OPERATIONAL AMPLIFIERS

ECEN-489

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Differential Pair

If both transistors are saturated

\[ i_{d1} + i_{d2} = IB \]
\[ i_{d1} = \frac{\mu_n C_{OX}}{2} \frac{W}{L} (V_{gs1} - V_T)^2 \]
\[ i_{d2} = \frac{\mu_n C_{OX}}{2} \frac{W}{L} (V_{gs2} - V_T)^2 \]

Solving these equations =>

\[ i_{d1} = \frac{IB}{2} + g_{ml} \frac{(v_1 - v_2)}{2} \left( 1 + \sqrt{1 - \frac{(v_1 - v_2)^2}{2V_{DSAT1}}} \right) \]
\[ i_{d2} = \frac{IB}{2} - g_{ml} \frac{(v_1 - v_2)}{2} \sqrt{1 - \frac{(v_1 - v_2)^2}{2V_{DSAT1}}} \]
Differential Pair

Is a non-linear circuit

\[ i_{d1} - i_{d2} = g_m (v_1 - v_2) \sqrt{1 - \left( \frac{v_1 - v_2}{2V_{DSAT1}} \right)^2} \]

For \( v_d = v_1 - v_2 < V_{DSAT1} \implies \]

\[ i_{d1} - i_{d2} = \sqrt{\mu_n C_{OX} \frac{W}{L} I_B (v_1 - v_2)} \]

Note:
Linear range increases for large \( V_{DSAT1} \)
\( V_{GS} \) is also increased (limited by VSS)
**Basic Operational Transconductance Amplifier**

**DESIGN CONSIDERATIONS:**

- \( v_d = v_1 - v_2 < V_{DSAT} \)
- \( V_{1,2} > V_{GS1} + V_{DSATB} \)

\[
i_{out} = \sqrt{\mu_n C_{OX}} \frac{W}{L} I_B (v_1 - v_2)
\]

or

\[
i_{out} = g_m (v_1 - v_2)
\]

Sensitive to differential signals

\[
v_{out} = g_m r_{out} (v_1 - v_2)
\]

\[
r_{out} = r_{o1} II r_{op}
\]

\( i_{out} = 0 \) for \( v_1 = v_2 \) \( \Rightarrow \) rejection to common-mode signals
OTA based on 3 current mirrors

Differential pair + current gain $g_m$

Output conductance $g_{out}$

DC-gain

Gain – BW Product (GBW)

Slew – rate

NOISE: All transistors are important !!!!
OTA based on 3 current mirrors

Differential pair + current gain  \( g_{m} \)

Output conductance  \( g_{out} \)

DC-gain  \( A_{V-DC} \)

\[
B \sqrt{\frac{W}{L}} I_{TAIL}
\]

\[
\frac{B}{2} \left( \frac{I_{TAIL}}{V_{AN} L_{N}} + \frac{I_{TAIL}}{V_{AP} L_{P}} \right)
\]

\[
2 \sqrt{\frac{W}{L}} I_{TAIL} \left( \frac{1}{V_{AN} L_{N}} + \frac{1}{V_{AP} L_{P}} \right)
\]

\[
\frac{g_{out}}{C_{out}}
\]

\[
\frac{g_{mp}}{(1+B)C_{GS-P}}
\]

\[
B \sqrt{\frac{W}{L}} I_{TAIL}
\]

\[
C_{out}
\]

\[
\frac{BI_{TAIL}}{C_{out}}
\]

\[
\left[ B^2 g_{m1} + B^2 g_{mp1} + g_{mpB} + 1 g_{mn} \right]
\]

\[
\frac{16kT}{3} \left[ \frac{g_{mp1}}{g_{m1}} \left( 1 + \frac{1}{B} \right) + \frac{g_{mn}}{g_{m1}} \left( \frac{1}{B^2} \right) \right]
\]

Current=(1+B)I_{TAIL}

\[
\frac{i_{0,eq}^2}{3} = \frac{16kT}{3} \left[ B^2 g_{m1} + B^2 g_{mp1} + g_{mpB} + 1 g_{mn} \right]
\]

\[
\frac{v_{eq}^2}{3g_{m1}} = \frac{16kT}{3g_{m1}} \left[ \frac{g_{mp1}}{g_{m1}} \left( 1 + \frac{1}{B} \right) + \frac{g_{mn}}{g_{m1}} \left( \frac{1}{B^2} \right) \right]
\]
OTA based on 3 current mirrors using cascode transistors

Differential pair + current gain  \( g_m \)

Output conductance  \( g_{out} \)

DC-gain  \( A_{V-DC} \)

\[ BI_{TAIL} \]  

\[ \frac{V_B}{V_P} \]

\[ i_o \]

\[ v_o \]

Current = \((1+B)I_{TAIL}\)

\[ g_m \equiv \frac{16kT}{3} \left[ B^2 g_{m1} + B^2 g_{mp1} + g_{mpB} + 2g_{mn} \right] \]

\[ v_{eq}^2 \equiv \frac{16kT}{3g_{m1}} \left[ \frac{g_{mp1}(1 + \frac{1}{B})}{g_{m1}} + \frac{g_{mn} \left( \frac{2}{B^2} \right)}{g_{m1}} \right] \]
OTA based on 3 current mirrors using cascode transistors

\[
A_V \approx \frac{B g_{m1} R_{out}}{2} \left[ \frac{1}{1 + s R_{out} C_{out}} \right] \left[ \frac{1}{1 + s (1 + B) C_{GSP}} \right]^2 \left( \frac{1}{1 + s (2 + \Delta) C_{GSN}} \right) + \left( \frac{1}{1 + s (1 + \Delta) C_{GSP}} \right) \left( \frac{1}{1 + s (1 + \delta) C_{GSN}} \right)
\]

Phase Margin is limited

OTA-output

\[
\text{B : 1} \quad 1 : \text{B}
\]

\[
\begin{align*}
\text{P-type current mirror} & \quad \text{P-type cascode} \\
\text{N-type current mirror & cascode}
\end{align*}
\]

\[
\frac{B I_{TAIL}}{2}
\]

\[
\text{MN}
\]

\[
\text{OTA-CAS}
\]

\[
\text{OTA}
\]

\[
w
\]

\[
\begin{align*}
\text{OTA-output}
\end{align*}
\]
OTA based on 3 current mirrors using cascode transistors

Excess Phase is defined as (phase at 0 - phase at $\omega_u$)

Phase Margin = (180 – excess phase)

Gain margin = Gain measured at 180° excess phase
FOLDED-CASCODE OTA

Frequency response:

Right hand side

- Right hand side: 2 poles (at $V_W$ and $V_{out}$)
- Left hand side: 4 poles !!! (at $V_X$, $V_Y$, $V_Z$ and $V_{out}$)

- The poles at $V_Y$ and $V_Z$ are associated to N-type transistors $\rightarrow$ higher frequencies

$$A_V(s) \approx g_{m1}R_{out} \cdot \frac{1}{sC_{out}} \cdot \frac{1}{1+\frac{sC_{PC}}{g_{mp}}}$$

Noise: $i_{out}^2 = 2(i_{eq1}^2 + i_{eqp}^2 + i_{eqn}^2)$
Output referred noise

- M1 produces an output current given by
  \[ i_{01} = g_{m1} v_{n1} \]

- Each transistor M2 generates a differential output current
  \[ i_{02} = g_{m2} v_{n2} \]

- Similarly, for each transistor M5
  \[ i_{05} = g_{m5} v_{n5} \]

- At low and medium frequencies, noise contribution of the cascode transistors can be neglected (M3 and M4)
  \[ i_{\text{out}}^2 = 2(i_{\text{eq1}}^2 + i_{\text{eq2}}^2 + i_{\text{eqn}}^2) \]
Two-Stage Miller Amplifier

- Good for low-voltage applications
- Large gain can be achieved
- Low and Medium frequency applications
- Stability is an issue

Main equations: First order approximations ignoring CGD

\[ A_{VDC} = -\frac{g_{m1} g_{m3}}{g_1 g_L} \]

\[ \omega_{p1} = \frac{g_1}{C_1} \]

\[ \omega_{p2} = \frac{g_L}{C_L} \]

\[ GBW = (A_{VDC}) \cdot \min(\omega_{p1}, \omega_{p2}) \]

\[ \text{Phase margin} = 180 - \tan^{-1}\left( \frac{\omega_u}{\omega_{p1}} \right) - \tan^{-1}\left( \frac{\omega_u}{\omega_{p2}} \right) \]
GENERAL CIRCUIT

\[
\begin{bmatrix}
q_x + s(c_1 + c_3)
g_1 + s(c_1 + c_3)
\end{bmatrix}
\begin{bmatrix}
C_0
\end{bmatrix}
= \begin{bmatrix}
0
\end{bmatrix}
\]

\[
\frac{dx}{dt} = \begin{bmatrix}
0
\end{bmatrix}
\]

\[
\frac{\sigma_0}{x_0} \bigg|_{\omega = \omega_c} = -\frac{g_{in}r_2}{j_0\omega_1 R_2}
\]

\[
\omega_{2\omega_0} = \frac{g_{in}r_2}{C_3}
\]

If \( \omega_1 < \omega_{\omega_0} \)

\[
\omega_{\omega_1} > \frac{1}{R_2(c_2 + c_3) + R_1(c_1 + c_3) + g_{in}r_2 R_2 (R_1 C_3)}
\]

\[
\omega_{sp_1} = \frac{1}{\omega_{p1}} \left( \frac{1}{R_1 R_2 (c_1 (c_2 + c_3) + c_2 c_3)} \right)
\]

\[
\frac{d\omega}{dt} \bigg|_{\omega = \omega_1} = -\frac{g_{in}r_2}{R_2c_3}
\]

\[
\frac{d\omega}{dt} \bigg|_{\omega = \omega_{p2}} = -\frac{40 \text{ dB} / \text{dec}}{-20 \text{ dB} / \text{dec}}
\]
Phase compensation ➔ Pole splitting techniques!!

\[ A_{VDC} = -\frac{g_{m1} g_{m3}}{g_1 g_L} \]

\[ \omega_{p1} \approx \frac{g_1}{C_1 + \frac{g_{m3}}{g_L} C_M} \]

\[ \omega_{p2} = \frac{g_{m2}}{C_1 + C_L} \]

\[ \text{GBW} = A_{VDC} \cdot \omega_{p1} \approx \frac{g_{m1}}{C_M} \]

\[ \text{Phase margin in} = 180 - \tan^{-1}\left(\frac{\omega_u}{\omega_{p1}}\right) - \tan^{-1}\left(\frac{\omega_u}{\omega_{p2}}\right) \]
After compensation

> Phase Margin > 45 degrees
> Bandwidth is reduced!!

Parasitic (bad) RHP zero!!

\[ \omega_{\text{ZERO}} = \frac{g_{m3}}{C_M} \]

\[ A_{\text{VDC}} = -\frac{g_{ml} g_{m3}}{g_1 g_L} \]

\[ \omega_{p1} \approx \frac{g_1}{C_1 + \frac{g_{m3}}{g_L} C_M} \]

\[ \omega_{p2} = \frac{g_{m3}}{C_1 + C_L} \]

\[ \text{GBW} = A_{\text{VDC}} \times \omega_{p1} \approx \frac{g_{ml}}{C_M} \]

\[ \text{Phase margin in} = 180 - \tan^{-1}\left( \frac{\text{GBW'}}{\omega_{p1'}} \right) - \tan^{-1}\left( \frac{\text{GBW'}}{\omega_{p2'}} \right) - \tan^{-1}\left( \frac{\text{GBW'}}{\omega_{\text{ZERO}}} \right) \]
Parasitic (bad) RHP zero!!
Can be catastrophic if close or below \( w_u \! \! \! \! \! \! \) 

\[
\omega_{\text{ZERO}} = \frac{g_{m3}}{C_M}
\]

Phase margin = \( 180 - \tan^{-1}\left(\frac{\omega_u}{\omega_{p1}'}\right) - \tan^{-1}\left(\frac{\omega_u}{\omega_{p2}'}\right) - \tan^{-1}\left(\frac{\omega_u}{\omega_{\text{ZERO}}}\right) \)

After compensation
- Phase Margin > 45 degrees
- Phase is equivalent to having 3 poles below unity gain frequency \( \Rightarrow \) Unstable!

After compensation
- \( \Rightarrow \) Phase Margin < 45 degrees
- \( \Rightarrow \) Bandwidth is reduced!!!

\( \omega_u \)
\( \omega_{p1} \)
\( \omega_{p1}' \)
\( \omega_{p2} \)
\( \omega_{p2}' \)
\( \omega_{\text{ZERO}} \)
**Design Example:**

- \( \text{GBW} = 10 \text{ MHz} \)
- Phase Margin > 45 degrees, all conditions
- \( A_{\text{DC}} > 80 \text{ dB} \)
- Minimize offset voltage
- \( C_{\text{LOAD}} = 0.5-10 \text{ pF} \)
- \( R_{\text{LOAD}} \sim 10 \text{ kOhms} \)
- Input referred noise density < 10 nV/Hz\(^{1/2}\)
- Maximize the output swing

**Additional considerations:**

\[ |V_{\text{ICM}} - \text{VSS}| > V_{\text{GS1}} + V_{\text{DSAT\_current source}} + \Delta \]

\[ V_{\text{GS2}} = V_{\text{GS3}} \quad \text{(low offset)} \]

\[ V_{\text{DSAT2}}, V_{\text{DSAT3}}, V_{\text{DSAT4}} \text{ have to be minimized} \]

\( \text{(max swing)} \)

Can you solve the system for all these conditions??

**BE SURE ALL TRANSISTORS OPERATE IN THE PROPER REGION!!!**
Adding a series resistance

Why the series resistance compensates the effect of the zero?

How this is done?

What is unique here?

Design Trade-offs?
Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase

Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase

Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase

Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase

Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase

Magnitude and Phase Response for the two-stage OPAMP

Frequency

Magnitude
Phase
Magnitude Response for the two-stage OPAMP:

- Without resistive compensation
- With resistive compensation
Phase Response for the two-stage OPAMP:

Without resistive compensation

With resistive compensation
Two-stage OTA

\[ A_{VDC} = -\frac{g_{m1} \cdot g_{m3}}{g_1 \cdot g_L} \]

\[ \omega_{p1} \approx \frac{g_1}{C_1 + \frac{g_{m3}}{g_L} C_M} \]

\[ \omega_{p2} = \frac{g_{m3}}{C_1 + C_L} \]

\[ GBW = A_{VDC} \cdot \omega_{p1} \approx \frac{g_{m1}}{C_M} \]

\[ \text{Phase margin in} = 180 - \tan^{-1}\left(\frac{GBW}{\omega_{p1}}\right) - \tan^{-1}\left(\frac{GBW}{\omega_{p2}}\right) \]

Phase compensation ➔ Pole splitting!!

After compensation
➔ Phase Margin > 45 degrees
➔ Bandwidth is reduced!!
Buffered OTA = Operational Voltage Amplifier (OPAMP)

SIMPLE MACROMODEL:

\[
\begin{align*}
    A_{\text{VDC}} &= (g_m R_1) A_2 \\
    A_2 &\approx 1 - 0.5 \\
    \text{GBW} &\approx \frac{g_m}{C_1} \\
    R_{\text{out}} &= r_0 \approx 1/g_m \quad \text{(1)}
\end{align*}
\]

Notice that load capacitors must satisfy the following condition, otherwise phase margin is not good enough:

\[
\text{GBW} < \frac{1}{r_0 C_L}
\]
OPAMP Characterization (This section is optional)

Main parameters to be measured:

- DC gain \((10^4-10^6) \text{ V/V}\)
- Frequency Limitations
  - Bandwidth (Few Hertz~1kHz)
  - Gain-Bandwidth product (1~100 Mhz)
- Output resistance
- Input Impedance
- Signal Swing
  - Common-mode input range
  - Output swing
- Stability
- DC Offset
- Slew-rate
- CMRR
- PSRR

For this section, see:
CMOS Analog design, Allen & Holberg
If the OPAMP is not precise, how we can design accurate systems?

Answer is **FEEDBACK!!!**

Examples of our daily life
• Can you shave yourself closing your eyes?
• Can you drive your car closing your eyes?
• Can you adjust the supply voltages without a voltage indicator?
• Can you measure (without any equipment) the magnetic field generated by your cellular phone?
• Can you control properly your daily activities without feedback?

• **If you measure the output at the time you apply the stimuli you can better control the system!!**
Typical values for low-frequency opamps:
- $A_V \sim 10^5$
- $\omega_p \sim 100$ rad/sec
- $r_0 \sim < 100$ Ohms

If $A_V \sim \infty$ then
- $v_- = 0$  VIRTUAL GROUND
- $i_1 = -i_2$
- $v_i/Z_1 = -v_0/Z_2$
If $A_V$ is finite then $v_0 \neq 0$

$$i_1 = -i_2 \quad \text{if} \quad Z_{\text{in}} = \infty$$

or

$$ (v_i - (-v_0/A_V)) Y_1 = (v_0 - (-v_0/A_V)) Y_2$$

$$v_0 = -\frac{Y_1}{Y_2} \left[ \frac{1}{1 + \frac{1}{A_V (1 + \frac{Y_1}{Y_2})}} \right] = -\frac{Z_2}{Z_1} \left[ \frac{1}{1 + \frac{1}{A_V (1 + \frac{Z_2}{Z_1})}} \right]$$

$$\text{Error} = -\frac{1}{A_V} \left( 1 + \frac{Z_2}{Z_1} \right) = -\frac{1 + \frac{s}{\omega_P}}{A_{\text{DC}}} \left( 1 + \frac{Z_2}{Z_1} \right)$$

$A_V = \frac{(g_m R_1) A_2}{1 + \frac{s}{\omega_P}} \approx g_m R_1$

$R_{\text{out}} = 0$ and $Z_{\text{in}} = \infty$
FEEDBACK:

• If you measure the output at the time you apply the stimuli you can better control the system!!

Definitions:

A(s): Amplifier gain (very large but not very well controlled)

B(s): Feedback Factor (Very well controlled)

T(s) = A(s)B(s): Loop Gain (Extremely important parameter!!)

Applying Mason Rule:

\[
\frac{v_o}{v_i} = \frac{\text{Direct trajectory}}{1-\text{loop gain}} = \frac{A(s)}{1-(-T(s))} = \frac{A(s)}{1+A(s)B(s)}
\]
STABILITY: \[
\frac{v_o}{v_i} = \frac{A(s)}{1 - A(s)B(s)}
\]

If:

\[
\frac{v_o}{v_i} = \frac{A_0}{s^2 + BWs + \omega_0^2}
\]

\[
1 - \frac{A_0 B_0}{s^2 + BWs + \omega_0^2}
\]

Real and negative

Imaginary and negative

Real and positive (=1)

Enough phase margin
Real and negative
Imaginary and negative
Unstable (denominator goes to zero and overall gain goes to infinity)

\[ \frac{v_o}{v_i} = \begin{vmatrix} A_0 \\ s^2 + BWs + \omega_0^2 \end{vmatrix} \]

Not Enough phase margin!!!
Key points:

If you want to amplify your signal: \( B(s) \) must be an attenuator (voltage divider!!)

The error is determined by the overall loop gain: \( T(s) = A(s)B(s) \)

\[
\frac{v_o}{v_i} = \frac{A(s)}{1-(-T(s))}
\]

If \( T(s) \gg 1 \), then

\[
\frac{v_o}{v_i} \approx \frac{A(s)}{T(s)} = \frac{1}{B(s)}
\]

Error \( \approx \frac{1}{T(s)} \)

\[
B(s) = \frac{Z_1}{Z_1 + Z_2} = \frac{1}{1 + \frac{Z_2}{Z_1}}
\]

\[
\frac{v_o(s)}{v_i(s)} \approx 1 + \frac{Z_2}{Z_1}
\]
Inverting Amplifier: Apply superposition

A(s) and B(s) are sharing some elements!!

A(s) = ? B(s) = ? T(s) = ?

Why not to open the loop here?

Loop can also be opened here

\[
A(s) = -\frac{Z_2 A_V}{Z_1 + Z_2}
\]

\[
T(s) = \frac{Z_1 A_V}{Z_1 + Z_2}
\]

\[
\frac{v_0}{v_i} = -\frac{A(s)}{1 + T(s)} = -\frac{Z_2 A_V}{Z_1 + Z_2} = \frac{Z_2}{Z_1} \left( 1 + \frac{Z_2}{Z_1 A_V} \right)
\]
Inverting Amplifier: consider the non-zero output impedance!!

\[ \frac{v_o'}{v_i} = \frac{(Z_2 + r_0)A_V}{Z_1 + Z_2 + r_0} = \frac{Z_2}{Z_1} \left[ \frac{1 + \frac{r_0}{Z_2}}{1 + \frac{Z_2 + r_0}{Z_1}} \right] \]

\[ \frac{v_o}{v_i} = \left( \frac{v_o'}{v_i} \right) \frac{Z_1 + Z_2 + r_0}{Z_1 + Z_2 + r_0} = \frac{Z_2}{Z_1} \left[ \frac{1 + \frac{r_0}{Z_2}}{1 + \frac{Z_2 + r_0}{Z_1}} \right] \left[ \frac{1}{1 + \frac{Z_2 + r_0}{A_V}} \right] \]
**Inverting Amplifier:** consider the non-zero output impedance!!

\[
\frac{v_o(s)}{v_i} = \left( \frac{v_o}{v_i} \right) \frac{Z_1 + Z_2}{Z_1 + Z_2 + r_0} = -\frac{Z_2}{Z_1} \left[ \frac{1 + \frac{r_0}{Z_2}}{1 + \frac{Z_2 + r_0}{Z_1 + Z_2}} \right] \left[ \frac{1}{1 + \frac{Z_2 + r_0}{Z_1}} \frac{1}{A_V} \right]
\]

The error can be approximated as:

\[
\text{Error} = \frac{r_0}{Z_2} - \frac{r_0}{Z_1 + Z_2} \frac{1 + \frac{Z_2 + r_0}{Z_1}}{A_V}
\]

IDEAL GAIN

Determined by the OPAMP open-loop gain

Determined by the OPAMP output impedance

What are the effects of the finite OPAMP input impedance and load (externally connected) impedance?

These could be exam questions!!
Benefits of the Feedback

\[
\frac{v_o(s)}{v_i} = \left( \frac{v_o}{v_i} \right) \frac{Z_1 + Z_2}{Z_1 + Z_2 + r_0} = -\frac{Z_2}{Z_1} \left[ 1 + \frac{r_0}{Z_2} \right] \left[ \frac{1}{1 + \frac{Z_2 + r_0}{Z_1 + Z_2}} \right] \frac{1}{1 + \frac{Z_1}{A_V}}
\]

The error can be approximated as:

\[
\text{Error} = \frac{r_0}{Z_2} - \frac{r_0}{Z_1 + Z_2} \frac{1 + \frac{Z_2 + r_0}{Z_1}}{A_V}
\]

IDEAL GAIN

Determined by the OPAMP open-loop gain

Determined by the OPAMP output impedance
OPAMP Characterization

DC gain \((10^4\text{-}10^6) \text{ V/V})\ :

• Very difficult to measure in open-loop due to DC offsets.

\[ V_0 \neq 0 \]

How to measure/characterize it?

• Stabilize for DC

\[ V_i \]

• For DC, the OPAMP operates in closed loop!!
• For frequencies higher than \(1/R_C C_C\), the OPAMP operates in open-loop with a grounded load given by \(R_C\).
How to measure/characterize it?

• At DC

\[
\frac{v_o}{v_i} = \frac{A(s)}{1 + A(s)} \approx 1
\]

• If \( A(s)B(s) \ll 1 \) then the measured gain is dominated by the OPAMP transfer function!

\[
\frac{v_o}{v_i} \approx A(s)
\]
OPAMP Characterization: GBW and stability

\[ V_{in} \rightarrow +A(s) \rightarrow V_0 \rightarrow - \]

\[ A_{VDC} \]

\[ \omega \]

\[ \omega_{p1} \]

GBW

Enough phase margin

OPAMP (open-loop)
OPAMP Characterization: GBW and stability

Not enough phase margin
OPAMP Characterization: Slew-Rate (max speed)

\[ V_{\text{in}} \rightarrow A(s) \rightarrow V_0 \]

- Slew-Rate: \( \max \frac{d}{dt} V_0(t) \)
- Max output level: VDD
- Min output level: VSS

\[ t \quad V_{\text{SS}} \quad V_{\text{DD}} \]
OPAMP Characterization: Output Swing

Use a slow triangular input signal such that the raising and falling edges are not determined by slew rate limitations.

\[
\begin{align*}
A(s) & \triangleq \frac{\Delta V_o}{\Delta V_i} \\
V_{in} & \rightarrow V_{o} \\
V_{min} & \rightarrow V_{max}
\end{align*}
\]
OPAMP Characterization: Input and Output impedance

At Low frequencies:
\[ Z_{\text{measured}} \neq Z_o \] (Why????)

At medium frequencies:
\[ Z_{\text{measured}} = Z_o \parallel R_C \]

Be sure that the OPAMP (all internal transistors) is properly biased during characterization!!

\[ R_C \text{ and } C_C \text{ as large as possible!!} \]
OPAMP Characterization: Input and output impedances

Use a slow triangular input signal such that the raising and falling edges are not determined by slew rate limitations.