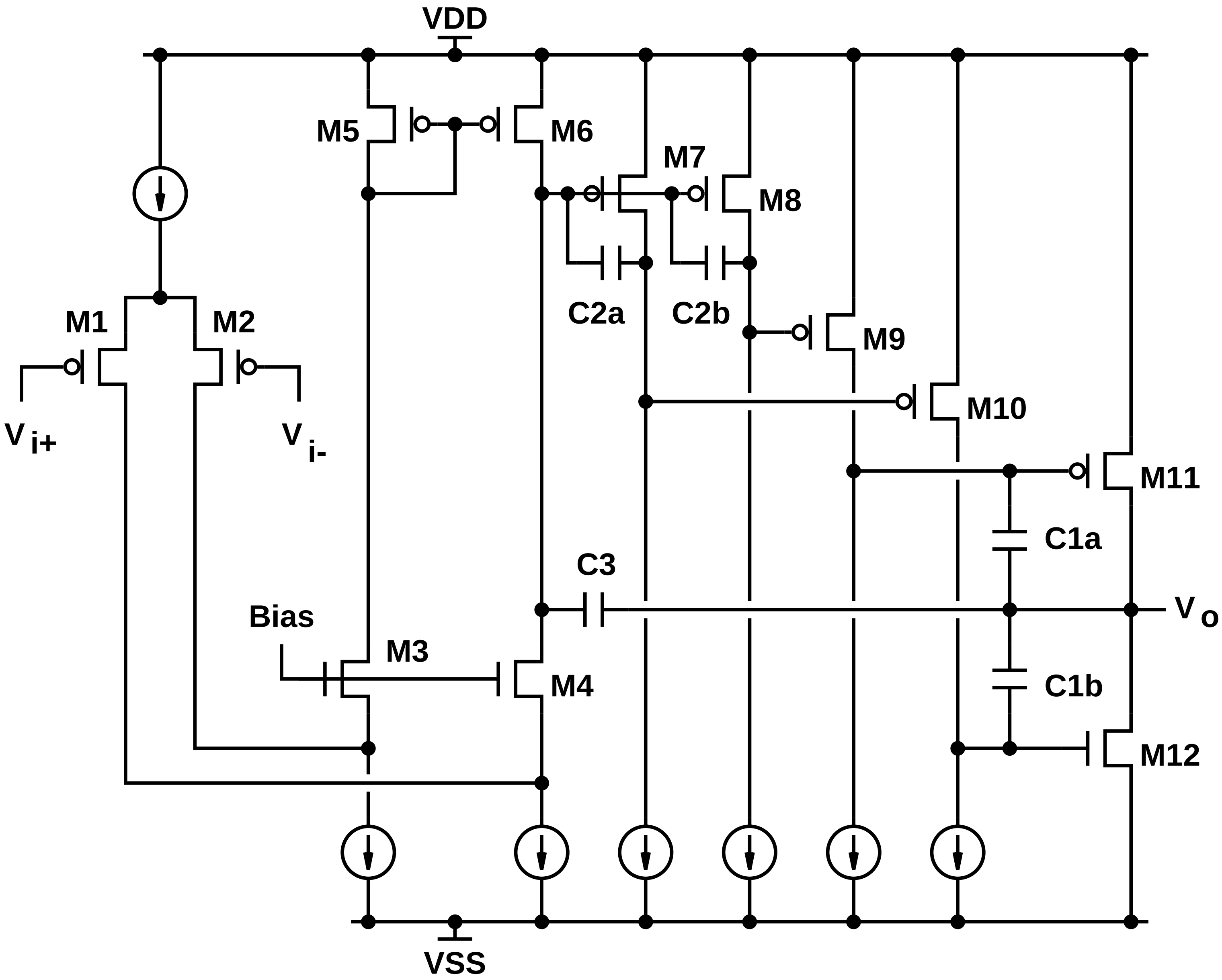
A Rail-to-Rail Input/Output Opamp

- Two cascaded gain stages.
- Noises in V_{bop} and V_{bon} are canceled at output.
- The bias of the output stage is insensitive to variations in I_p, I_n and supply voltage.
- The two C_c are connected as Miller frequency compensation using common-gate stages.
- The output pole is

$$p_2 = \frac{C_c}{C_{gso}} \times \frac{g_{mo}}{C_L}$$

  where g_{mo} and C_{gso} are respectively the total g_m and C_{gs} of the output stage.
- Reference: Hogervorst, et al., JSSC 12/94, pp. 1505–1513.

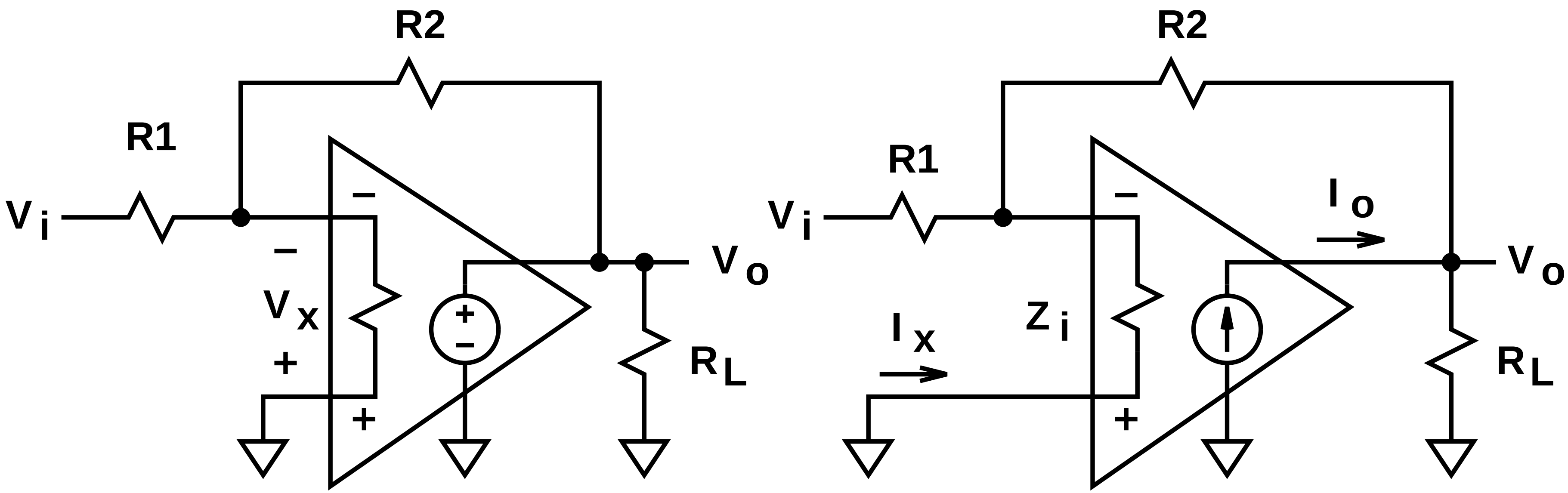
Low-Voltage Multi-Stage Opamp



Low-Voltage Multi-Stage Opamp

- Four cascaded gain stages.
- Hybrid nested Miller compensation.
- Class-AB output stage.
- A supply voltage below 1.5 V is possible.
- Reference: Eschauzier, et al., JSSC 12/94, pp. 1497–1504.

Current-Feedback Configuration



Voltage-Feedback Opamp **Current-Feedback Opamp**

For the voltage-feedback opamp, let $V_o/V_x = A \approx \omega_u/s$ and $Z_i \rightarrow \infty$, then

$$A_v = \frac{V_o}{V_i} = -\frac{R_2}{R_1} \cdot \frac{1}{1 + \frac{1}{A}\left(1 + \frac{R_2}{R_1}\right)} \approx -\frac{R_2}{R_1} \cdot \frac{1}{1 + \frac{s}{\omega_u}\left(1 + \frac{R_2}{R_1}\right)}$$

- Trade-off between closed-loop gain and closed-loop bandwidth.

Current-Feedback Configuration

For the current-feedback opamp, let $I_o/I_x = A \approx \omega_u/s$, then

$$A_v = \frac{V_o}{V_i} = -\frac{R_2}{R_1} \cdot \frac{1 - \frac{Z_i}{AR_2}}{1 + \frac{1}{A}\left[1 + \frac{R_1R_2 + Z_i(R_1+R_2+R_L)}{R_1R_L}\right]} \approx -\frac{R_2}{R_1} \cdot \frac{1}{1 + \frac{s}{\omega_u}\left[1 + \frac{R_2 + Z_i\left(1+\frac{R_2+R_L}{R_1}\right)}{R_L}\right]}$$

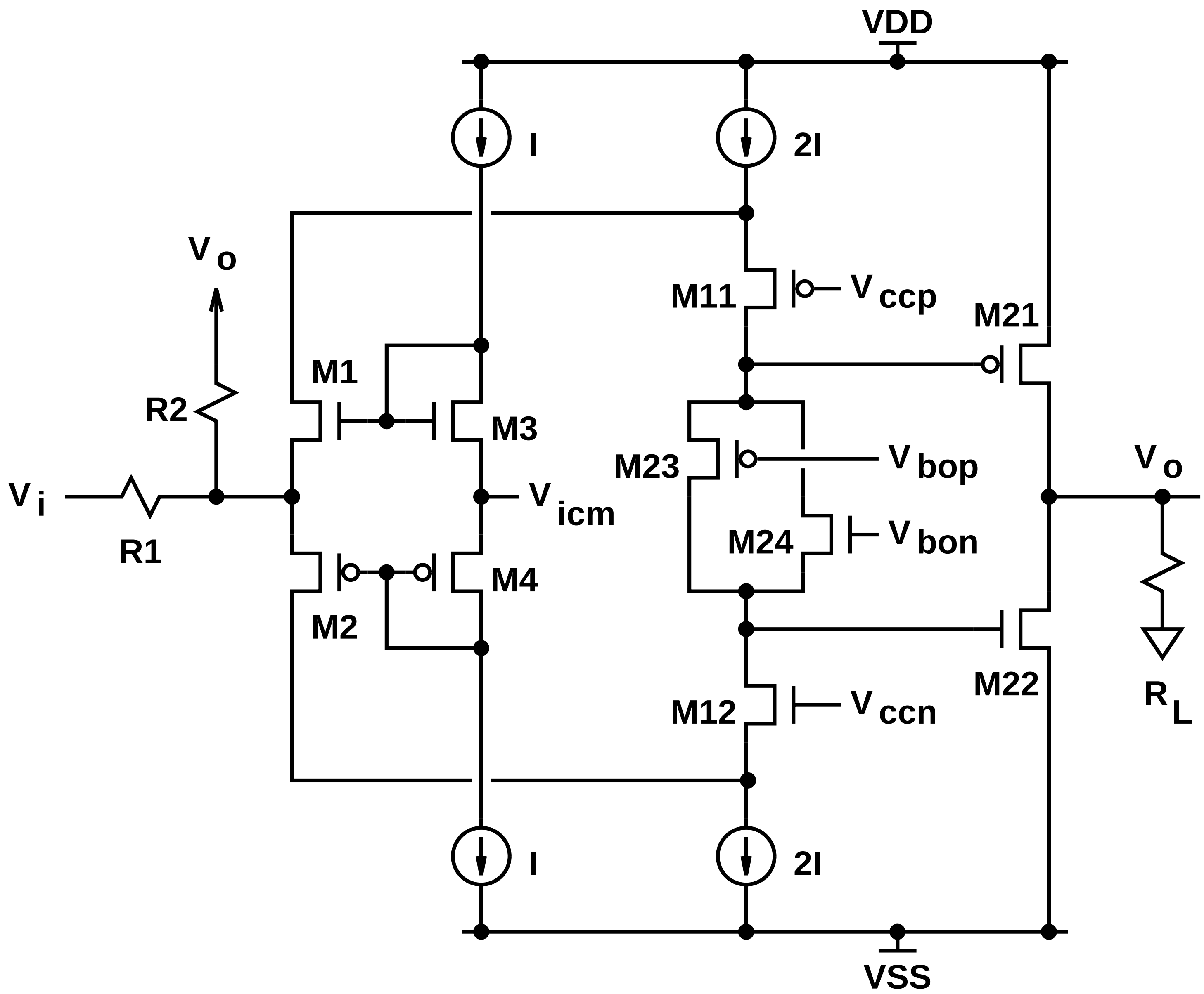
If $Z_i \longrightarrow 0$,

$$A_v \approx -\frac{R_2}{R_1} \cdot \frac{1}{1 + \frac{s}{\omega_u}\left(\frac{R_2}{R_L}\right)}$$

- The closed-loop gain can be modified by changing R_1, leaving the closed-loop bandwidth unchanged.

- For a given R_2, frequency compensation can be optimized. Suitable for high-frequency applications.

A CMOS Current-Feedback Driver



A CMOS Current-Feedback Driver

- This opamp has been designed to drive $R_L = 25\ \Omega$ and provide 50 mA of output current.

- Two-stage opamp with only one high-impedance node.

- C_{gs} and C_{gd} of M21 and M22 are large enough to provide adequate frequency compensation.

- The class-AB common-gate input stage provides large internal slew rate.

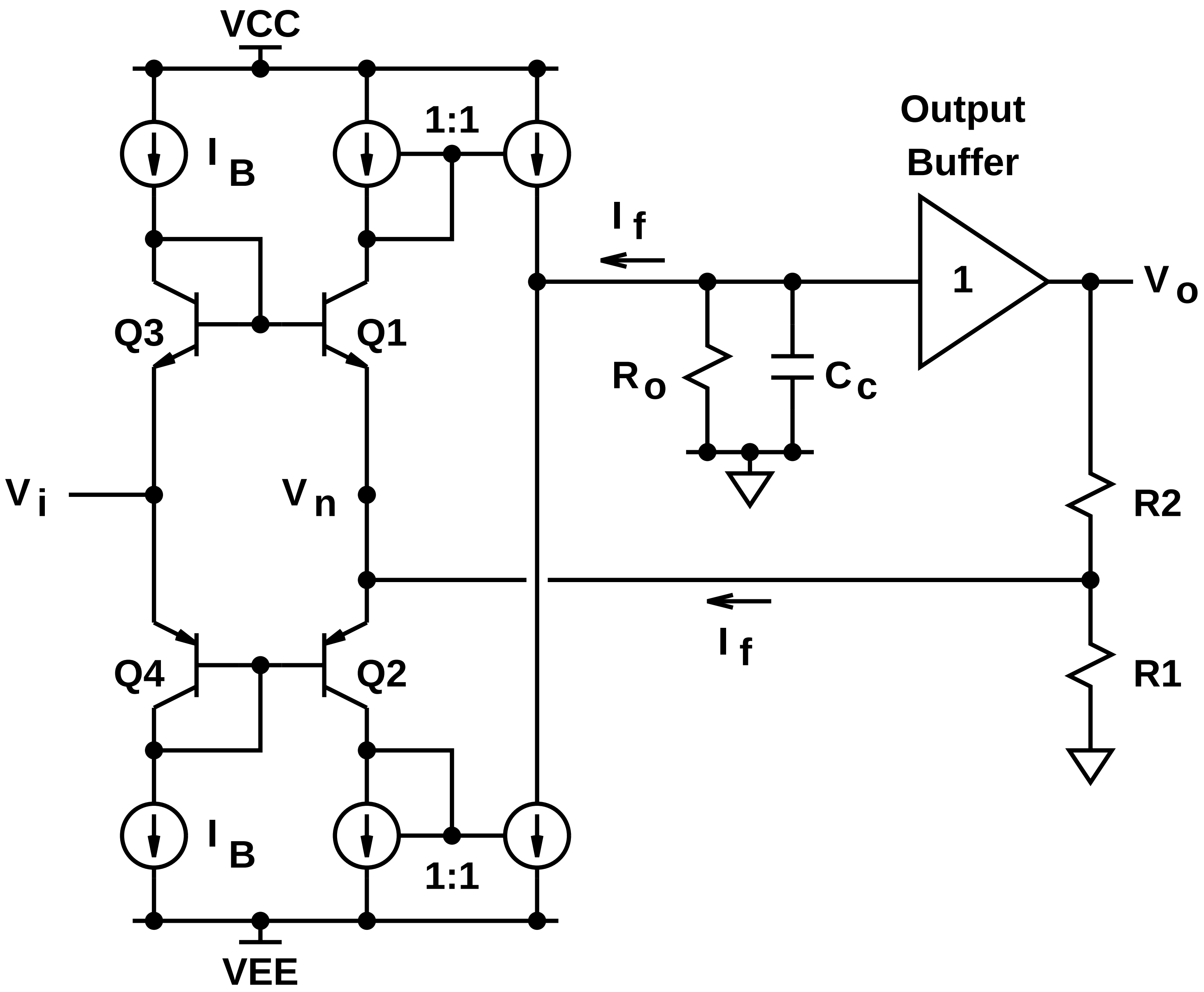
- Large voltage swing of V_{gs21} and V_{gs22} are required.

- Open-loop current gain is determined by the output stage,

$$A(s) \approx \frac{g_{mo}}{sC_{go}} = \frac{\omega_u}{s} \qquad \omega_u = \frac{g_{mo}}{C_{go}}$$

- Loop gain $T(s) \approx A(s)R_L/(R_L + R_2)$ is independent of R_1.

A General-Purpose BJT Current-Feedback Opamps



A General-Purpose BJT Current-Feedback Opamps

Due to the symmetry of the input stage, we have $V_i = V_n$.

If $R_1 \parallel R_2 \gg 1/(g_{m1} + g_{m2})$, we have

$$I_f = \frac{V_o - V_n}{R_2} - \frac{V_n}{R_1} = V_o\left(\frac{1}{R_2}\right) - V_i\left(\frac{1}{R_1} + \frac{1}{R_2}\right) \qquad V_o = -I_f\left(\frac{1}{sC_c + 1/R_o}\right)$$

$$A_v = \frac{V_o}{V_i} = \left[\frac{R_o(R_1 + R_2)}{(R_o + R_2)R_1}\right]\left[\frac{1}{1 + sC_c(R_o \| R_2)}\right]$$

If $R_o \gg R_2$,

$$A_v \approx \left(1 + \frac{R_2}{R_1}\right)\left(\frac{1}{1 + sC_cR_2}\right)$$

Also the loop gain is

$$T(s) = \left(\frac{1}{sC_c + 1/R_o}\right)\left(\frac{1}{R_2}\right) \approx \frac{1}{sC_cR_2}$$