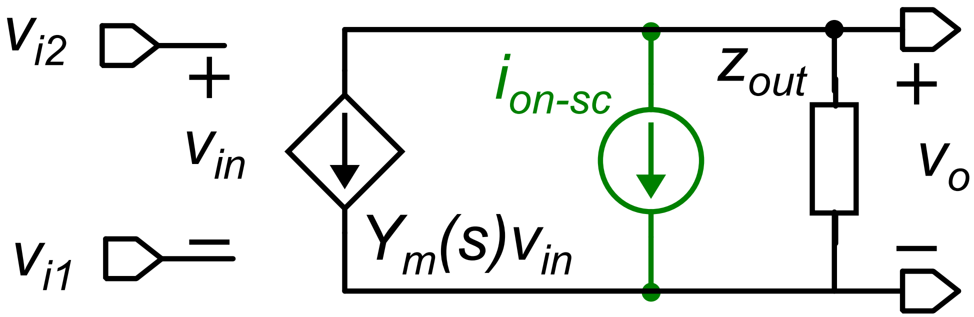
Amplifier Norton schematization with output referred noise source



It is possible to model any amplifier (whole amplifier or single amplifier stage) with a Norton equivalent circuit of the output port

and take into account noise with an additional current source $\boldsymbol{i_{on\text{-}sc}}$

$$v_o = -\left(Y_m v_{in} + i_{on-sc}\right) Z_{out} = -Y_m Z_{out}\left(v_{in} + \frac{i_{on-sc}}{Y_m}\right)$$

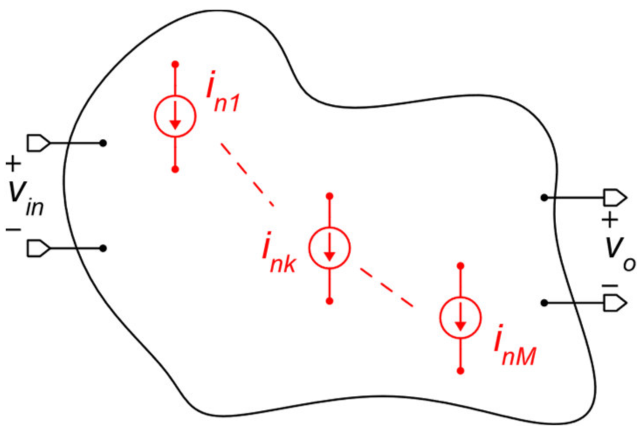
General input-output law of a voltage amplifier with noise/offset: $v_o = A_V\left(v_{in} - v_n\right)$

$$A_V = -Y_m Z_{out}$$

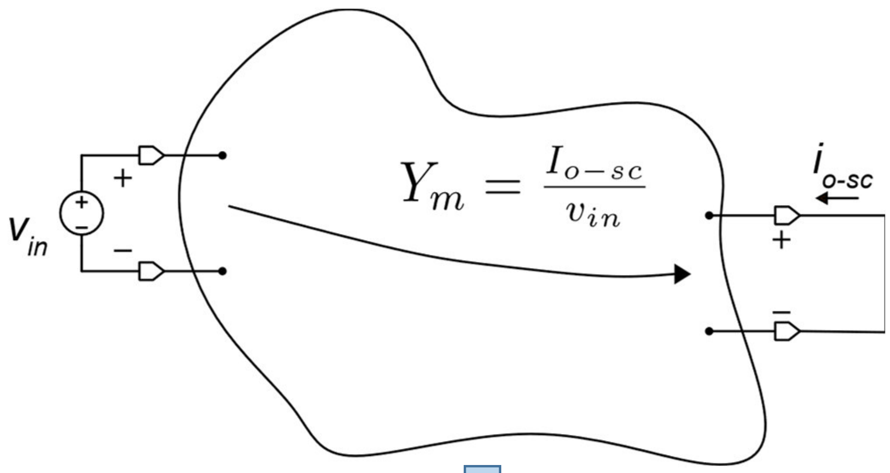
Input referred noise voltage

$$v_n = \frac{-i_{on-sc}}{Y_m}$$

General method to calculate the input referred noise / offset

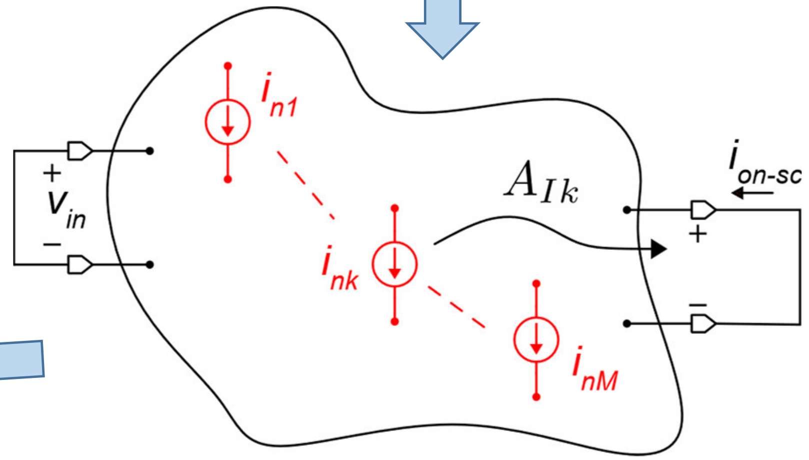


Generic amplifier with *M* internal noise current sources



Step 1. Noise sources off, input signal on, determine ***Ym***

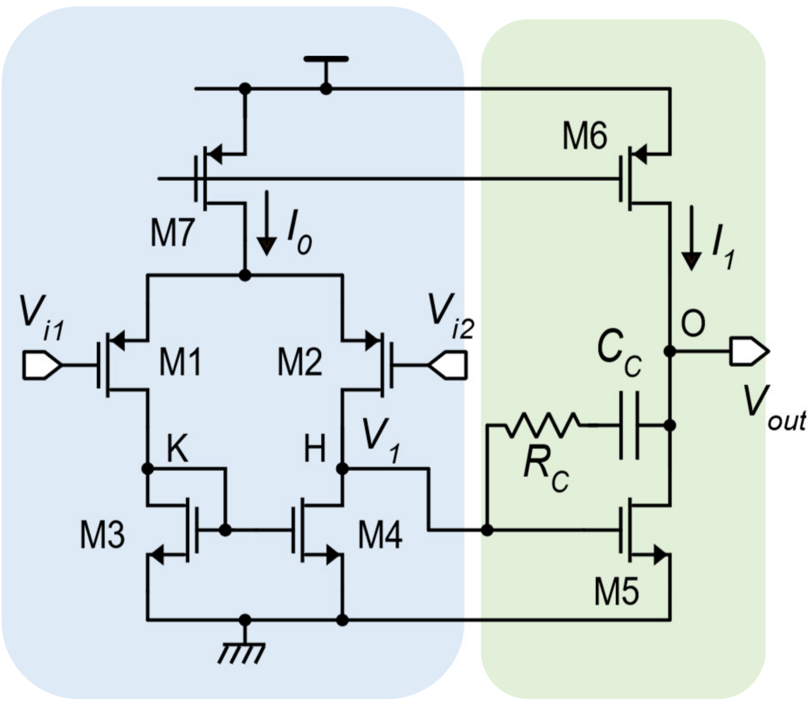
Step. 2 Signal off, determine $\boldsymbol{i_{on\text{-}sc}}$ with the superposition theorem



$$i_{on-sc} = \sum_{k=1}^{M} A_{Ik} i_{nk}$$

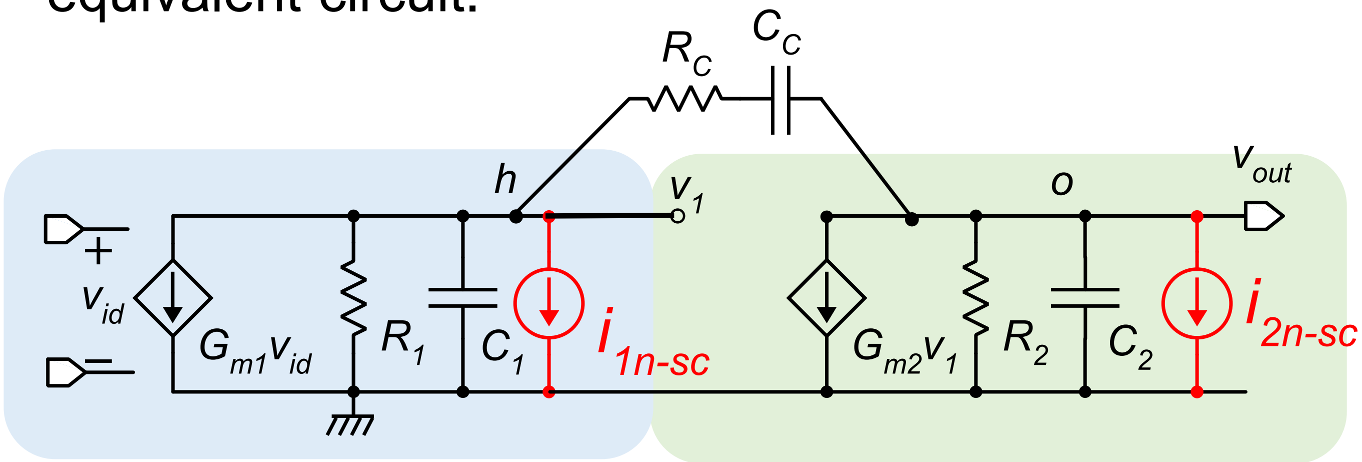
$$v_n = \frac{-i_{on-sc}}{Y_m}$$

Application of the method to the two-stage op-amp



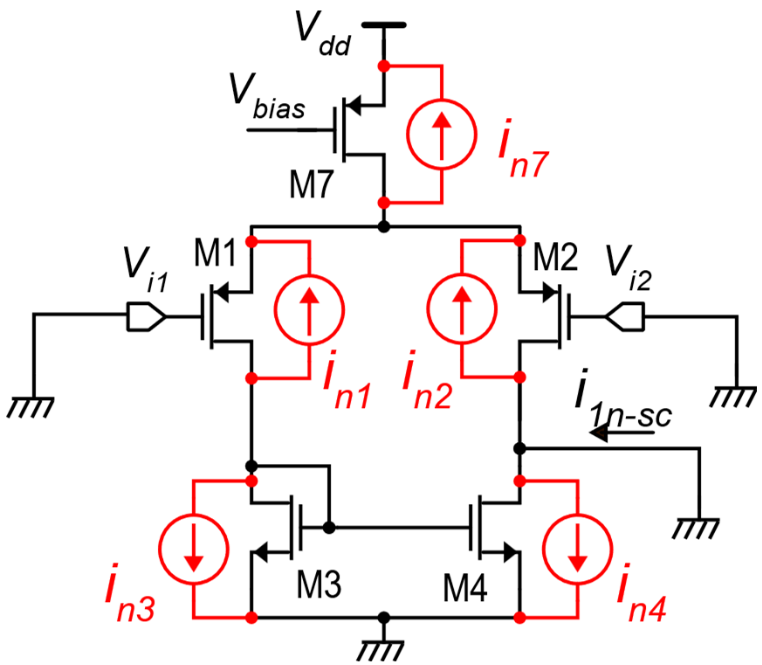
First stage Second stage

It is convenient to calculate the equivalent output noise currents of the two stages individually, and then study the whole amplifier using the following equivalent circuit.



First stage Second stage

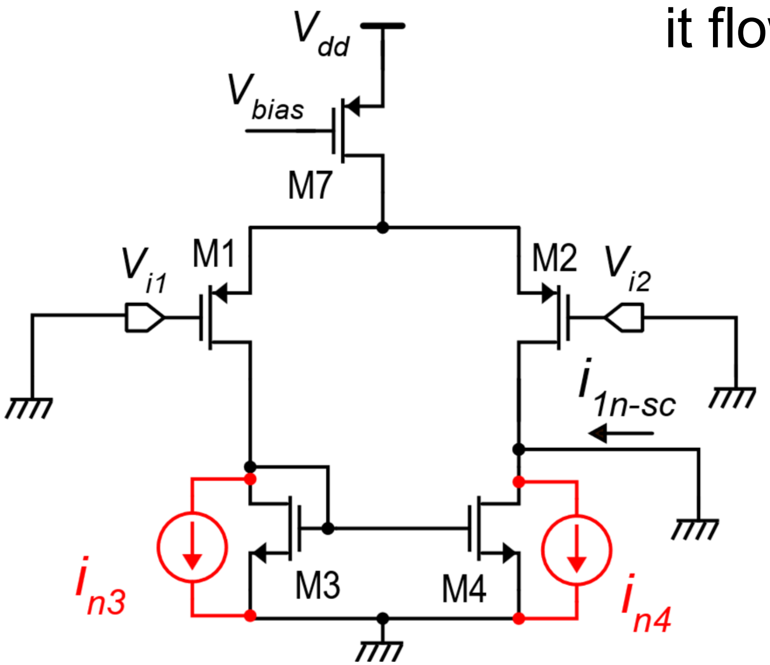
Output noise short circuit current of the first stage



Input stage with noise current sources of all devices

In order to calculate the output noise short-circuit current, we need to calculate the current gains $\boldsymbol{A_{Ik}}$, from each one of the MOSFET noise sources to the output short circuit current.

Effect of i_{n3}, i_{n4}



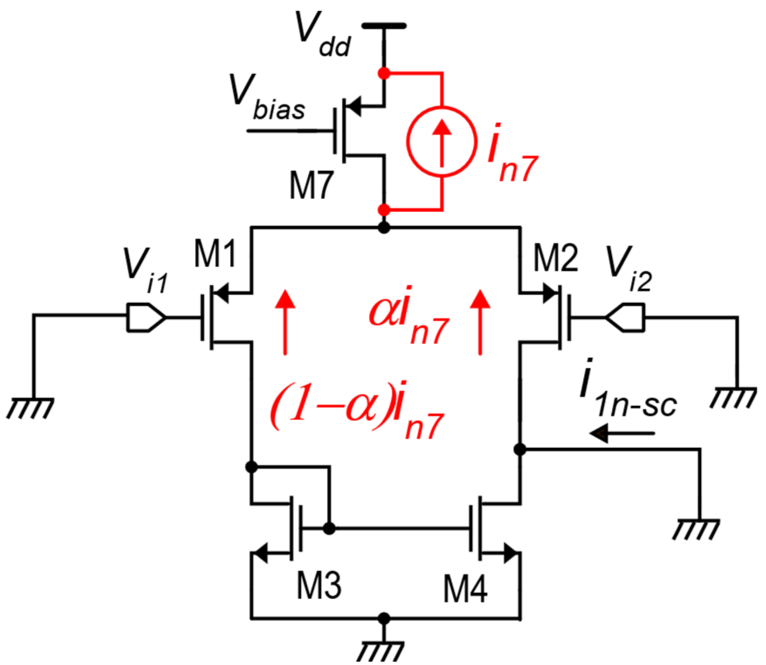
i_{n4} is directly connected to the output port, then it flows directly into the output short circuit:

$$A_{I4} = 1$$

i_{n3} is directly connected to the input of the current mirror. It sees a low resistance towards the mirror and high resistance towards M1 ($2r_{d1}$). Then it flows almost completely into the mirror and reaches the output port after an inversion (caused by the mirror).

$$A_{I3} \cong -1$$

Effect of i_{n7}



$$i_{on-cc}\left(i_{n7}\right) \cong i_{n7}\left[\alpha-\left(1-\alpha\right)\right]=i_{n7}\left[2\alpha-1\right]$$

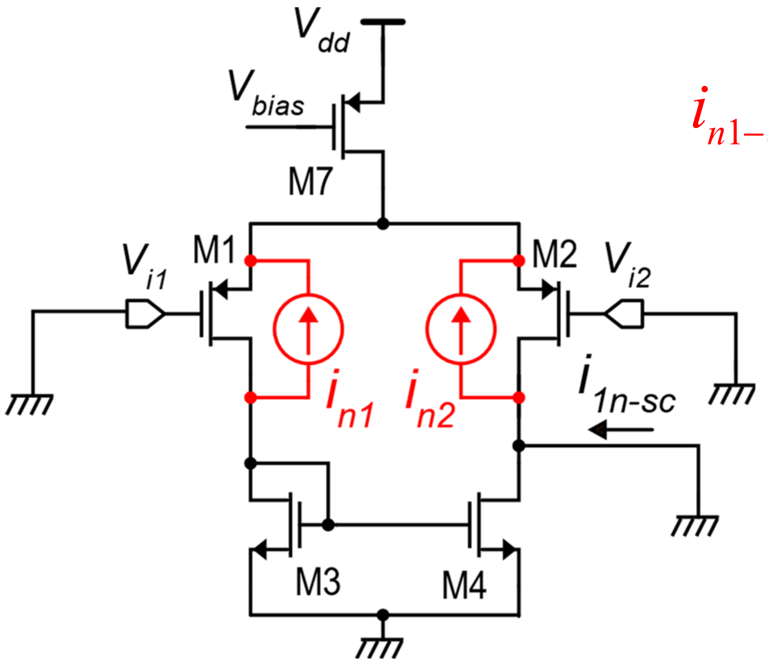
In the case of perfect symmetry and zero input differential voltage (V_{id}=0), which is the case that we are analyzing:

$$\alpha=\frac{1}{2} \Rightarrow i_{1n-sc}\left(i_{n7}\right) \cong 0 \qquad A_{I7} \cong 0$$

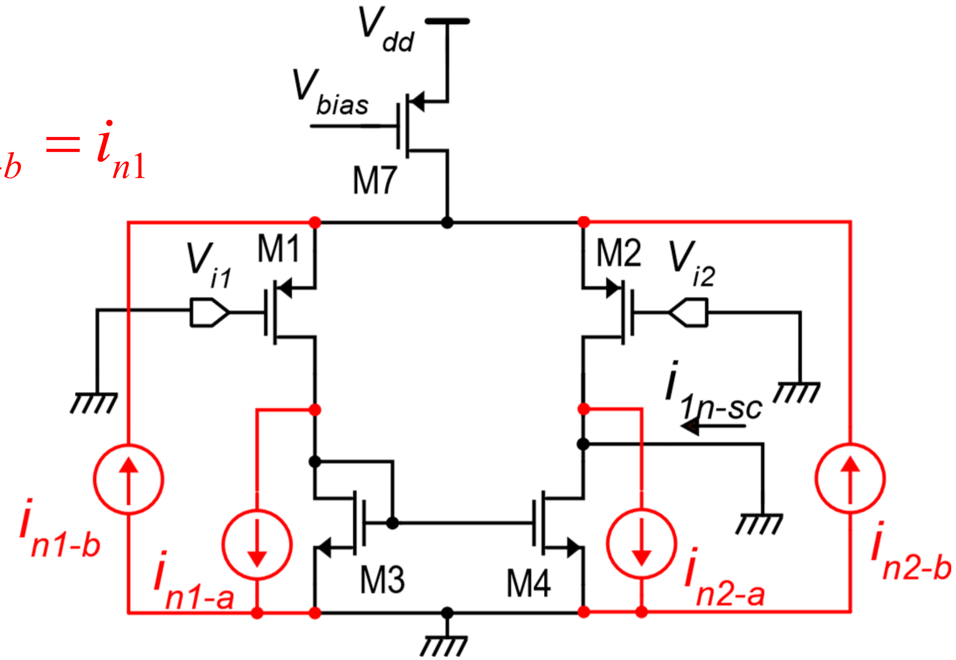
If a relatively large input differential voltage is present, α can be significantly different from 0.5 and the effect of i_{n7} is no more negligible.

In the following part of this analysis, we will consider α=0.5

Effect of i_{n1}, i_{n2}



$$i_{n1-a} = i_{n1-b} = i_{n1}$$



Since i_{n1} and i_{n2} are floating, we can split them into two sources with a terminal at *gnd*.

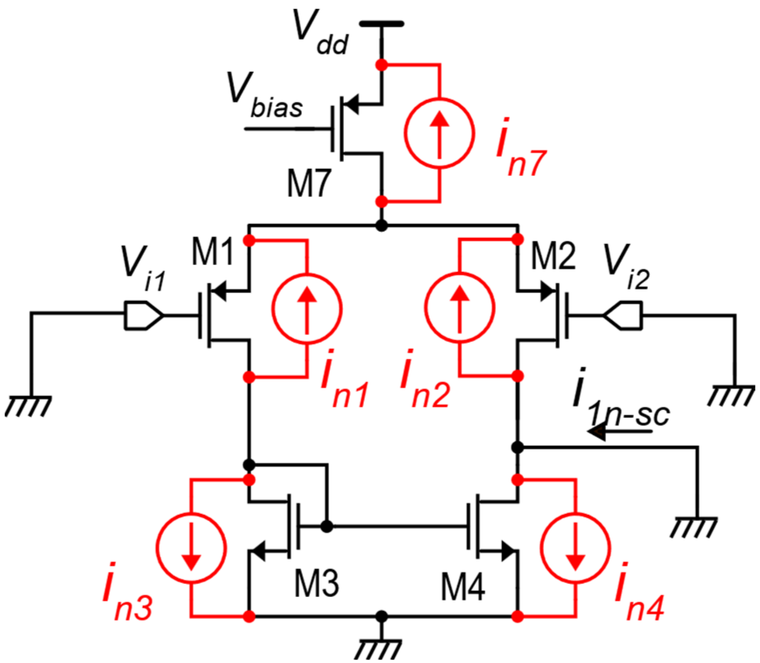
$i_{n1\text{-}a}$: same effect as i_{n3}: $A_{I1\text{-}a} \cong -1$

$i_{n1\text{-}b}$: same effect as i_{n7}: $A_{I1\text{-}b} \cong 0$

$$A_{I1} \cong -1$$

Repeating the procedure for i_{n2} $\quad A_{I2} \cong 1$

Putting all contribution together for the first stage:



$$A_{I1} \cong -1$$

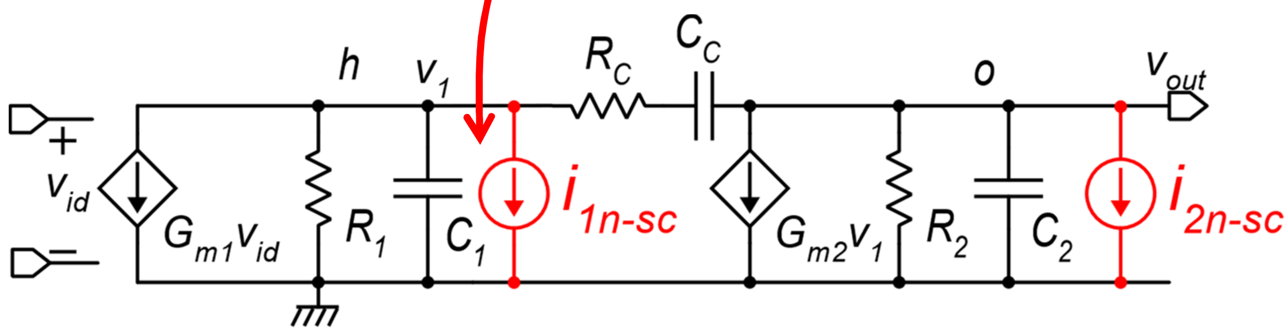
$$A_{I2} \cong 1$$

$$A_{I3} \cong -1$$

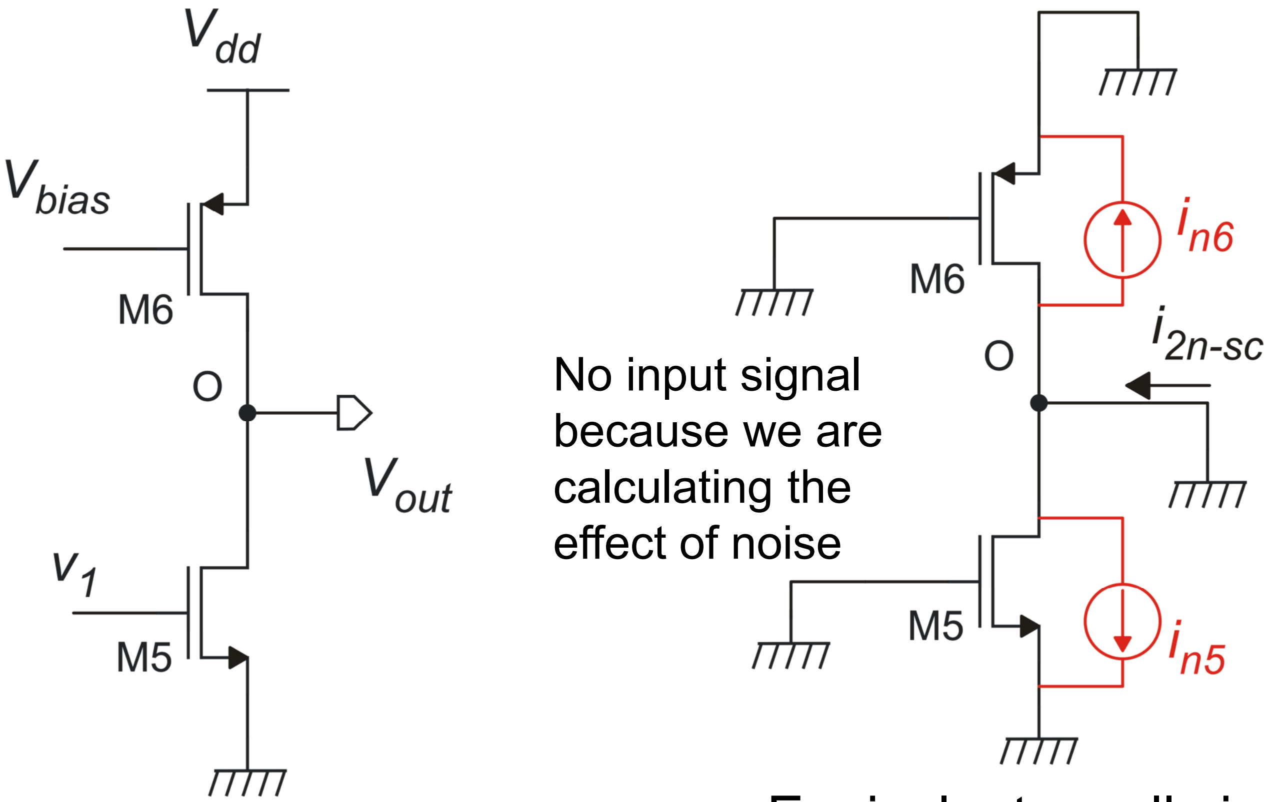
$$A_{I4} \cong 1$$

$$A_{I7} \cong 0$$

$$i_{1n-sc} \cong i_{n2} - i_{n1} + i_{n4} - i_{n3}$$



Equivalent output noise current of the second stage

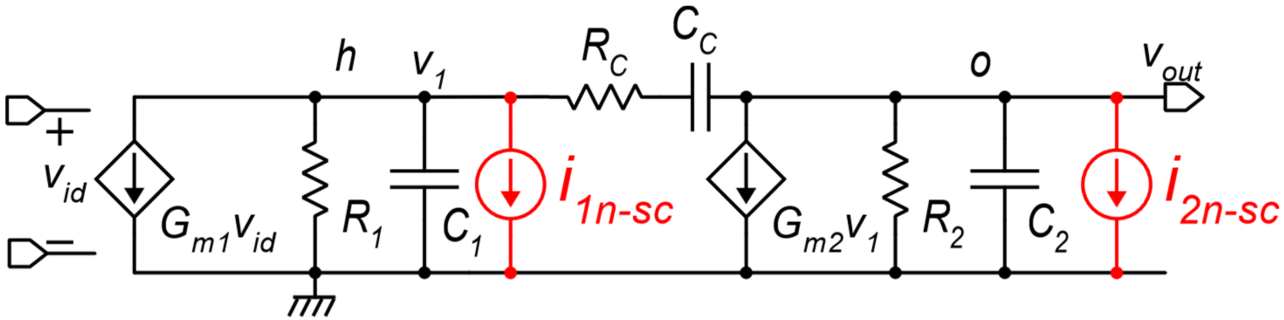


No input signal because we are calculating the effect of noise

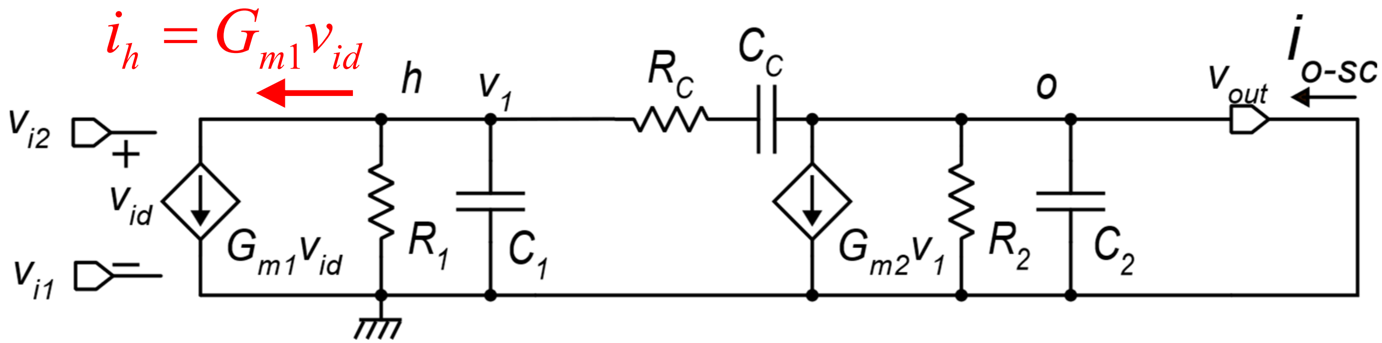
Equivalent small signal circuit with current sources

$$i_{2n-sc} \cong i_{n5} + i_{n6}$$

Putting the two stages together



Step 1. Calculate the Y_m of the op-amp



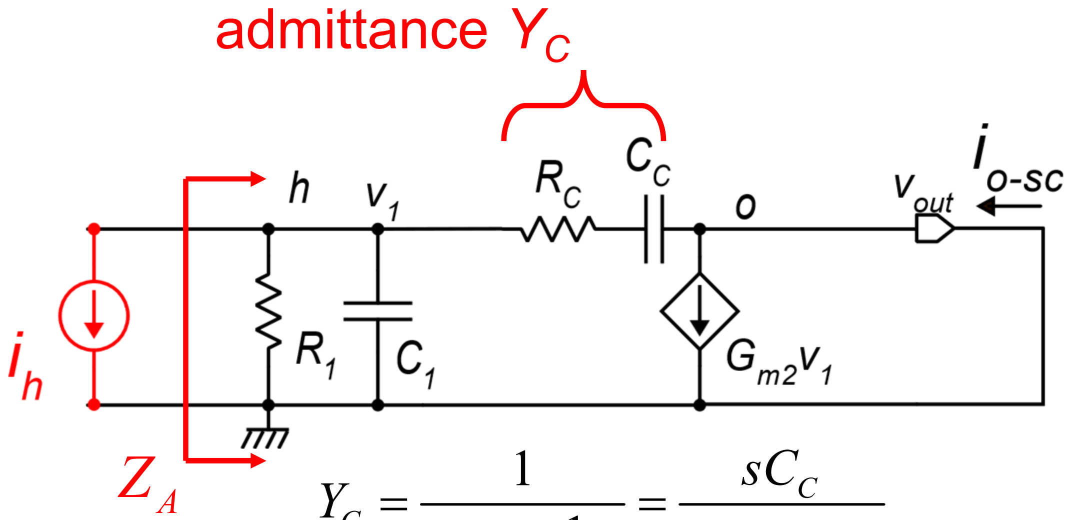
$$i_{o-sc} = G_{m1}v_{id} \cdot A_{Ih}$$

$$A_{Ih} = \frac{i_{o-sc}}{i_h}$$

$$Y_m = \frac{i_{o-sc}}{v_{id}} = G_{m1}A_{Ih}$$

A_{Ih} is the transfer function (current gain) from a current source connected between node h and gnd to the output short circuit current

Calculation of A_{Ih}



$$Y_C = \frac{1}{R_C + \frac{1}{sC_C}} = \frac{sC_C}{1 + sR_C C_C}$$

$$Z_A = \frac{1}{Y_C + \frac{1}{R_1} + sC_1}$$

$$i_{o-sc} = v_1 G_{m2} - v_1 Y_C = v_1 \left(G_{m2} - Y_C\right)$$

$$v_1 = -Z_A i_h \Rightarrow i_{o-sc} = -Z_A i_h \left(G_{m2} - Y_C\right)$$

$$A_{Ih} = \frac{i_{o-sc}}{i_h} = -Z_A \left(G_{m2} - Y_C\right)$$

$$G_{m2} - Y_C = \left(G_{m2} - \frac{sC_C}{1 + sR_C C_C}\right) = \frac{G_{m2} - sC_C\left(1 - G_{m2} R_C\right)}{1 + sR_C C_C}$$

since we set $R_C = \frac{1}{G_{m2}} \Rightarrow G_{m2} - Y_C = \frac{G_{m2}}{1 + sR_C C_C}$

Calculation of A_{Ih}

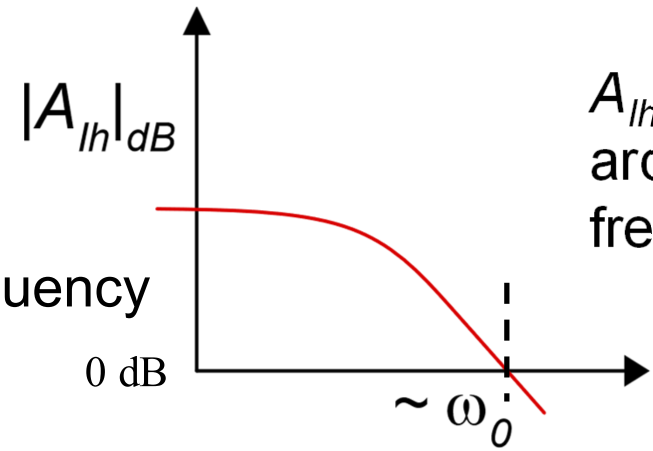
$$Z_A = \frac{1}{\dfrac{sC_C}{1+sC_CR_C} + \dfrac{1}{R_1} + sC_1} = \frac{R_1\left(1+sC_CR_C\right)}{1+s\left(C_CR_1 + C_CR_C + C_1R_1\right) + s^2R_CC_CR_1C_1}$$

$$G_{m2} - Y_C = \frac{G_{m2}}{1+sR_CC_C}$$

$$A_{Ih} = -Z_A\left(G_{m2} - Y_C\right) = \frac{-G_{m2}R_1}{1+s\left(C_CR_1 + C_CR_C + C_1R_1\right) + s^2R_CC_CR_1C_1}$$

dc value: $A_{Ih}(0) = -G_{m2}R_1$

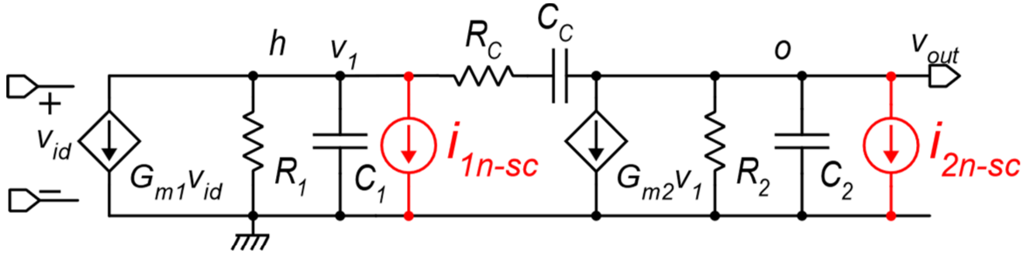
$\left|A_{Ih}(0)\right| >> 1$

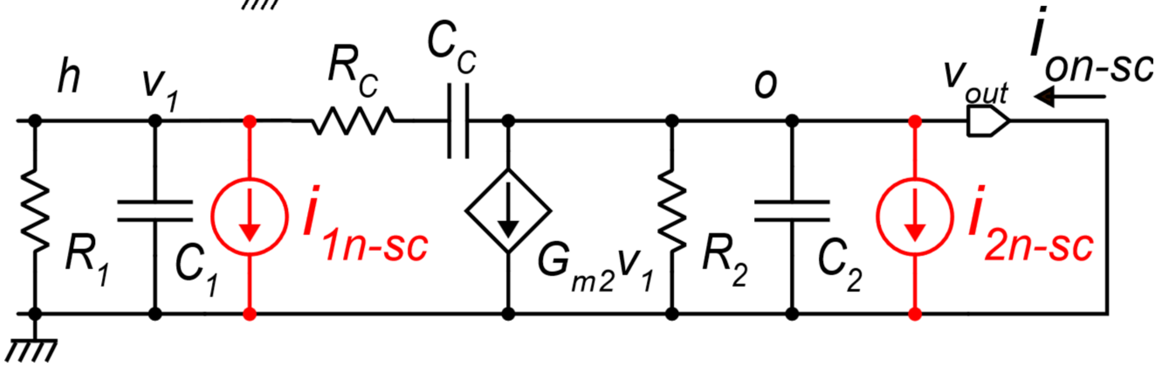


qualitative frequency dependence

A_{Ih} drops below 1 (0 dB) around the unity-gain frequency of the op-amp:

Let us come back to the noise





Circuit for calculation of the total output noise current

$$Y_m = \frac{i_{o-sc}}{v_{id}} = G_{m1} A_{Ih}$$

$$v_n = \frac{-i_{on-sc}}{Y_m}$$

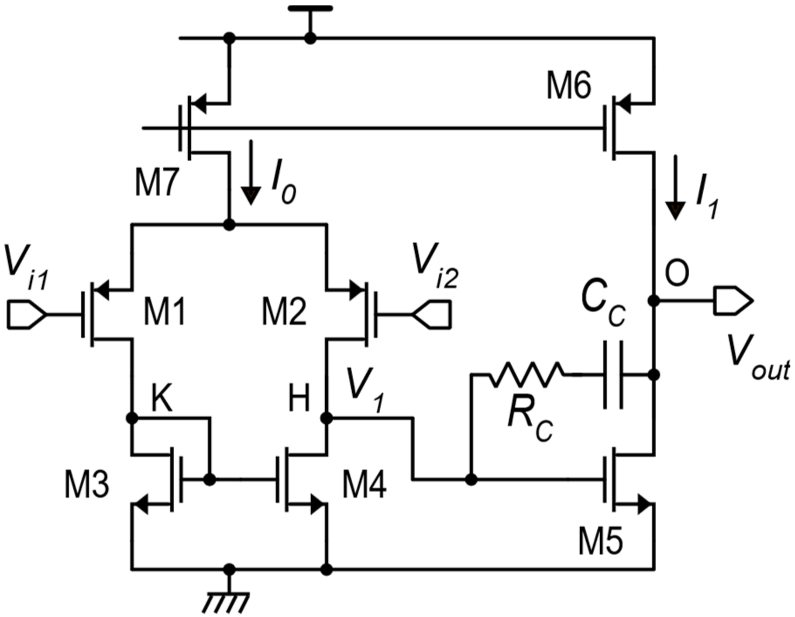
$$i_{on-sc} = i_{2n-sc} + A_{Ih} i_{1n-sc}$$

$$v_n = -\frac{i_{2n-sc}}{G_{m1} A_{ih}} - \frac{i_{1n-sc}}{G_{m1}}$$

$$v_n = -\frac{1}{G_{m1}} \left(i_{1n-sc} + \frac{i_{2n-sc}}{A_{ih}} \right)$$

up to frequencies where $\left| A_{ih} \right| >> 1 \Rightarrow v_n \cong \frac{-i_{1n-sc}}{G_{m1}}$

In the simple 2-stage op-amp



$$v_n \cong \frac{-i_{1n-sc}}{G_{m1}} = \frac{i_{n1} - i_{n2} + i_{n3} - i_{n4}}{g_{m1}}$$

$$S_{vn}\left(f\right) = \frac{S_{I1} + S_{I2} + S_{I3} + S_{I4}}{g_{m1}^2}$$

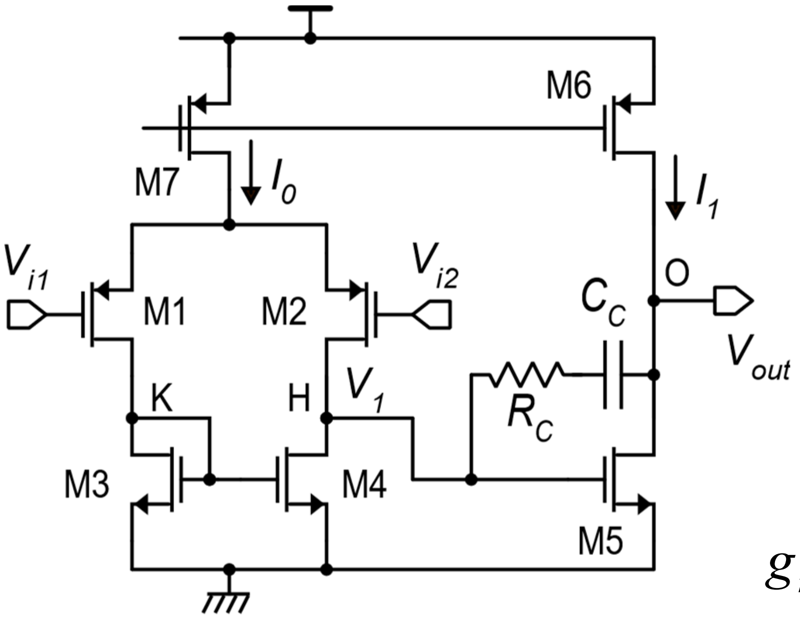
$$S_{vn}\left(f\right) = 2\frac{S_{I1} + S_{I3}}{g_{m1}^2}$$

A more interesting formula can be obtained substituting the equivalent gate noise voltage

$$S_{In} = g_m^2 S_{Vn}$$

$$S_{vn}\left(f\right) = 2\frac{g_{m1}^2 S_{v1} + g_{m3}^2 S_{v3}}{g_{m1}^2} = 2\left(S_{v1} + \frac{g_{m3}^2}{g_{m1}^2} S_{v3}\right)$$

Input noise density of the op-amp



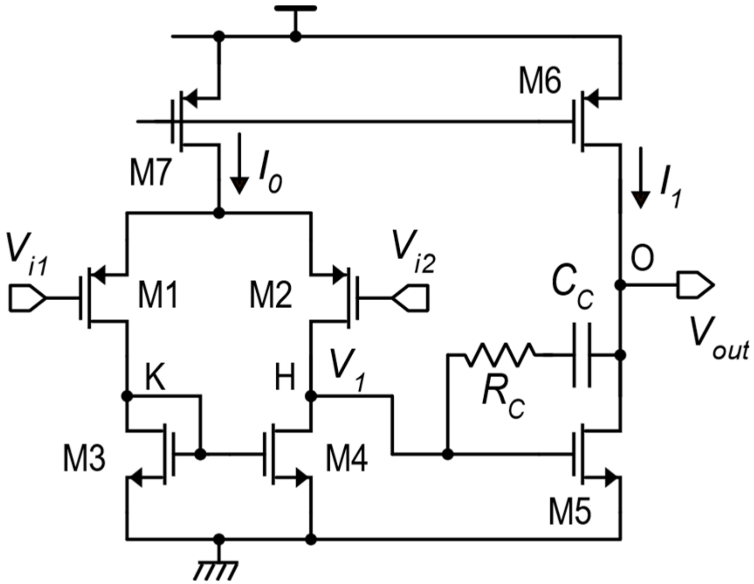
$$S_{vn}(f) = 2\left(S_{v1} + \frac{g_{m3}^2}{g_{m1}^2} S_{v3}\right)$$

$$S_{vn}(f) = 2\left(S_{v1} + F^2 S_{v3}\right) \qquad F = \frac{g_{m3}}{g_{m1}}$$

$$g_m = \frac{I_D}{V_{TE}} \Rightarrow F = \frac{I_{D3}}{V_{TE3}} \frac{V_{TE1}}{I_{D1}}$$

$$F = \frac{I_{D3}}{I_{D1}} \frac{V_{TE1}}{V_{TE3}}$$

For this amplifier $I_{D3}=I_{D1}$, then: $F = \frac{V_{TE1}}{V_{TE3}}$



Thermal noise

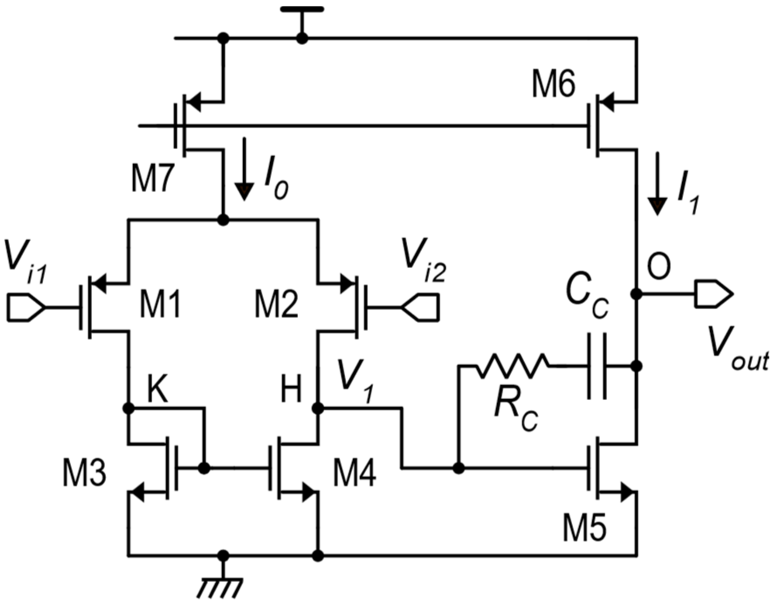
$$S_{vn}(f) = 2\left(S_{v1} + F^2 S_{v3}\right)$$

using: $S_v(f) = \frac{8}{3} kT \frac{1}{g_m}$

$$S_{vn-th}(f) = 2\left(\frac{8}{3} kT \frac{1}{g_{m1}} + F^2 \frac{8}{3} kT \frac{1}{g_{m3}}\right)$$

$$S_{vn-th}(f) = 2 \cdot \frac{8}{3} kT \frac{1}{g_{m1}}\left(1 + F^2 \frac{g_{m1}}{g_{m3}}\right) \qquad F = \frac{g_{m3}}{g_{m1}}$$

$$S_{vn-th}(f) = 2 \cdot \frac{8}{3} kT \frac{1}{g_{m1}}(1 + F)$$



Flicker noise

$$S_{vn}(f) = 2\left(S_{v1} + F^2 S_{v3}\right)$$

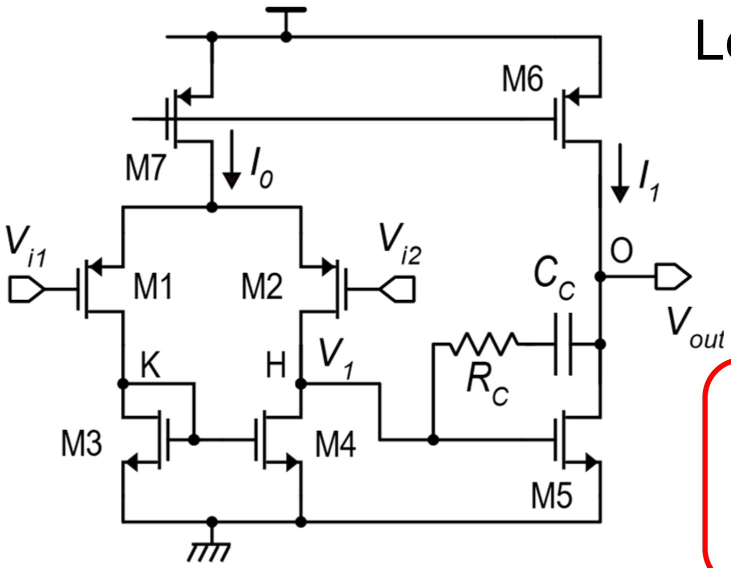
using: $S_v(f) = \dfrac{N_f}{WL}\dfrac{1}{f}$

with: p-MOS $\Rightarrow N_{fp}$

n-MOS $\Rightarrow N_{fn}$

$$S_{vn-F}(f) = 2\left(\frac{N_{fp}}{W_1L_1} + F^2\frac{N_{fn}}{W_3L_3}\right)\frac{1}{f}$$

General considerations about the op-amp noise:



Let us recall the thermal noise and use: $g_{m1}=I_{D1}/V_{TE1}$)

$$S_{vn-th}(f) = 2\cdot\frac{8}{3}kT\frac{1}{g_{m1}}(1+F) = 2\cdot\frac{8}{3}kT\frac{V_{TE1}}{I_{D1}}(1+F)$$

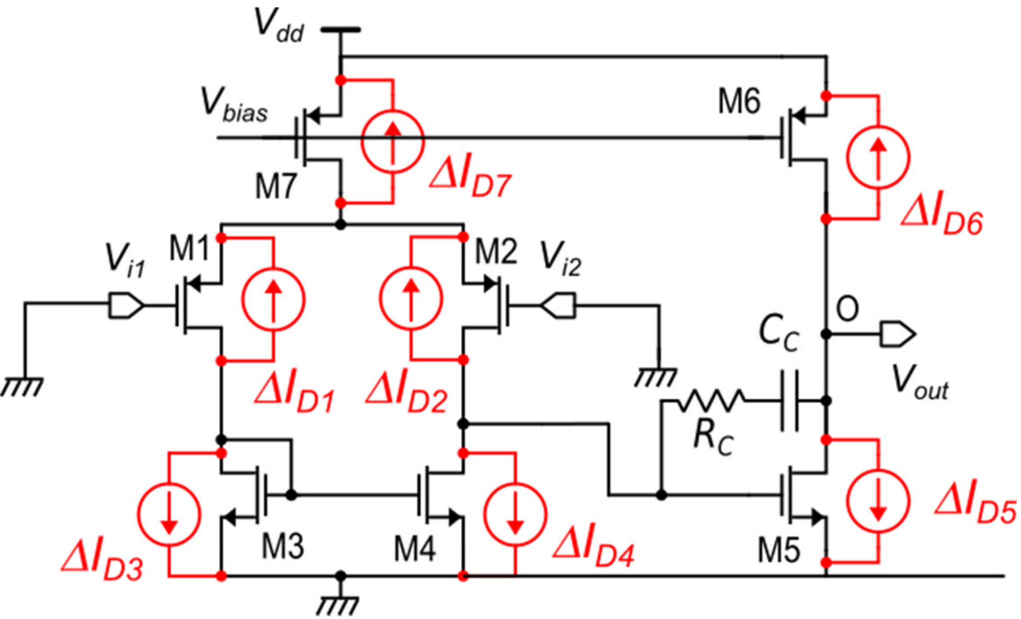
Thermal

$$S_{vn-F}(f) = 2\left(\frac{N_{fp}}{W_1L_1} + F^2\frac{N_{fn}}{W_3L_3}\right)\frac{1}{f}$$

Flicker

- For both the thermal and flicker noise, it is convenient to set F<<1 (V_{TE1}<<V_{TE3})
- The larger I_{D1}, the lower the input thermal noise voltage density
- A small V_{TE1} helps obtaining small thermal noise densities with lower current
- A small flicker noise density can be obtained using large M1 and M3 areas

Input offset voltage of the op-amp



Op-amp with the equivalent current sources that takes into account parameter variations

$$v_n \cong \frac{-i_{1n-sc}}{G_{m1}} = \frac{i_{n1} - i_{n2} + i_{n3} - i_{n4}}{g_{m1}}$$

Let us just replace the noise current sources with the equivalent current sources of parameter variations

$$v_{io} \cong \frac{\Delta I_{D1} - \Delta I_{D2} + \Delta I_{D3} - \Delta I_{D4}}{g_{m1}}$$

Note that M1,M2 and M3,M4 form pairs of matched devices.

Then, we can group their parameter variation sources into single contributions that contain only matching errors

$$v_{io} \cong \frac{\Delta I_{D1,2} + \Delta I_{D3,4}}{g_{m1}}$$

Input offset voltage of the op-amp

$$v_{io} \cong \frac{\Delta I_{D1,2} + \Delta I_{D3,4}}{g_{m1}} = \frac{I_{D1}\left[\frac{\Delta\beta_{1,2}}{\beta_1} - \frac{2\Delta V_{t1,2}}{\left|V_{GS}-V_t\right|_1}\right] + I_{D3}\left[\frac{\Delta\beta_{3,4}}{\beta_3} - \frac{2\Delta V_{t3,4}}{\left(V_{GS}-V_t\right)_3}\right]}{g_{m1}}$$

$$v_{io} \cong \frac{I_{D1}}{g_{m1}}\left\{\left[\frac{\Delta\beta_{1,2}}{\beta_1} - \frac{2\Delta V_{t1,2}}{\left|V_{GS}-V_t\right|_1}\right] + \frac{I_{D3}}{I_{D1}}\left[\frac{\Delta\beta_{3,4}}{\beta_3} - \frac{2\Delta V_{t3,4}}{\left(V_{GS}-V_t\right)_3}\right]\right\}$$

in strong inversion

$$\frac{I_{D1}}{g_{m1}} = V_{TE1} = \frac{\left|V_{GS}-V_t\right|_1}{2}$$

$$v_{io} \cong \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{1,2}}{\beta_1} - \Delta V_{t1,2} + $$

$$+ \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{I_{D3}}{I_{D1}}\frac{\Delta\beta_{3,4}}{\beta_3} - \frac{I_{D3}}{I_{D1}}\frac{\left|V_{GS}-V_t\right|_1}{\left(V_{GS}-V_t\right)_3}\Delta V_{t3,4}$$

Input offset voltage of the op-amp

$$v_{io} \cong \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{1,2}}{\beta_1}-\Delta V_{t1,2}+\frac{\left|V_{GS}-V_t\right|_1}{2}\frac{I_{D3}}{I_{D1}}\frac{\Delta\beta_{3,4}}{\beta_3}-\underbrace{\frac{I_{D3}}{I_{D1}}\frac{\left|V_{GS}-V_t\right|_1}{\left(V_{GS}-V_t\right)_3}}_{F}\Delta V_{t3,4}$$

$$F=\frac{g_{m3}}{g_{m1}}=\frac{I_{D3}}{I_{D1}}\frac{V_{TE1}}{V_{TE3}}$$

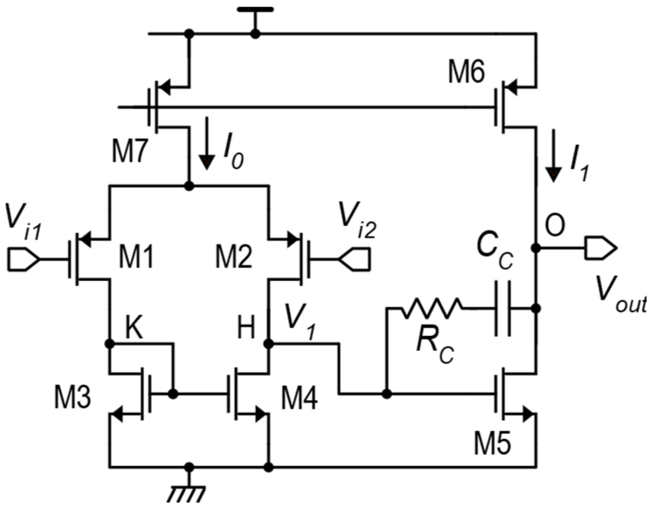
in strong inversion: $V_{TE1}=\frac{\left|V_{GS}-V_t\right|_1}{2}$, $V_{TE3}=\frac{\left(V_{GS}-V_t\right)_3}{2}$

$$v_{io} \cong \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{1,2}}{\beta_1}-\Delta V_{t1,2}+\left(\frac{I_{D3}}{I_{D1}}\right)\frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{3,4}}{\beta_3}-F\Delta V_{t3,4}$$

Contribution of the input pair devices (first two terms)

Contribution of the mirror devices (last two terms)

Input offset voltage of the op-amp: standard deviation



$$v_{io} \cong \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{1,2}}{\beta_1} - \Delta V_{t1,2} + \left(\frac{I_{D3}}{I_{D1}}\right)\frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{3,4}}{\beta_3} - F\Delta V_{t3,4}$$

$$I_{D1} = I_{D3}$$

$$v_{io} \cong \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{1,2}}{\beta_1} - \Delta V_{t1,2} + \frac{\left|V_{GS}-V_t\right|_1}{2}\frac{\Delta\beta_{3,4}}{\beta_3} - F\Delta V_{t3,4}$$

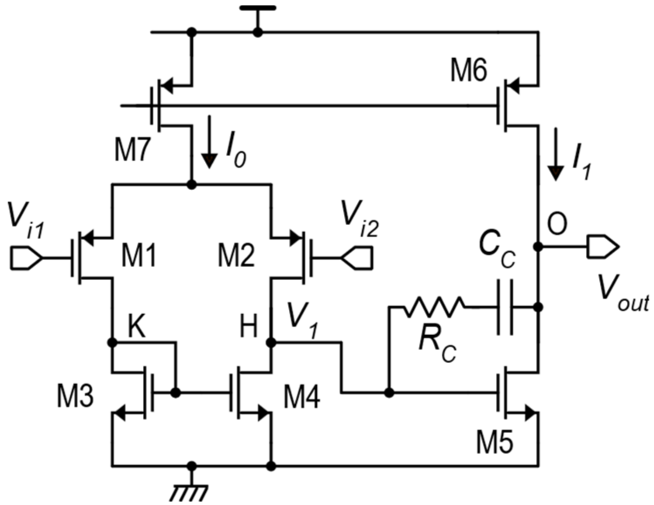
$$\sigma_{vio}^2 \cong \frac{\left(V_{GS}-V_t\right)_1^2}{4}\sigma^2_{\frac{\Delta\beta_{1,2}}{\beta_1}} + \sigma^2_{\Delta V_{t1,2}} + \frac{\left(V_{GS}-V_t\right)_1^2}{4}\sigma^2_{\frac{\Delta\beta_{3,4}}{\beta_3}} + F^2\sigma^2_{\Delta V_{t3,4}}$$

$$\sigma_{\Delta V_t} = \frac{C_{Vt}}{\sqrt{WL}}$$

$$\sigma_{\frac{\Delta\beta}{\beta}} = \frac{C_\beta}{\sqrt{WL}}$$

$$\sigma_{vio}^2 \cong \frac{\left(V_{GS}-V_t\right)_1^2}{4}\frac{C^2_{\beta p}}{W_1L_1} + \frac{C^2_{Vtp}}{W_1L_1} + \frac{\left(V_{GS}-V_t\right)_1^2}{4}\frac{C^2_{\beta n}}{W_3L_3} + F^2\frac{C^2_{Vtn}}{W_3L_3}$$

Design for input offset voltage



$$\sigma_{vio}^2 \cong \frac{\left(V_{GS}-V_t\right)_1^2}{4}\frac{C_{\beta p}^2}{W_1 L_1}+\frac{C_{Vtp}^2}{W_1 L_1}+\frac{\left(V_{GS}-V_t\right)_1^2}{4}\frac{C_{\beta n}^2}{W_3 L_3}+F^2\frac{C_{Vtn}^2}{W_3 L_3}$$

$$\sigma_{vio}^2 \cong \frac{A}{W_1 L_1}+\frac{B}{W_3 L_3}$$

$$A=\frac{\left(V_{GS}-V_t\right)_1^2}{4}C_{\beta p}^2+C_{Vtp}^2 \; ; \; B=\frac{\left(V_{GS}-V_t\right)_1^2}{4}C_{\beta n}^2+F^2 C_{Vtn}^2$$

Total gate area of the input pair and mirror:

$$S=2\left(W_1 L_1+W_3 L_3\right)$$

Offset voltage: area optimization procedure

$$\sigma_{vio}^2 \cong \frac{A}{W_1L_1} + \frac{B}{W_3L_3}$$

$$W_3L_3 = a \cdot W_1L_1$$

Example: if *a*=1, we are assigning the same area to the input pair and to the mirror

$$S = 2\left(W_1L_1 + W_3L_3\right)$$

Optimization problem: find the value of *a* that allows obtaining the required σ_{vio} with the minimum area occupation.

$$\sigma_{vio}^2 \cong \frac{A}{W_1L_1} + \frac{B}{aW_1L_1} = \frac{1}{W_1L_1}\left(A + \frac{B}{a}\right)$$

$$S = 2\left(W_1L_1 + aW_1L_1\right) = 2W_1L_1\left(1+a\right)$$

$$W_1L_1 = \frac{1}{\sigma_{vio}^2}\left(A + \frac{B}{a}\right)$$

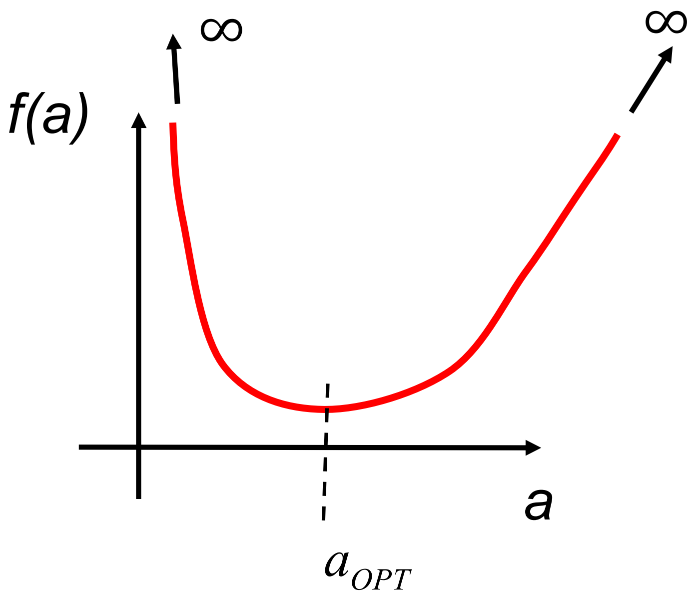
$$S = 2\frac{1}{\sigma_{vio}^2}\left(A + \frac{B}{a}\right)\left(1+a\right)$$

We need to find the minimum of this function of *a*

Offset voltage: area optimization procedure

$$S = 2\frac{1}{\sigma_{vio}^2}\left(A+\frac{B}{a}\right)(1+a) = 2\frac{1}{\sigma_{vio}^2}\left(A+\underbrace{aA+\frac{B}{a}}+B\right)$$

$$\sigma_{vio}^2 \cong \frac{A}{W_1L_1}+\frac{B}{W_3L_3}$$



only these two terms depend on a. then we have to find the minimum of:

$$f(a) = aA+\frac{B}{a} \qquad \frac{df(a)}{da} = A-\frac{B}{a^2} = 0$$

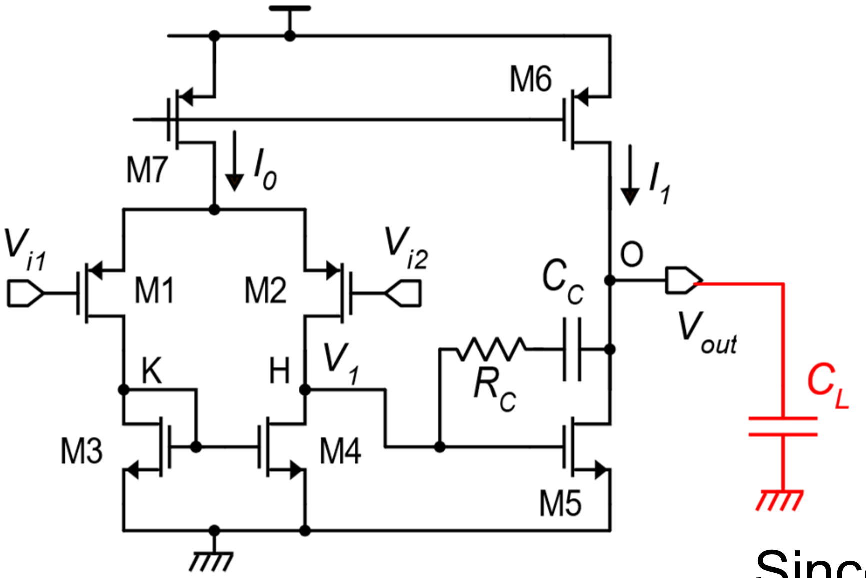
$$a_{OPT} = \sqrt{\frac{B}{A}} \qquad (a>0)$$

$$\begin{cases} W_1L_1 = \dfrac{1}{\sigma_{vio}^2}\left(A+\dfrac{B}{a_{OPT}}\right) \\ W_3L_3 = a\cdot W_1L_1 \end{cases}$$

Current consumption of the op-amp

- In this section, we will consider the main factors that affect the current consumption of the operational amplifier.
- We have already found an expression that ties the current consumption with the GBW specification
- Here, we will review that expression, introducing also the role of the "*F*" parameter that comes from the noise and offset analysis
- After that, we will find an expression of the current consumption that highlights the relationship with the thermal noise specification

GBW and supply current (from the *GBW* and φ_m design procedure)



$$I_{supply} = 2\pi\sigma \cdot GBW \cdot C_L \cdot \left(V_{TE5} + 2\frac{g_{m1}}{g_{m5}} V_{TE1} \right)$$

$$I_{supply} = 2\pi\sigma \cdot GBW \cdot C_L \cdot V_{TE5} \left(1 + 2\frac{g_{m1}}{g_{m5}} \frac{V_{TE1}}{V_{TE5}} \right)$$

define: $\frac{g_{m1}}{g_{m5}} = \frac{1}{\sigma}\frac{C_C}{C_L} = r_{gm}$

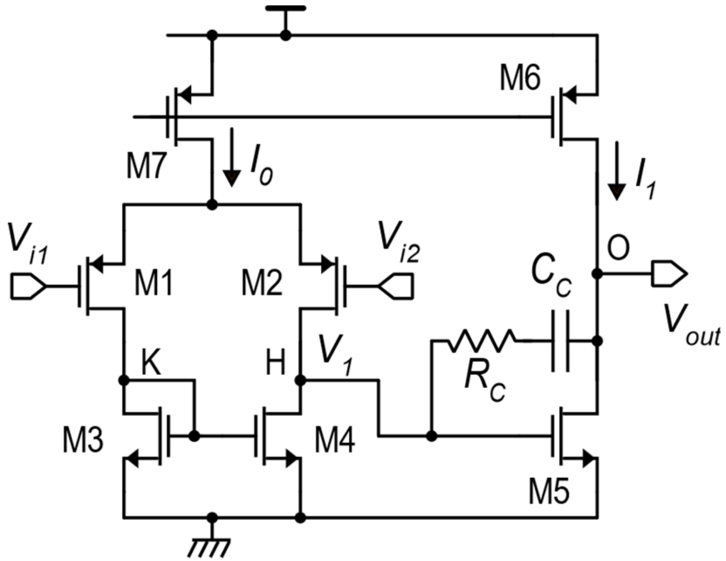
Since $V_{GS5}=V_{GS3}$ $\quad V_{TE5} = V_{TE3}$ $\quad \frac{V_{TE1}}{V_{TE5}} = \frac{V_{TE1}}{V_{TE3}} = F$

$$I_{supply} = 2\pi\sigma \cdot GBW \cdot C_L \cdot V_{TE5} \left(1 + 2Fr_{gm} \right)$$

Fraction due to the second stage

Fraction due to the first stage

Current consumption of the op-amp - role of the thermal noise spec



If the "dominant" specification is the thermal noise PSD (S_{Vn-th}), we can use a different expression

Let us start again from the general formula:

$$I_{supply} = 2g_{m1}V_{TE1} + g_{m5}V_{TE5}$$

We need to highlight the role of g_{m1}

... and consider the expression of the input thermal noise voltage PSD:

$$S_{vn-th} = 2 \cdot \frac{8}{3} kT \frac{1}{g_{m1}} (1+F) \Rightarrow g_{m1} = 2 \cdot \frac{8}{3} kT \frac{1}{S_{vn-th}} (1+F)$$

Current consumption of the op-amp

$$I_{supply} = 2g_{m1}V_{TE1} + g_{m5}V_{TE5} = 2g_{m1}V_{TE1}\left(1+\frac{1}{2}\frac{g_{m5}}{g_{m1}}\frac{V_{TE5}}{V_{TE1}}\right) = 2g_{m1}V_{TE1}\left(1+\frac{1}{2r_{gm}F}\right)$$

$$\frac{1}{r_{gm}} \qquad \frac{1}{F}$$

$$g_{m1} = 2\cdot\frac{8}{3}kT\frac{1}{S_{vn-th}}(1+F)$$

$$I_{supply} = 2\cdot 2\cdot\frac{8}{3}kT\frac{1}{S_{vn-th}}(1+F)V_{TE1}\left(1+\frac{1}{2r_{gm}F}\right)$$

Note: F and 1/F appear: an optimum F value can be calculated

Note: $I_{supply} \propto \frac{1}{S_{vn-th}}$

low thermal noise (S_{vn-th}) means high current consumption

$$I_{supply} = \frac{32}{3}kT(1+F)\frac{V_{TE1}}{S_{vn-th}}\left(1+\frac{1}{2r_{gm}F}\right)$$

Fraction due to the first stage

Fraction due to the second stage

Examples

Case 1: $GBW = 10\,\text{MHz},\ C_{L-\max} = 10\,\text{pF},\ \sigma = 3\ V_{TE5} = 150\ \text{mV},\ V_{TE1} = 50\,\text{mV},\ \text{C}_C = C_L$

$$F = \frac{V_{TE1}}{V_{TE3}} = \frac{V_{TE1}}{V_{TE5}} = \frac{1}{3} \qquad r_{gm} = \frac{g_{m1}}{g_{m5}} = \frac{1}{\sigma}\frac{C_C}{C_L} = \frac{1}{3} \qquad I_{supply} = 2\pi\sigma \cdot GBW \cdot C_L \cdot V_{TE5}\left(1 + 2Fr_{gm}\right)$$

$$I_{supply} = 2\pi\sigma \cdot GBW \cdot C_L \cdot V_{TE5}\left(1 + \frac{2}{9}\right) = 6.28 \cdot 3 \cdot 10\times10^{6} \cdot 10\times10^{-12} \cdot 0.15 \cdot \frac{11}{9} = 345\ \mu\text{A}$$

Case 2: as above, but the GBW specification is replaced by noise specs:

$$\sqrt{S_{vn-th}} = 1\ \text{nV}/\sqrt{\text{Hz}} \ \Rightarrow S_{vn-th} = 10^{-18}\ \ \text{V}^2/\text{Hz}$$

$$I_{supply} = \frac{32}{3}kT\left(1+F\right)\frac{V_{TE1}}{S_{vn-th}}\left(1 + \frac{1}{2r_{gm}F}\right) \qquad = \frac{32}{3}4\times10^{-21}\,\frac{4}{3}\,\frac{0.05}{10^{-18}}\left(1+\frac{9}{2}\right) = 15.6\ \text{mA}$$

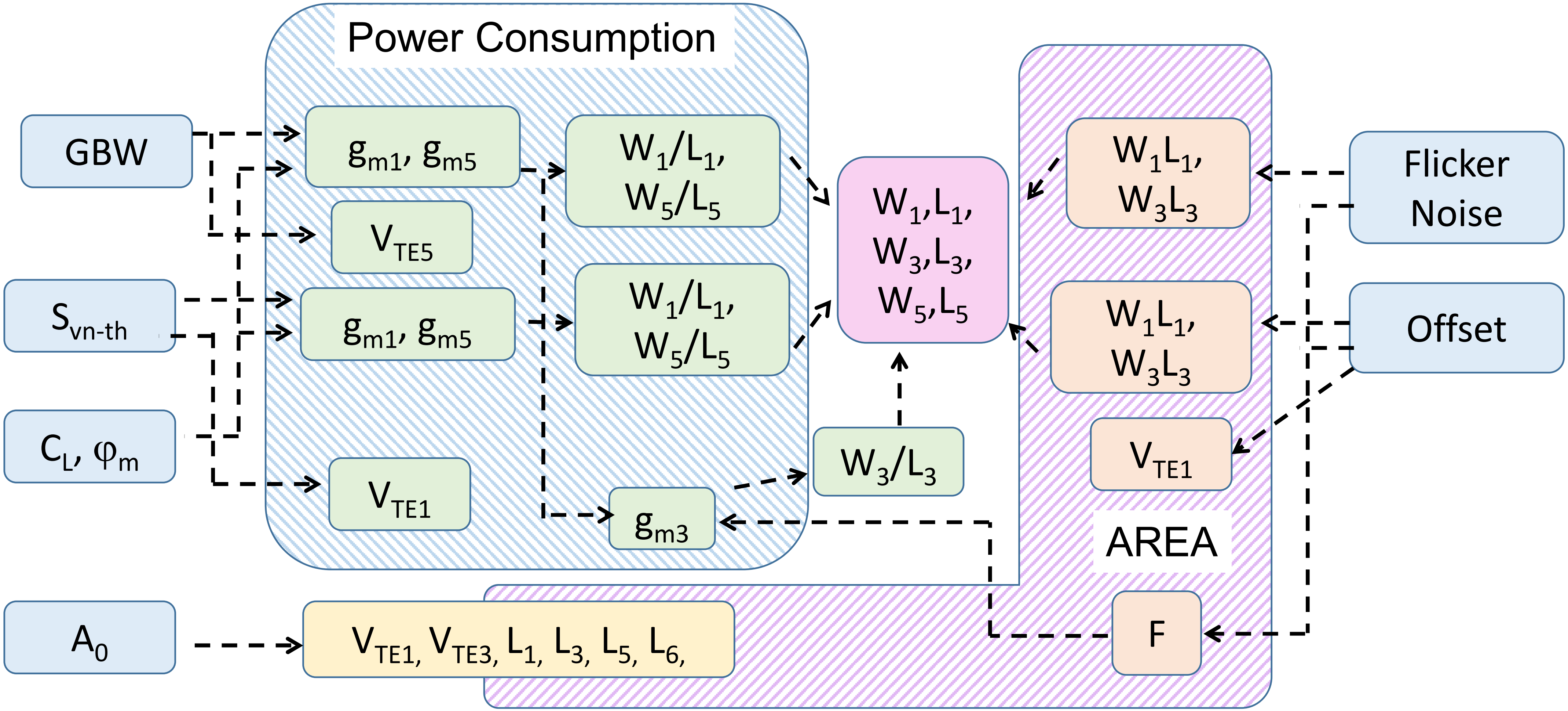
Consistent and contrasting specifications

As we have seen in previous example, the GBW specification and thermal noise density specification are consistent, since for both the following rule holds: the stricter the specification, the higher the required supply-current.

If the design include both specifications, then one of them is likely to be dominant. In the previous example, the noise specification dominates: the minimum supply current required to meet the required noise specification is much larger the current required for the given GBW-C_L combination. Then, if we design the amplifier for the noise density, we certainly meet the GBW requirement.

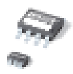
Other specification pairs are likely to be contrasting: thermal noise and speed are in contrast with the supply current specification. The same can be said about flicker noise and area.

Action of various specifications:



Commercial products: high speed - low thermal noise CMOS op-amp

Burr-Brown Products
from Texas Instruments



OPA300, OPA2300
OPA301, OPA2301

SBOS271D – MAY 2003 – REVISED JUNE 2007

Low-Noise, High-Speed, 16-Bit Accurate, CMOS OPERATIONAL AMPLIFIER

FEATURES

- High Bandwidth: 150MHz
- 16-Bit Settling in 150ns
- Low Noise: 3nV/√Hz
- Low Distortion: 0.003%
- Low Power: 9.5mA (typ) on 5.5V ← low power?
- Shutdown to 5μA
- Unity-Gain Stable
- Excellent Output Swing: (V+) – 100mV to (V–) + 100mV
- Single Supply: +2.7V to +5.5V
- Tiny Packages: MSOP and SOT23

APPLICATIONS

- 16-Bit ADC Input Drivers
- Low-Noise Preamplifiers
- IF/RF Amplifiers
- Active Filtering

DESCRIPTION

The OPA300 and OPA301 series high-speed, voltage-feedback, CMOS operational amplifiers are designed for 16-bit resolution systems. The OPA300/OPA301 series are unity-gain stable and feature excellent settling and harmonic distortion specifications. Low power applications benefit from low quiescent current. The OPA300 and OPA2300 feature a digital shutdown (Enable) function to provide additional power savings during idle periods. Optimized for single-supply operation, the OPA300/OPA301 series offer superior output swing and excellent common-mode range.

The OPA300 and OPA301 series op amps have 150MHz of unity-gain bandwidth, low 3nV/√Hz voltage noise, and 0.1% settling within 30ns. Single-supply operation from 2.7V (±1.35V) to 5.5V (±2.75V) and an available shutdown function that reduces supply current to 5μA are useful for portable low-power applications. The OPA300 and OPA301 are available in

Commercial product: low power op-amp

ANALOG DEVICES

1 µA Micropower CMOS Operational Amplifiers

AD8502/AD8504

FEATURES

- **Supply current: 1 µA maximum/amplifier**
- **Offset voltage: 3 mV maximum**
- **Single-supply or dual-supply operation**
- **Rail-to-rail input and output**
- **No phase reversal**
- **Unity gain stable**

PIN CONFIGURATIONS

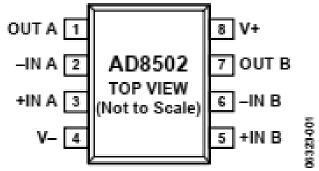


Figure 1. 8-Lead SOT-23

		−40°C < T_A < +125°C			2	µA
DYNAMIC PERFORMANCE						
Slew Rate	SR	R_{LOAD} = 1 MΩ		0.004		V/µs
Gain Bandwidth Product	GBP			7		kHz
Phase Margin	Øo			60		Degrees

Parameter	Symbol	Conditions	Min	Typ	Max	Unit
NOISE PERFORMANCE						
Peak-to-Peak Noise		0.1 Hz to 10 Hz		6		µV p-p
Voltage Noise Density	e_n	f = 1 kHz		190		nV/√Hz
Current Noise Density	i_n	f = 1 kHz		0.1		pA/√Hz