ELECTRONIC SYSTEMS

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ES01 - COMPONENTS AND AMPLIFIERS

14/09/2021

temperature ranges
- military $-55° - +125°C$
- industrial $-25° - +85°C$
- consumer $0° - +70°C$

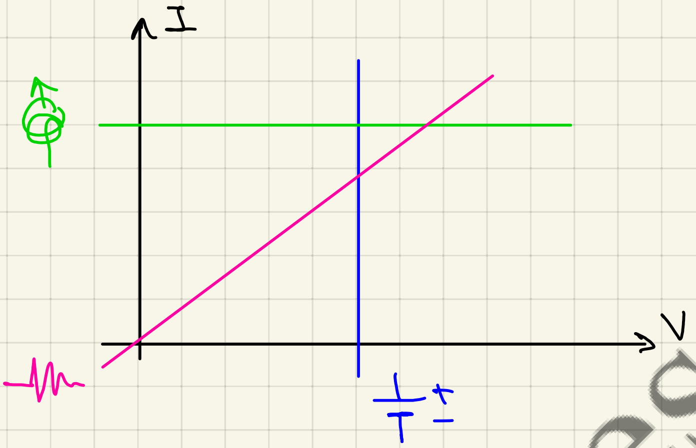
2 different sets of components in PCBs
1) Through-hole
2) SMT - SURFACE MOUNTING TECHNOLOGY

The PCB can be a multi-layer PCB <
- Through-hole via
- Buried via
- Blind via

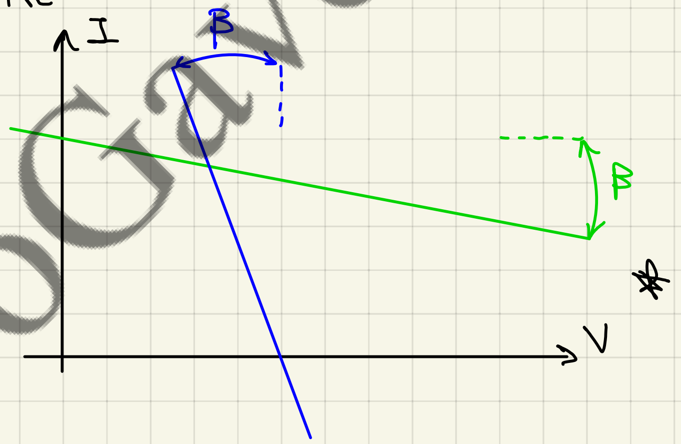
Types of packages:
- DIL - Dual In-Line
- SOT -
- QFP - Quad Flat-Package

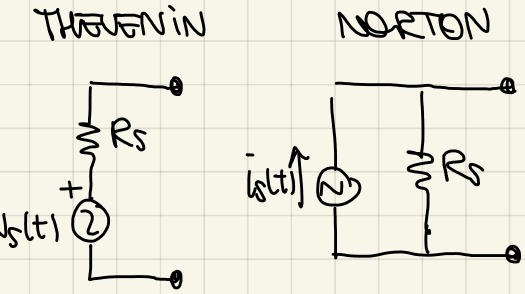
IDEAL & REAL VOLTAGE AND CURRENT SOURCES

- IDEAL

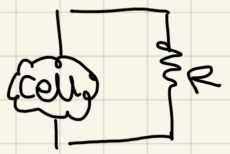


- REAL





* What's the voltage source and what's the current one?
In principle, they can be both: the blue line can be a voltage source w/ a very small Rs or a current source w/ a very large Rs (the same for the green line)
↓
Beware: in reality they don't, let's take for example a biological cell in parallel w/ a resistor (bc then I have an amp. that reads voltage in input)



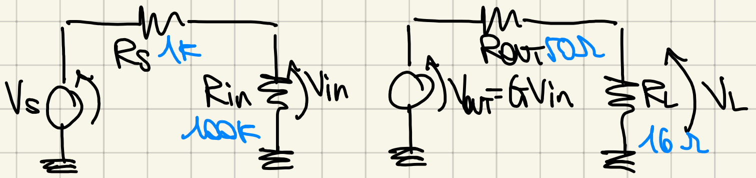
let's suppose the cell feeds the circuit w/ 100nA/excitation and the resistance value is 4.7MΩ
This means that across the resistor, and so across the cell too, a 0.47V voltage arises. The cell can survive to such a voltage and that's ok. But if, for instance, we had a very much higher resistor, the voltage across it would much larger and the cell could not survive to such a voltage bc the cell membrane would destroy.

⇒ If the cell is a current source → let's read the current, not the voltage!

In other words: if you need to read a current, don't create a voltage, let's use a zero impedance input amplifier

AMPLIFIERS

- VOLTAGE TO VOLTAGE



where $\boxed{G = A_V = \frac{V_L}{V_S}}$

ideal voltage amplifier $\boxed{\begin{cases} R_S = 0, \ R_{in} = \infty \\ R_{out} = 0, \ R_L = \text{whatever} \end{cases}}$

Unfortunately there are voltage drops $\boxed{V_{in} = V_S \frac{R_{in}}{R_{in} + R_S}}$

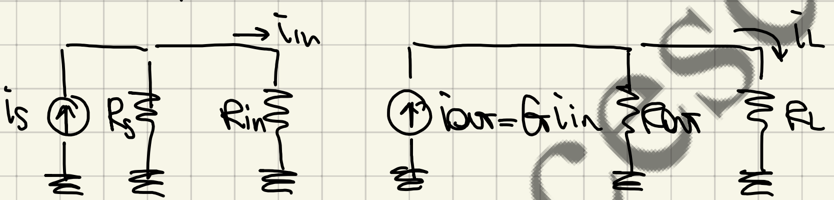
$$V_{out} = G V_{in} = A_V V_{in} = V_L/V_S \, V_{in}$$

$$\boxed{V_L = V_{out} \frac{R_L}{R_L + R_{out}}}$$

Hence, a good voltage amplifier would have:

$\boxed{\begin{cases} R_{in} \gg R_S & \text{very high input impedance} \rightarrow V_{in} \simeq V_S \\ R_{out} \ll R_L & \text{very low output impedance} \rightarrow V_L \simeq V_{out} \end{cases}}$

- CURRENT TO CURRENT



where $\boxed{G = A_i = i_L / i_S}$

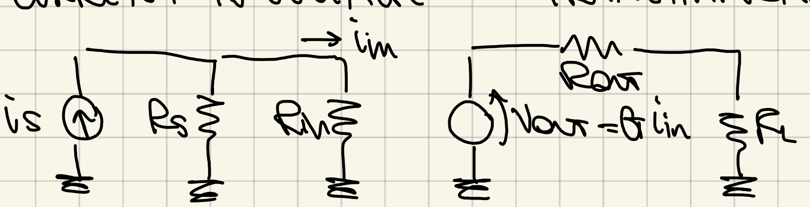
Ideal current generator $\boxed{\begin{cases} R_S = \infty, \ R_{in} = 0 \\ R_{out} = \infty, \ R_L = \text{whatever} \end{cases}}$

$\boxed{i_{in} = i_S \frac{R_S}{R_S + R_{in}}}$ $\quad i_{out} = G \, i_{in} = A_i \, i_{in}$ $\quad \boxed{i_L = i_{out} \frac{R_{out}}{R_{out} + R_L}}$

Hence a good current amplifier should have

$\boxed{\begin{cases} R_{in} \ll R_S & \text{very low input impedance} \rightarrow i_{in} \simeq i_S \\ R_{out} \gg R_L & \text{very high output impedance} \rightarrow i_L \simeq i_{out} \end{cases}}$

- CURRENT TO VOLTAGE → TRANSIMPEDANCE AMPLIFIER



where $\boxed{G = A_R = V_L / I_S}$ $\left[\frac{V}{A} = \Omega\right]$

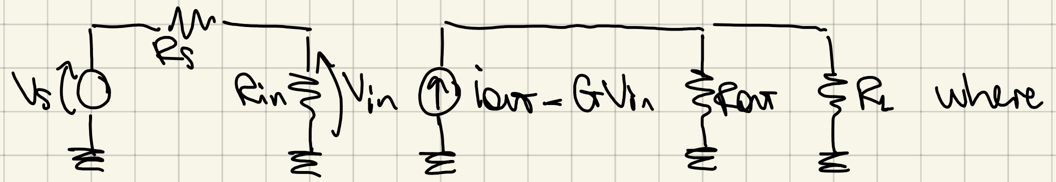
Ideal transimpedance amplifier $\boxed{\begin{cases} R_s = \infty, \; R_{in} = 0 \\ R_{OUT} = 0, \; R_L = \text{whatever} \end{cases}}$

$\boxed{i_{in} = i_s \dfrac{R_s}{R_s + R_{in}}}$ $\quad V_{OUT} = R \, i_{in} = A_R \, i_{in}$ $\quad \boxed{V_L = V_{OUT} \dfrac{R_L}{R_L + R_{OUT}}}$

Hence a good transimpedance amplifier should have:

$\boxed{\begin{cases} R_{in} \ll R_s & \text{very low input impedance} \rightarrow i_{in} \simeq i_s \\ R_{OUT} \ll R_L & \text{very low output impedance} \rightarrow i_L \simeq i_{OUT} \end{cases}}$

- VOLTAGE TO CURRENT → TRANSCONDUCTANCE AMPLIFIER



where $\boxed{G = A_G = \dfrac{i_L}{V_s}}$ $\left[\dfrac{A}{V} = S \right]$

ideal transconductance amplifier $\boxed{\begin{cases} R_{in} = \infty, \; R_s = 0 \\ R_{OUT} = \infty, \; R_L = \text{whatever} \end{cases}}$

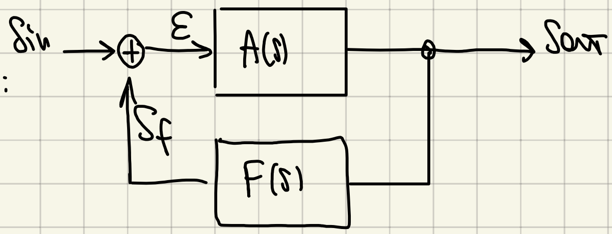
$\boxed{V_{in} = V_s \dfrac{R_{in}}{R_{in} + R_s}}$ $\quad i_{OUT} = G V_{in} = A_G V_{in}$ $\quad \boxed{i_L = i_{OUT} \dfrac{R_{OUT}}{R_{OUT} + R_L}}$

Hence, a good transconductance amplifier should have

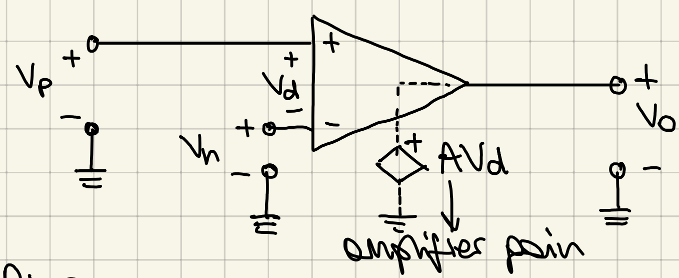
$\boxed{\begin{cases} R_{in} \gg R_s & \text{very high input impedance} \rightarrow V_{in} \simeq V_s \\ R_{OUT} \gg R_L & \text{very high output impedance} \rightarrow i_L \simeq i_{OUT} \end{cases}}$

Instead of designing a lot of different amplifier, we want to design just a single amplifier which is able to have different behaviors depending to what components we're gonna connect to it.
This amplifier is called OPAMP.

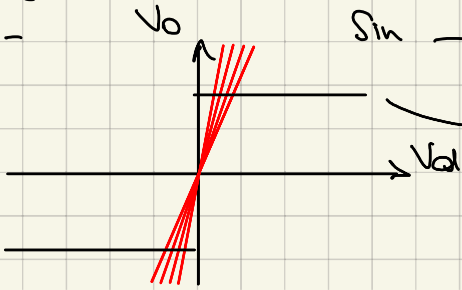
The idea is based on the FEEDBACK:

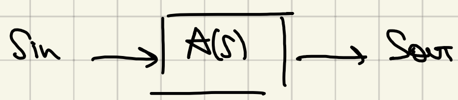


Open loop OPAMP (differential amplifier)



gain $= \dfrac{V_o}{V_d} = A$

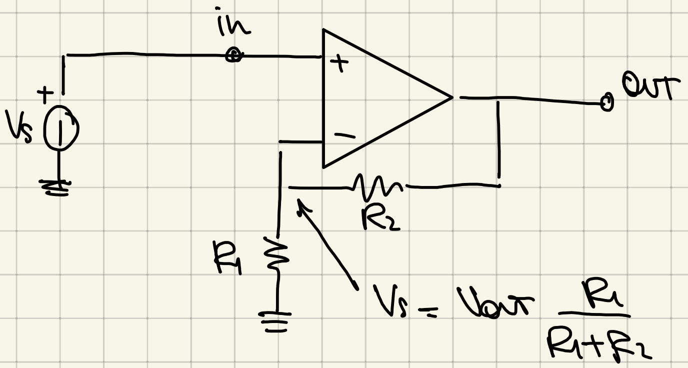




saturation is due to power supply limitation

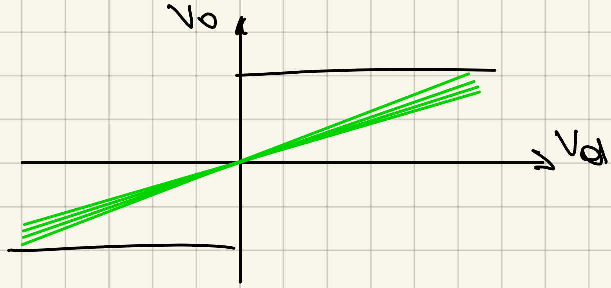
$R_{in} \simeq \infty$
$R_{OUT} \simeq 0$

Basic "closed-loop" (negative feedback) OPAMP:



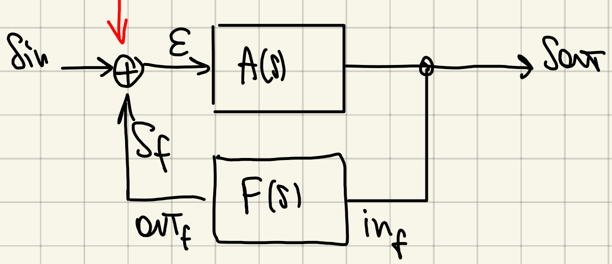
$$V_s = V_{OUT} \frac{R_1}{R_1+R_2}$$

$$Gain = \frac{V_{OUT}}{V_s} = \frac{R_1+R_2}{R_1} = 1 + \frac{R_2}{R_1}$$



We have a lower gain, but anyway it worths it bc:
1) thanks to the FB we can control the gain (just by changing R_1 and R_2)
2) now we can design just one amplifier and thanks to the FB we can "simulate" whatever amplifier
3) thanks to FB we have a very low tollerance

Beware: if here we consider a sum, so we have to consider $A(s) \cdot F(s) < 0$



$$\begin{cases} S_{OUT} = A(s) \cdot \varepsilon \\ \varepsilon = S_{in} + S_f \\ S_f = F(s) \cdot S_{OUT} \end{cases}$$

REAL (CLOSED-LOOP) GAIN:

$$G = \frac{S_{OUT}}{S_{in}} = ? \qquad S_{OUT} = A(s)\cdot\varepsilon = A(s)[S_{in} + S_f] = A(s)S_{in} + A(s)[F(s)S_{OUT}]$$

$$S_{OUT}[1 - A(s)F(s)] = A(s)S_{in}$$

$$\boxed{G(s) = \frac{S_{OUT}}{S_{in}} = \frac{A(s)}{1 - A(s)F(s)} = \frac{A(s)/G_{LOOP}}{\frac{1 - G_{LOOP}}{G_{LOOP}}} = \frac{A(s)}{\underbrace{G_{LOOP}}_{1/F(s)}} \cdot \frac{1}{\frac{1}{G_{LOOP}} - 1} = -\frac{1}{F(s)} \frac{1}{1 - 1/G_{LOOP}}}$$

$G_{LOOP} = A(s)F(s) < 0$

$$\boxed{G_{ideal} = -\frac{1}{F(s)}} \qquad \boxed{\frac{1}{1 - 1/G_{LOOP}} = \text{CORRECTION FACTOR}}$$

$$\Rightarrow \boxed{G_{real} = G_{ideal} \frac{1}{1 - 1/G_{LOOP}}} \quad \text{REAL GAIN}$$

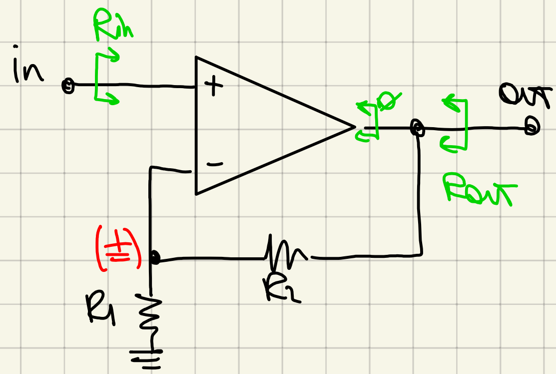
Notice: if $G_{LOOP} \to \infty \Rightarrow CF \to 1 \Rightarrow G(s) = -\frac{1}{F(s)} = G_{ideal}$

G_{LOOP} measures the feedback effectiveness, the higher the better!

if $G_{LOOP} \to \infty \Rightarrow \varepsilon \to 0 \equiv$ the negative FB forces S_f to mimic S_{in}

Notice: G_{LOOP} doesn't depend on the OPAMP

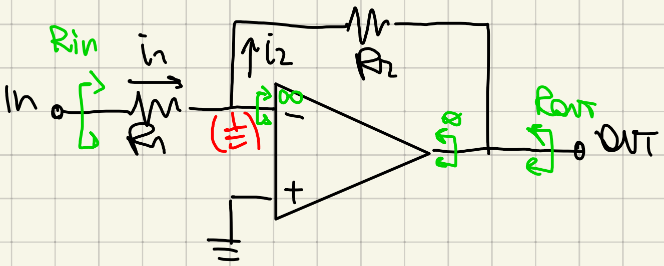
- IDEAL NON INVERTING CONFIGURATION



$$\boxed{G = A_{vr} = \frac{V_{OUT}}{V_{in}} = 1 + \frac{R_2}{R_1}}$$

$$\boxed{R_{in} \approx \infty \qquad R_{OUT} \approx 0}$$

- IDEAL INVERTING CONFIGURATION

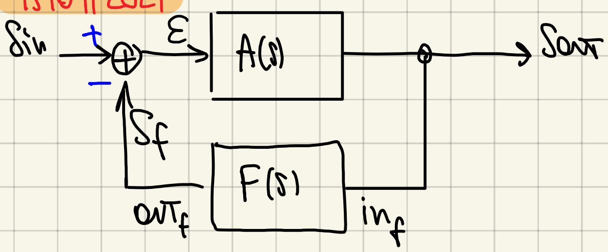


$V_{OUT} = -i_2 R_2$
$i_2 = \frac{V_{in}}{R_1}$ $\rightarrow$ $\boxed{G = A_{vr} = \frac{V_{OUT}}{V_{in}} = -\frac{R_2}{R_1}}$

$$\boxed{R_{in} = R_1 \qquad R_{OUT} \approx 0}$$

ES02 - NEGATIVE FEEDBACK

15/09/2021



- Gloop lowers the gain but adds many advantages
- gain is independent of the OPAMP and depends just on FB
- stronger FB, smaller CF

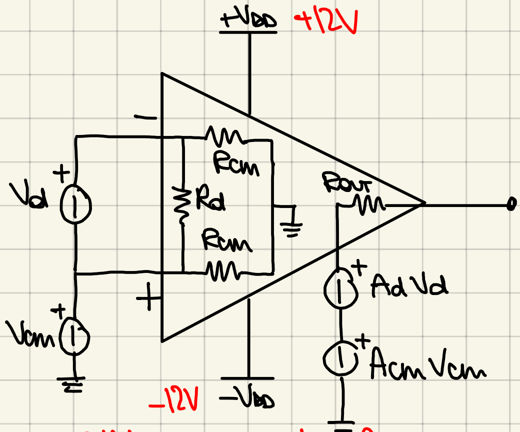
$$G(s) = \frac{S_{OUT}}{S_{in}} = \frac{A(s)}{1 \underset{(+)}{-} A(s)F(s)} = \frac{A(s)/G_{loop}}{\frac{1 \underset{(+)}{-} G_{loop}}{G_{loop}}} = \frac{A(s)}{\underbrace{G_{loop}}_{1/F(s)}} \cdot \frac{1}{\frac{1}{G_{loop}} \underset{(+)}{-} 1} \overset{(+)}{=} \frac{1}{F(s)} \frac{1}{1 \underset{(+)}{-} 1/G_{loop}}$$

$G_{loop} = A(s)F(s) < 0$

if Gloop is very high $\rightarrow \frac{1}{G_{loop}} \simeq 0 \rightarrow 1 - \frac{1}{G_{loop}} \simeq 1 \rightarrow CF \simeq 1 \rightarrow G(s) \simeq -\frac{1}{F(s)} = G_{ideal}$

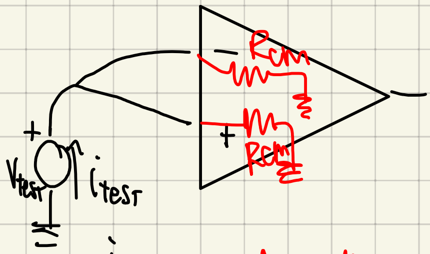
ADVANTAGES IN USING FB:
1. larger BW (high speed)
2. lower Rout (excellent voltage source)
3. { Rin = ∞ (excellent voltage reader)
   Rin = 0 (excellent current reader)

REAL OPAMP

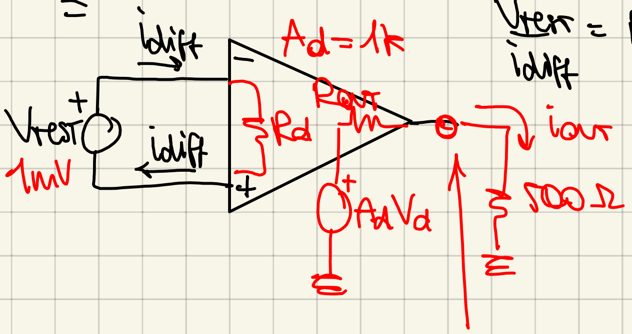


Ad = differential gain > 100dB
Acm = common mode gain < 20dB
BW = bandwidth 10Hz ÷ 1kHz
Rd =
Rcm =
Ro =
Temperature drifts some %/°C

Check if our amplifier is ideal or not



$\frac{V_{test}}{I_{test}} = R_{cm} \to \infty$

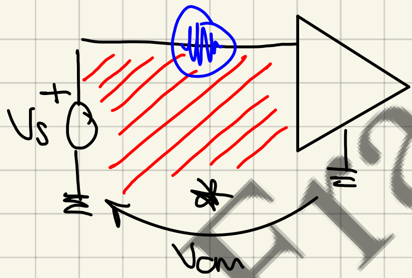


$\frac{V_{test}}{i_{diff}} = R_{diff}$

$V_{out} = 1k \cdot 1mV$

Why do we prefer to use a differential input amplifier instead a single input one?

SINGLE-ENDED INPUT



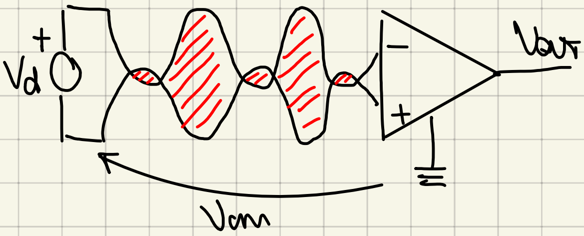
if the ground of my source is different from the ground of the amplifier a common mode difference btw the two grounds arises
↓
At the input of the amplifier we'll have Vd + Vcm so at the output we'll not have just Vd properly amplified but we'll have some disturbances

Practical view: it seems that someone is speaking through the mic, but it's just a fluctuation of the cm signal

*Furthermore any electromagnetic change w/in this area will generate a voltage drop that can be modelled w/ a noisy generator applied in series to the source

So the output of the stage will experience different problems:

DIFFERENTIAL INPUT



Using a diff. input we can twist the two wires:

- The overall area is much more reduced
- an electromagnetic field entering in this area will cause two opposite effects that cancel out each other.
  ↓
  immunity to disturbances
- we can exploit the negative feedback

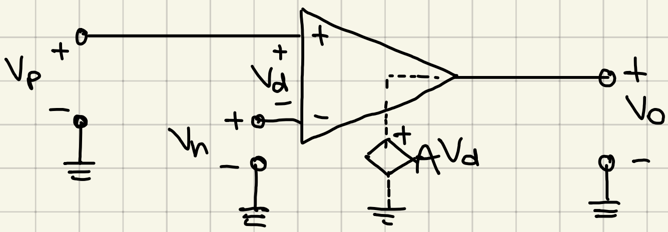
$V_{OUT} = A_d V_d + A_{cm} V_{cm}$

In order to reject the CM:

$CMRR = \frac{A_d}{A_{cm}}$ should be very high ($90 \div 100\,dB$)

EFFECT OF FB ON AMPLIFIER'S MISMATCHES AND DRIFTS

- OPEN LOOP OPAMP

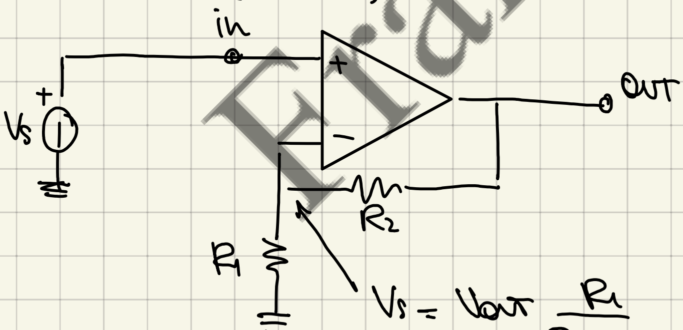


we have a nominal value but also some spreads up to 50%

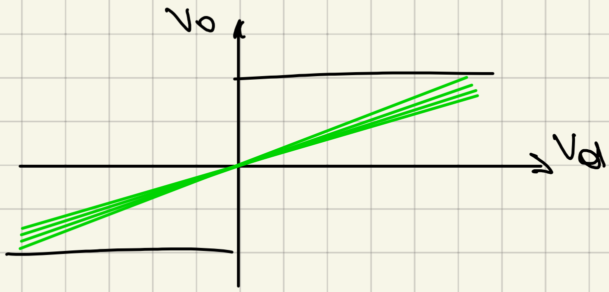
$G = \frac{V_o}{V_d} = A \rightarrow \frac{dG}{G} = \frac{dA}{A}$ → if the amplifier has a gain that varies of 50% for example, also the overall gain varies of 50%.

A ↓ gain of the amplifier

CLOSED LOOP (NEG. FB) OPAMP



$V_s = V_{out} \frac{R_1}{R_1+R_2}$



$$G = \frac{V_o}{V_d} = \frac{A}{1-AF} = \frac{A}{1-G_{loop}}$$

$$\frac{dG}{dA} = \frac{(1-AF)+AF}{(1-AF)^2} = \frac{1}{(1-AF)^2}\,\frac{A}{A} = \frac{G}{A}\,\frac{1}{1-G_{loop}}$$

$$\rightarrow \frac{dG}{G} = \frac{dA}{A} \cdot \frac{1}{1-G_{loop}}$$

if A varies of ±50% and if Gloop = 100 for example, so the overall gain varies of:

$$\frac{dG}{G} = \pm 50\% \cdot \frac{1}{1+100} \simeq \pm 5‰ \rightarrow \text{very very much reduced}$$

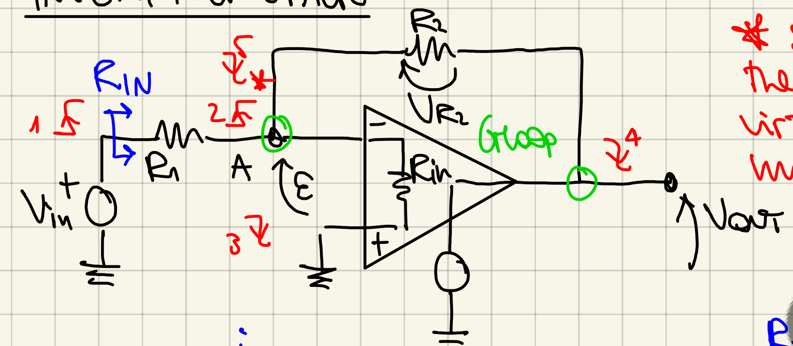
What happens if the feedback network has a fluctuation?

$$\frac{dG}{dF} = \dots \longrightarrow \frac{dG}{G} = \frac{df}{F}$$

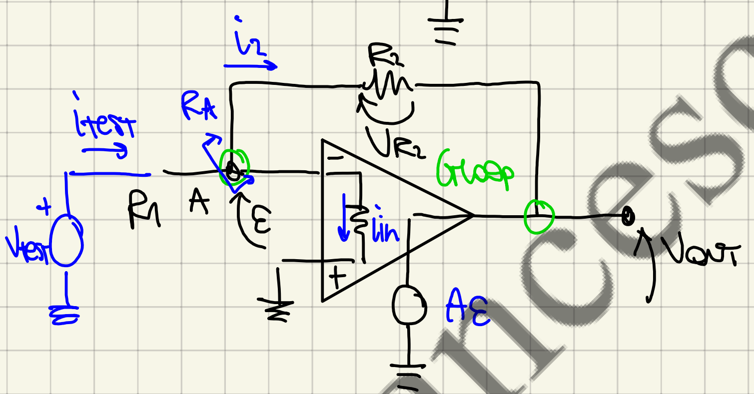
This means that we can buy an amplifier that can also be not so much precise, but the components that of the feedback network must be very precise!!

FEEDBACK EFFECT ON IMPEDANCES

- INVERTING STAGE



* since we are probing a feedback node, the feedback tries to push the node @ virtual ground, so the impedance we'll find must be divided for 1-Gloop
↓
The loop tends to decrease it!



$$R_A = \frac{\varepsilon}{i_{TEST}} = \frac{\varepsilon}{i_{in} + i_2} = \frac{\cancel{\varepsilon}}{\frac{\cancel{\varepsilon}}{R_{in}} + \frac{\cancel{\varepsilon} - A(s)\cdot\cancel{\varepsilon}}{R_2}} = \frac{1}{\frac{R_2 + R_{in}(1-A(s))}{R_{in}R_2}}$$

$$= \frac{R_{in}R_2}{R_2 + R_{in} - A(s)R_{in}} = *$$

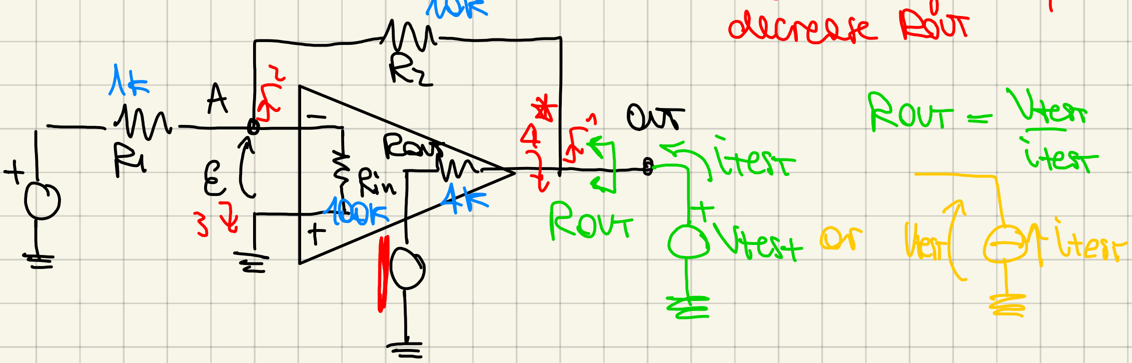
$$* = \frac{R_{in}R_2}{(R_{in}+R_2)} \frac{1}{1 - A(s)\frac{R_{in}}{(R_{in}+R_2)}} = \frac{R_{in}\|R_2}{1 - A(s)\frac{R_{in}}{R_{in}+R_2}} = \frac{R_{A\,STUPID}^{*}}{1 - G_{loop}}$$

* $R_{A\,STUPID}$ is the impedance I'd obtain turning off the output generator, so assuming (wrongly) that the output stays at ground
↓
That's wrong bcs the output doesn't remain @ ground, bcs ε changes and so the output changes

INPUT IMPEDANCE: $R_{IN} = R_1 + R_A = R_1 + \dfrac{R_{A\,STUPID}}{1 - G_{loop}}$

Notice: if Gloop is very high $\rightarrow R_A \simeq 0 \rightarrow R_{IN} \simeq R_1$

What about the output impedance?



Using what we have seen before → $R_{OUT} = \dfrac{R_{STUPID}}{1-G_{LOOP}}$

$R_{STUPID} = ? = R_{out} \parallel [R_2 + R_1 \parallel R_{in}] \simeq 3k$ → bad, we want 0Ω, but don't worry because it must be divided for $1-G_{LOOP}$

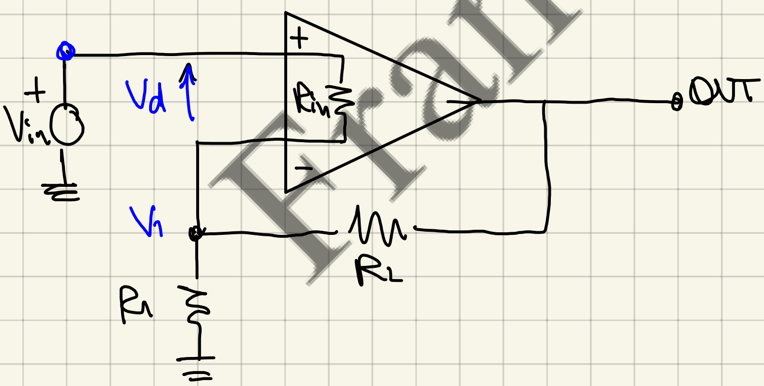
$$R_{OUT} = \frac{R_{out} \parallel [R_2 + R_1 \parallel R_{in}]}{1-G_{LOOP}}$$

if G_{LOOP} is very high → $R_{OUT} \simeq 0$

For what concerns the transfer function

$$\frac{V_{OUT}}{V_{in}} = \underbrace{\left(\frac{V_{OUT}}{V_{in}}\right)_{STUPID}}_{\text{inverting stage}} \cdot \underbrace{\frac{1}{1-1/G_{LOOP}}}_{CF} = \left(-\frac{R_2}{R_1}\right)\frac{1}{1-1/G_{LOOP}}$$

• NON-INVERTING STAGE



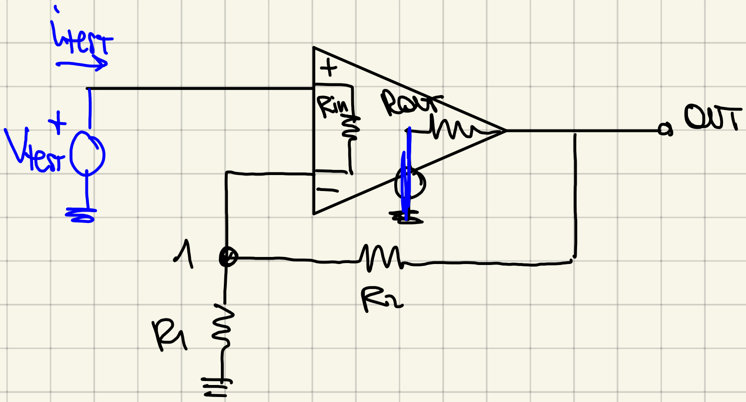
REAL TRANSFER FUNCTION

$$G_{real} = \frac{V_{OUT}}{V_{IN}} = G_{ideal} \cdot CF = \frac{1}{\beta}\,\frac{1}{1-1/G_{LOOP}}$$

$$\beta = \frac{V_f}{V_{OUT}} = \frac{R_1}{R_1+R_2} \Rightarrow \frac{1}{\beta} = \frac{R_1+R_2}{R_1} = 1 + \frac{R_2}{R_1}$$

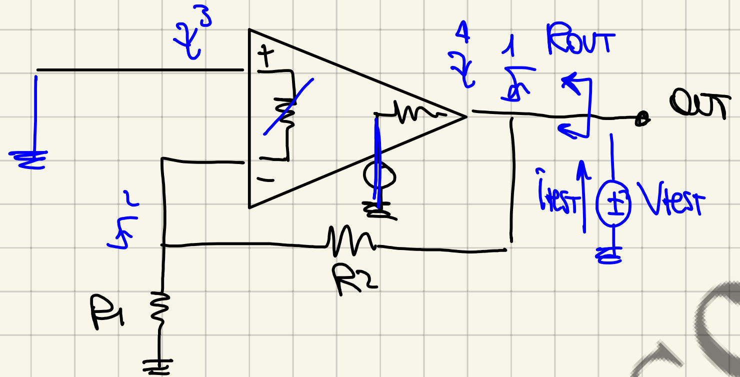
$$G_{real} = \left(1+\frac{R_2}{R_1}\right)\frac{1}{1-1/G_{LOOP}}$$

INPUT IMPEDANCE:



$$R_{IN} = R_{IN}^{*}\left[1 - G_{LOOP}\right] = \left\{R_{in} + \left[R_1 \,||\, (R_2 + R_{OUT})\right]\right\}\left[1 - G_{LOOP}\right] \qquad R_{IN} \to \infty$$

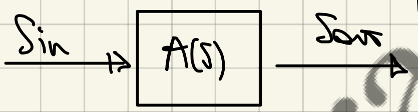
OUTPUT IMPEDANCE

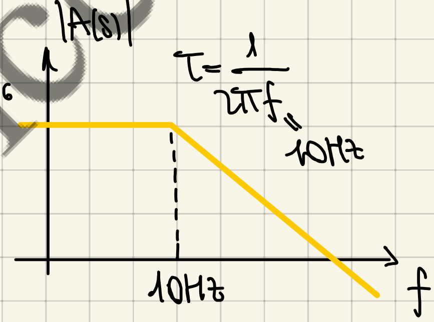


$$R_{OUT} = \frac{R_{OUT}^{*}}{1 - G_{LOOP}} = \frac{R_{out} \,||\, \left[R_2 + R_1 || R_{in}\right]}{1 - G_{LOOP}} \qquad R_{OUT} \approx 0$$

FB EFFECT ON BW

OPEN LOOP (just the OpAmp)



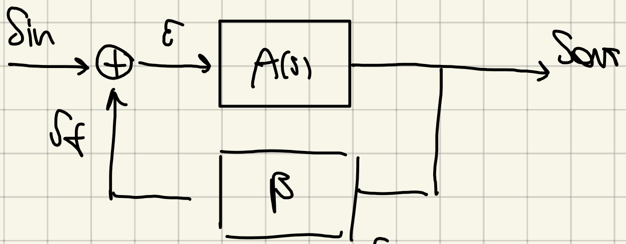


$$\tau = \frac{1}{2\pi f} \qquad f = 10\text{Hz}$$

$$\boxed{A = \frac{G_0}{1 + s\tau}}$$

$$s = -\frac{1}{\tau} = j2\pi f$$

CLOSED LOOP



$$G = \frac{A}{1 + A\beta} = \frac{\frac{G_0}{1 + s\tau}}{1 + \beta\frac{G_0}{1 + s\tau}} = \dots = \frac{G_0}{(1 - G_{LOOP})\left(1 + \frac{s\tau}{1 - G_{LOOP}}\right)}$$

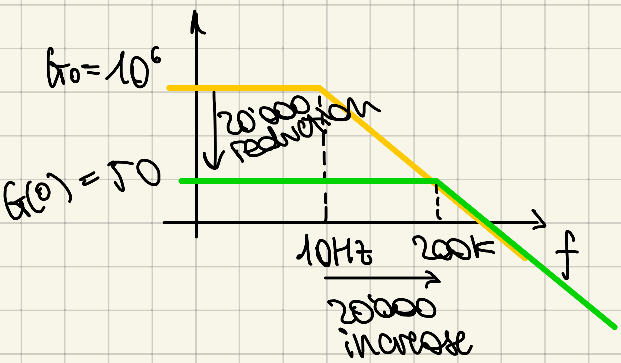
As we have already seen the DC gain decreases, since now it is:

$$G(0) = \frac{G_0}{1-G_{loop}} \left(= G_{ideal} = \frac{1}{F}\right)$$

but now we have a new time constant, and so a new pole:

$$s = j2\pi f\,(1-G_{loop}) \rightarrow f_p = \frac{1}{\tau}(1-G_{loop})$$

For example if: $F = 1/50$ $\quad G_{loop} = -20000$ $\quad G_{real} \cong 50$ $\quad$ pole $= 200kHz$

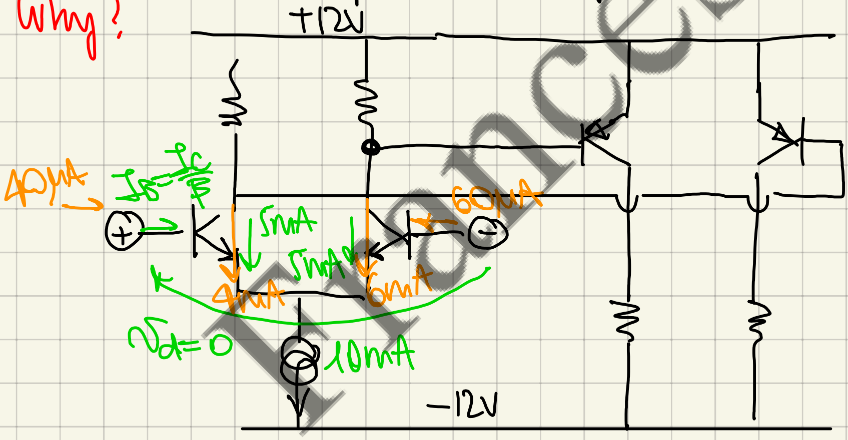


INPUT ERRORS IN REAL OPAMP

In a real OpAmp there are also bias currents and offset voltages

<u>Beware</u>: it's not the same current we modeled w/ R_{cm} inside the OpAmp
That current is zero if $V_{test} = 0$

The bias current is present even if we do not apply any differential voltage in input
Why?

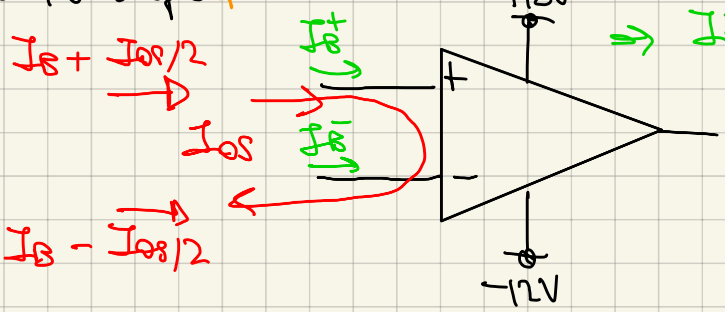


Even if $V_d = 0$ a current flows through the channel, which means that the BJT requires a base current:
$I_B = I_C/\beta$
↓
This means that in input of our real OpAmp we have a bias current due to the bias of the input transistors

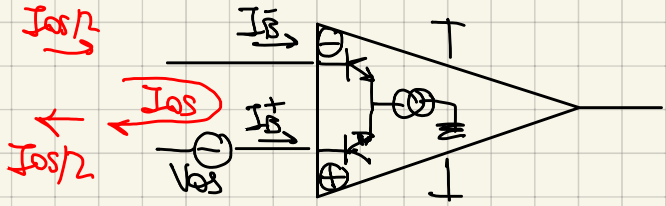
But when we use an OpAmp we apply a differential signal, bcz we want to amplify it.
↳ Doing this we unbalance the stage*

if $\beta = 100 \rightarrow I_B = \frac{5mA}{100} = 50\mu A$

$\Rightarrow I_B^+ = I_B^- = 50\mu A$



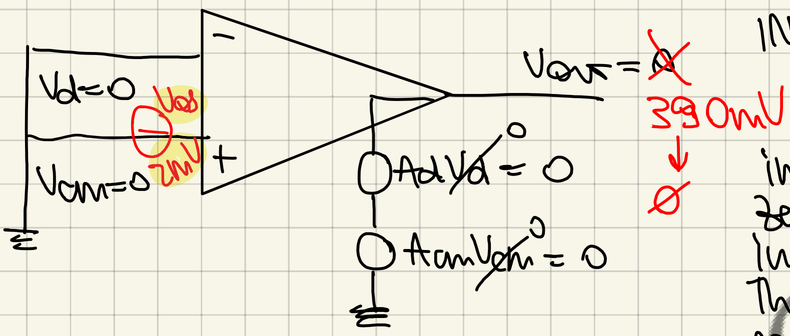
NON-IDEAL LARGE-SIGNAL SPECS :



INPUT BIAS CURRENTS : $I_B = (I^- + I^+)/2$
↓
average current that enters into the + and - input ($< 1\mu A$)

INPUT OFFSET CURRENT : $I_{os} = I^- - I^+$
↓
is the mismatch between the two inputs (is the unbalance) ($> 10nA$)

We have also another issue. In principle, if we do not apply any differential signal in input, the output should be zero:



INSTEAD THE OUTPUT IS NOT NIL
↓
in order to bring the output back to zero we have to unbalance the input stage: we have to apply a Vos that generates an unbalance that compensates the mismatches inside the stage

Beware: all the parameters change w/ temperature (few %/°C): TEMPERATURE DRIFT

$\underbrace{\text{few \%/°C}}$ TC = TEMPERATURE COEFFICIENT

example: if $I_B = 1\mu A$ $\Delta T = 10°C \rightarrow 50°C$ (40°C) and $TC = 20\%/°C$

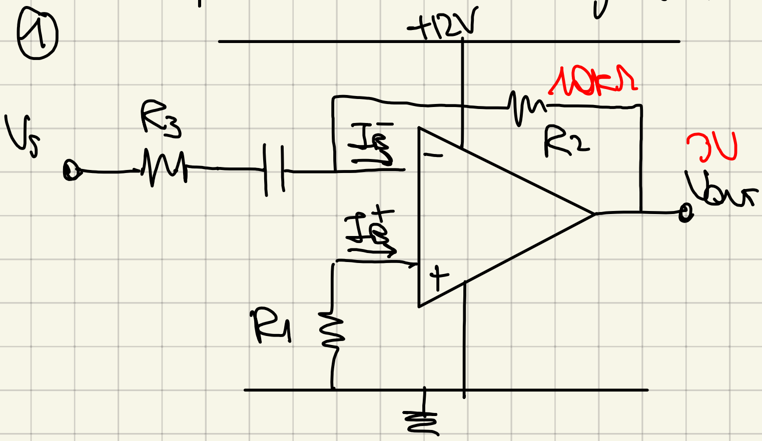
$$\Rightarrow \Delta I_B = I_B \cdot \Delta T \cdot TC = 1\mu A \; 40°C \; \frac{20\%}{°C} = 800\% \cdot 1\mu A$$

$$\Rightarrow I_B = 1\mu A \rightarrow 9\mu A$$

INPUT BIAS CURRENTS In reality bias currents are not a problem, because if we don't apply them the OpAmp cannot operate

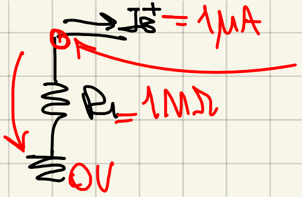
Even if from a schematical stand point these two circuits look correct, due to the input bias currents they cannot work:

1



@DC → source is switched off (C is open)
we need to provide $I_B^\pm$ to the opAmp

if $V_{OUT} = 3V$ for instance, the ⊖ input can drink I_B^- from its own output let's look at the ⊕ input



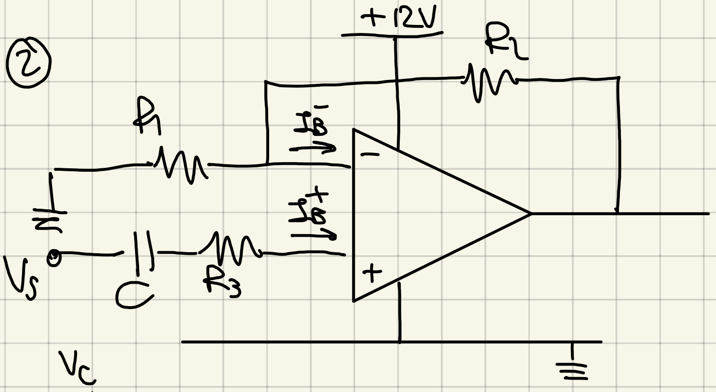
$\Delta V = I_B^+ R_1 = 1\mu A \cdot 1M\Omega = 1V$

→ but it cannot be negative since the power supply we are using is $0V \div +12V$

⇒ the ⊕ input cannot sink a current

⇒ this circuit cannot work

②

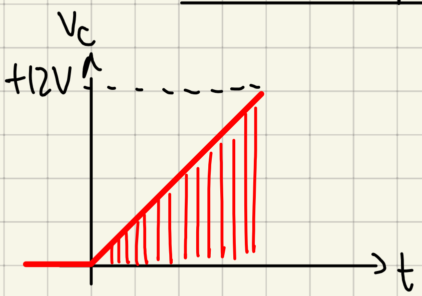


$$i_c(t) = C \frac{dV_c(t)}{dt}$$

if we force a constant current through a capacitor

$$1\mu A = C \frac{dV_c(t)}{dt}$$

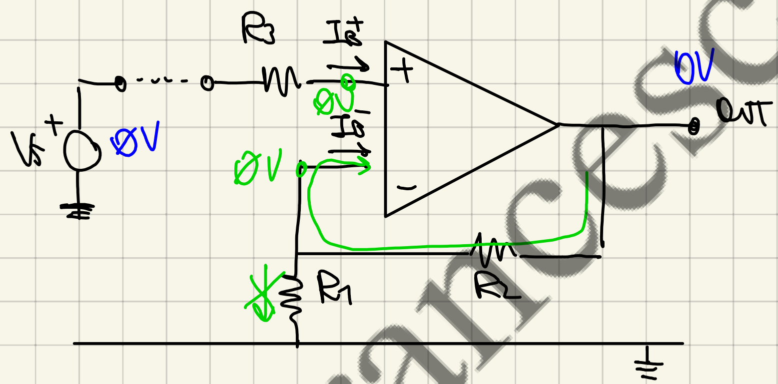
→ $dV_c(t)/dt$ is constant



When V_c reaches the power supply (+12V) then V^- reaches +12V and so the OpAmp is no longer able to operate bcz the + pin saturates due to the fact that the capacitor placed at the input stops any flow of DC current

$I_B^{\pm}$ is bad, but it's needed!

Why is it bad? → bcz it causes an error @ the output of the stage



If the input is 0V the output should be 0V, but let's consider the bias currents:

- I_B^+ : (neglect I_B^-)

$V^+ = -I_B^+ R_3$

$V_{OUT}(I_B^+) = -I_B^+ R_3 A = -I_B^+ R_3 \left(1 + \frac{R_2}{R_1}\right)$

- I_B^- (neglect I_B^+) : $V^+ = ØV \rightarrow V^- = ØV \Rightarrow V_{OUT}(I_B^-) = I_B^- R_2$

$\Rightarrow \tilde{V}_{OUT} = V_{OUT}(I_B^+) + V_{OUT}(I_B^-) = -I_B^+ R_3 \left(1 + \frac{R_2}{R_1}\right) + I_B^- R_2$

we could end up for example w/ : $= -20mV + 7mV = -13mV$

So even if the input is Ø, the output is not null

We would like to have: $\tilde{V}_{OUT} = 0 \rightarrow -I_B^+ R_3 \left(1 + \frac{R_2}{R_1}\right) = I_B^- R_2$

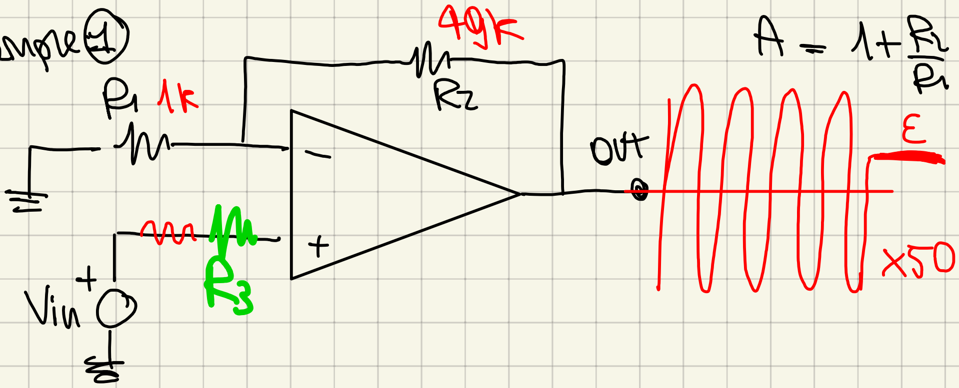
since $I_B^+ = I_B^- = 1\mu A$ (for example) $\rightarrow \left|-R_3\left(1 + \frac{R_2}{R_1}\right)\right| = |R_2|$

$\tilde{V}_{OUT} = 0$ when $R_3 = \frac{R_2}{1 + R_2/R_1} = \frac{R_1 R_2}{R_1 + R_2} = R_1 \| R_2$

if we choose $\boxed{R_3 = R_1 \| R_2}$ → The effects of the two bias currents compensate and $V_{OUT} = 0\,V$ (w/ $V_S = 0\,V$)

example ①

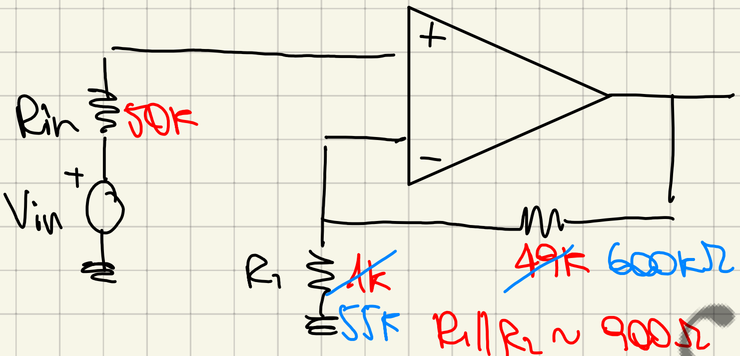
$$A = 1 + \frac{R_2}{R_1} = 1 + \frac{49k}{1k} = 50$$



In order to not have $V_{OUT} \neq 0$ when $V_{in} = 0$, we should add $R_3 = R_1 \| R_2 \simeq 900\Omega$

example ② : it may happen that our mic has already an input impedance of 50kΩ

What can we do?

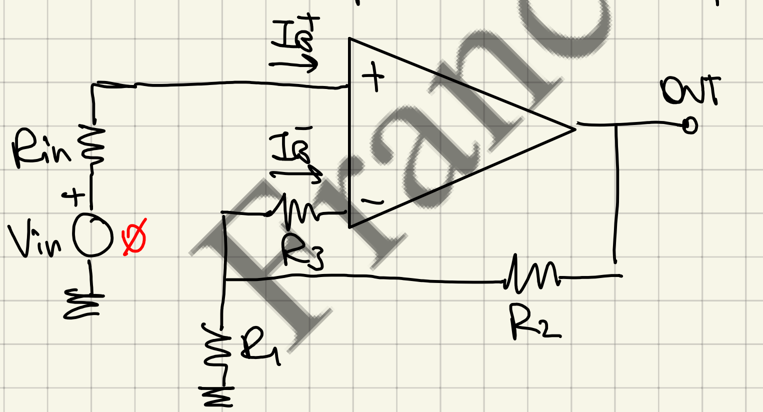


$R_1 \| R_2 \sim 900\Omega$

$R_1 \| R_2 \simeq 50k\Omega$

1. we can change mic w/ one that has 900Ω input impedance
2. or w/ one that has 100Ω input impedance and then we add a 800Ω impedance in series
3. if we cannot change mic we have to change our parameters*

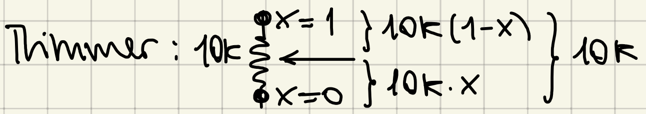
4. we add an impedance to the ⊖ pin :

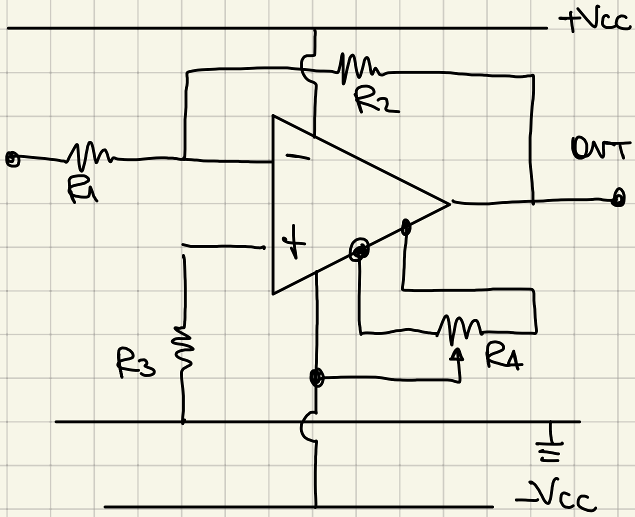


$$\boxed{R_{in} = R_3 + R_1 \| R_2}$$

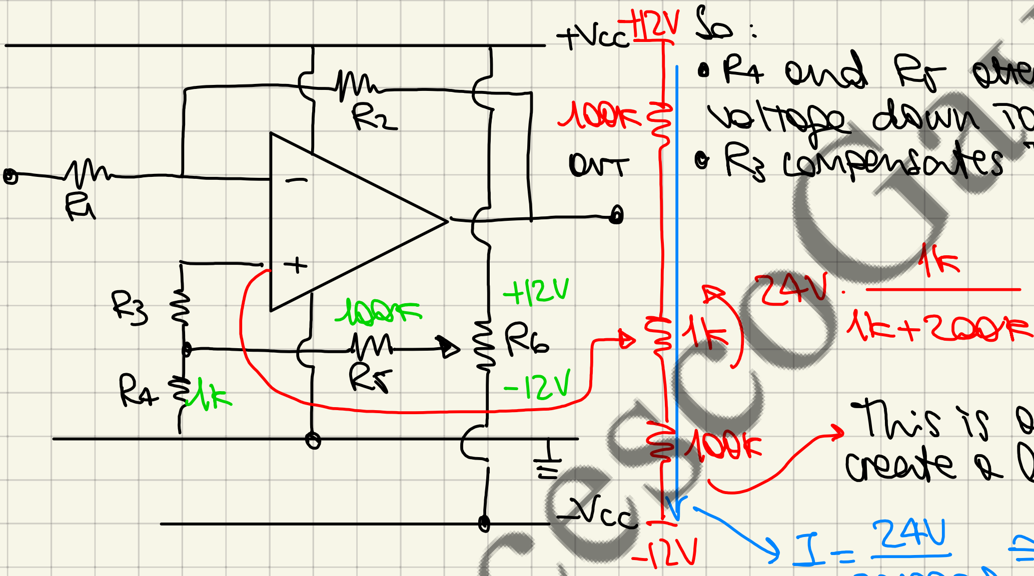
INPUT OFFSET VOLTAGE

There are some OpAmps that have some additional pins to which we can connect an external trimmer in order to properly trim that variable resistor and compensate the mismatches inside the opamp and bring the output back to zero (when no input signal is applied).





In case the OpAmp has no additional pins for the trimmer we can adjust the offset by simply using an external trimmer which is connected to ±Vcc. But, since such a swing is too high, we cannot connect the ⊕ straight to the trimmer, but we should attenuate it through R4 and R5:



So:
- R4 and R5 attenuates the trimming voltage down to some few mV
- R3 compensates the bias current

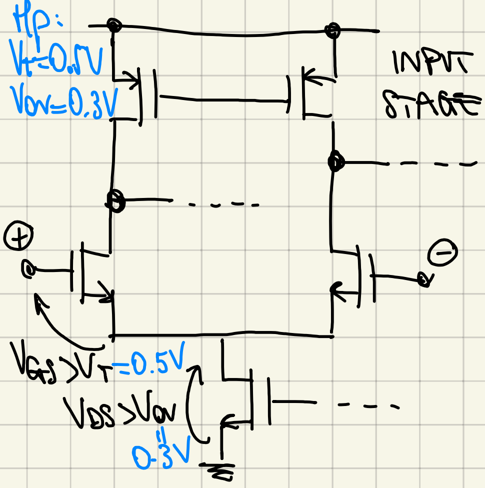
This is another way to properly create a low voltage

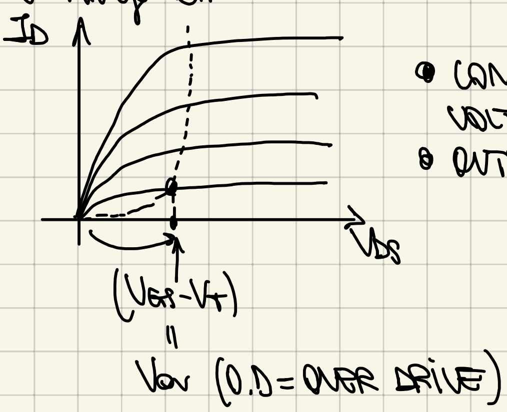
$I = \frac{24V}{201000\,\Omega} \cong 0.1\,\mu A$ GOOD!! (very low current) → low battery consumption

FINAL GOAL:
- reduce the full power supply variation into a small variation
- maximize the resistances seen at the input in order to cancel out the offset

DYNAMIC RANGES (there's a range for the input and one for the output)
We told that an OpAmp is something like:

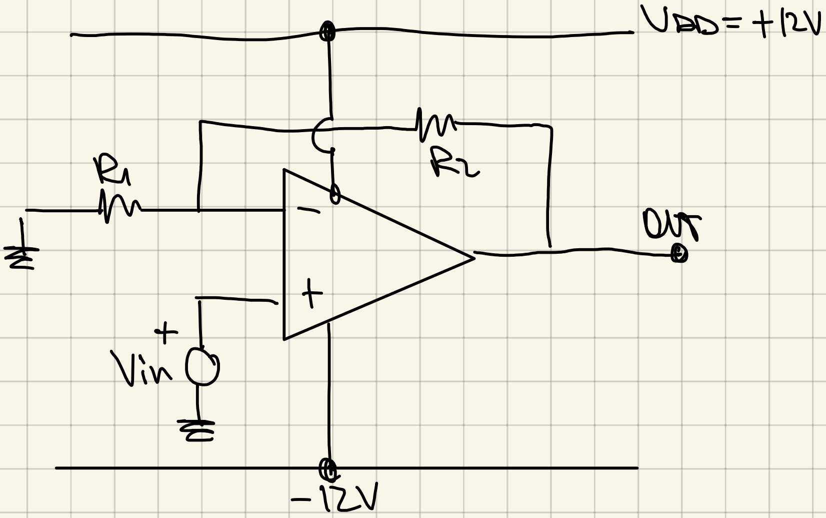
- COMMON MODE INPUT VOLTAGE RANGE
- OUTPUT VOLTAGE SWING





ES02 - NEGATIVE FEEDBACK (2)

21/09/2021



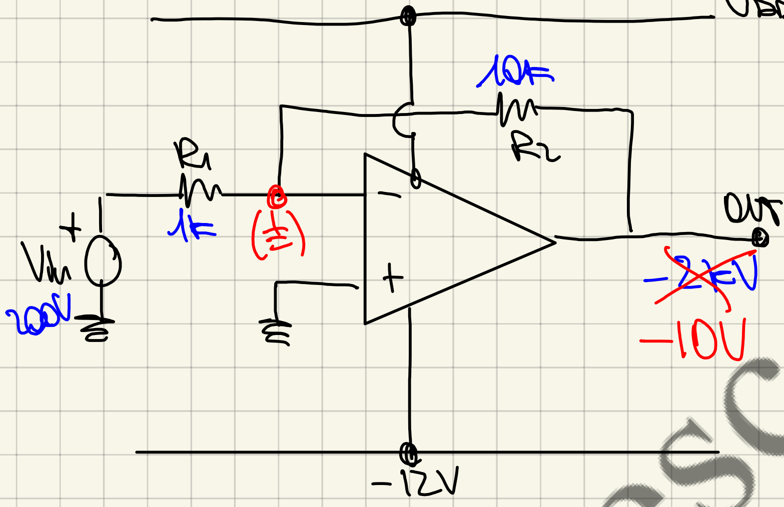
• rail-to-rail OpAmp

↓

CMVR = ±12V

$$\begin{cases} V_{in,max} \le +12V \\ V_{in,min} \ge -12V \end{cases}$$

Let's see what happens in an inverting configuration:



The ⊖ pin is @ v.g. (0V) whatever Vin is applied

↓

0V is w/in the CMVR, so there is no limitation on Vin

↓

Vin can be whatever!!!

$G(0) = -\frac{R_2}{R_1} = -10$

If we apply a 200V input in principle the output would reach −2kV, but it cannot due to the OVS (output voltage swing), so it will saturate to the most negative power supply (−12V in case of rail-to-rail OpAmp, or −10V in case of not rail-to-rail OpAmp, w/ a ±2V headroom)

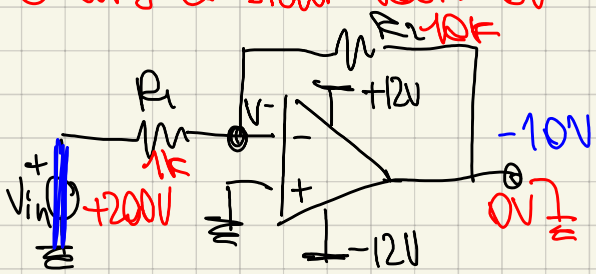
↓

The output will saturate to −10V (not rail-to-rail OpAmp)

↓

V^- will not be v.g. (0V), but will be given by the superposition of two effects

① Let's consider $V_{out} = 0V$ ② Let's consider $V_{out} = -10V$



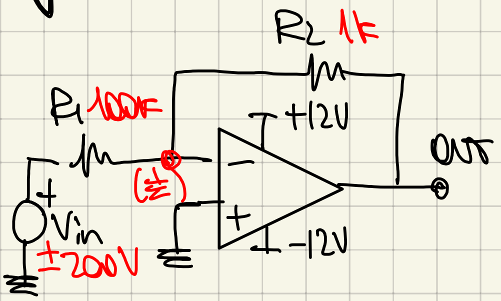
$$V^- = 200V \frac{R_2}{R_1+R_2} + (-10V)\frac{R_1}{R_1+R_2} =$$

$$\cong 180V - 0.9V = +179.1V$$

⇒ In this case the OpAmp would burn not because it exceeds the CMVR (it could also be correct), but bcz the gain is so high that the output would reach a →

→ too high value

How can we solve this issue?

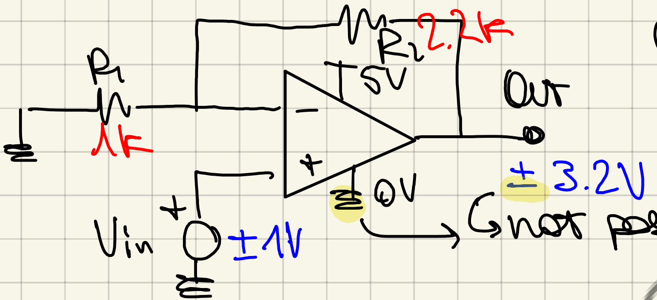


$$V_{out} = \pm 200V \left(-\frac{R_2}{R_1}\right) = \mp 2V$$ with $\frac{R_2}{R_1} \approx 0.01$ → The output is w/in the OVS

So even if the PS is ± 12V (and our CMVR is ± 10V) we can apply ±200V in input but we have to size the circuit in a correct way.

Beware: if we have a single-polarity OpAmp we can amplify only a single polarity input signal

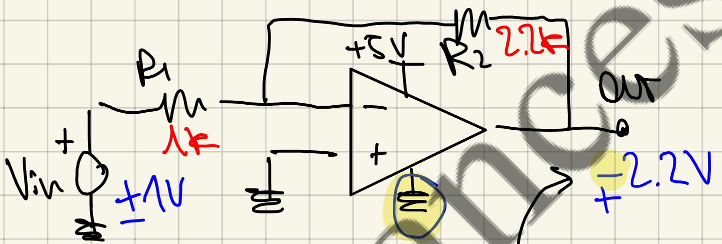
ex①: non-inverting single polarity OpAmp:



$$G = 1 + \frac{R_2}{R_1} = 1 + \frac{2.2}{1} = 3.2$$

not possible! ⇒ $\boxed{V_{in} \geq 0}$

ex②: inverting single-polarity OpAmp:

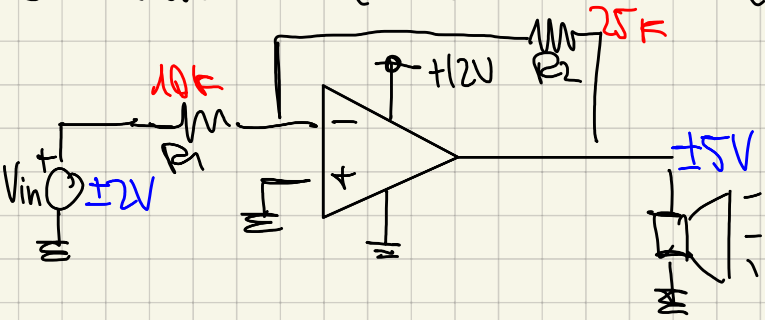


$$G = -\frac{R_2}{R_1} = -2.2$$

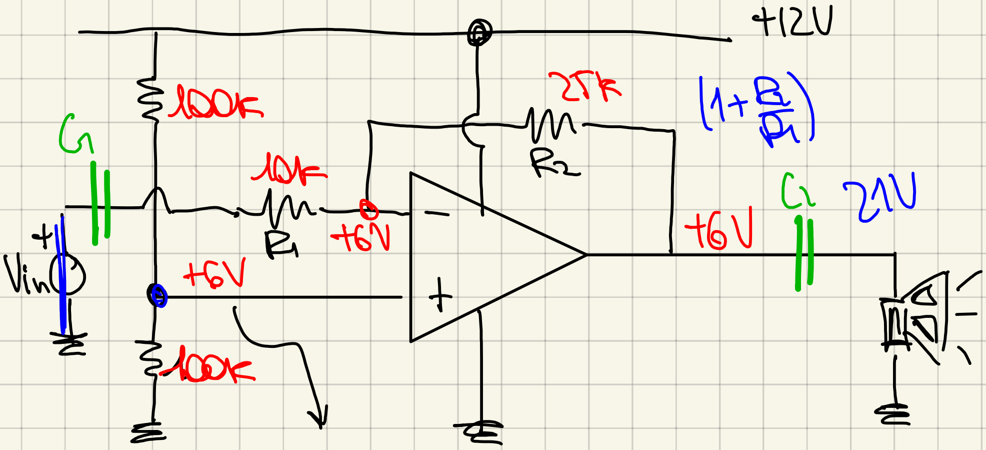
not possible !! ⇒ $\boxed{V_{in} \leq 0V}$

How can we solve this issue? How can we apply a dual-polarity signal to a single-polarity PS OpAmp?

① we want Vin (±2V) and Vout (±5V) ⇒ $|G| = \frac{5}{2} = 2.5$ ↓ desired amplification

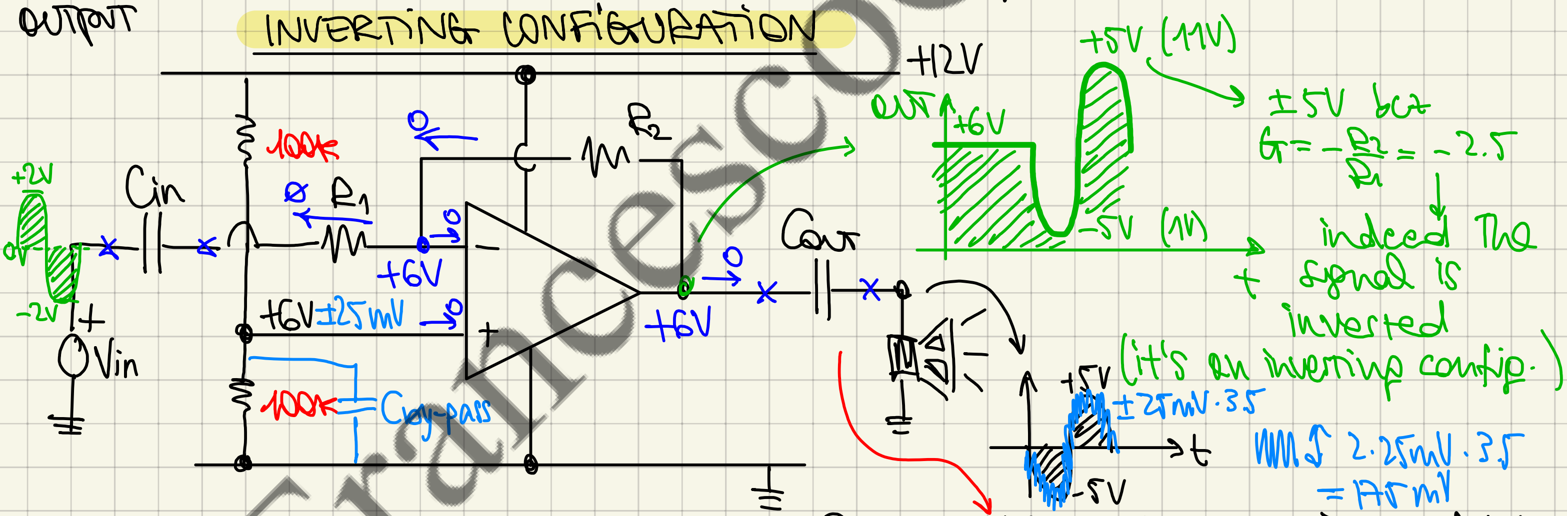


This configuration is not correct bc+ due to the PS limitation the output cannot reach −5V



we want to pre-bias the circuit @ half the dynamics

Now if we turn-off Vin, +6V will be amplified by $(1+R_2/R_1)$ (non-inverting stage) which is equal to 3.5 so at the output we would have 21V.
But the output cannot be @ 21V due to the PS limitation
↓
we do not want the output reaches an high value, we want that the output stays @ 0V if we apply 0V in input.
↓
to do that we need to place a capacitor @ the input and a capacitor @ the output

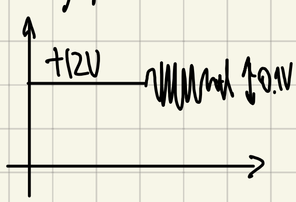


Cout is able to kill the DC and lets the signal to pass through

@DC → C open

In order to avoid any possible disturbance from the PS that can enter and gets amplified we have to add to the circuit a third capacitor Cby-pass

if the PS is affected by noise whose value is ↕ 0.1V peak-to-peak



so also V^+ is affected by noise whose value is ↕ 50mV peak-to-peak which means that $V^+ = +6V \pm 25mV$
↓
This variation gets amplified too and @ the output we would have a variation of $\pm 25mV\,(1+R_2/R_1) = \pm 25mV \cdot 3.5$ which means that @ the output we'll end up w/ a peak-to-peak variation of $2 \cdot 25mV \cdot 3.5 = 175mV$ ↕

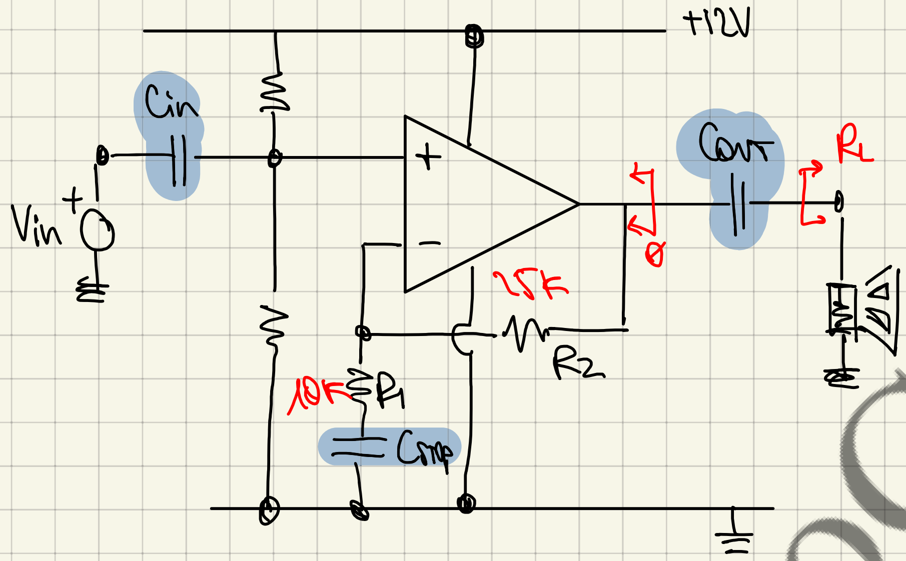
Let's suppose that this ripple is @ $f_{ripple} = 100Hz$

In order to kill it we should choose $C_{by\text{-}pass}$ whose pole is much lower than 1kHz, in such a way that @ $f = f_{ripple}$, $C_{by\text{-}pass}$ behaves as a short circuit

$$f_{P_{by\text{-}pass}} = \frac{1}{2\pi C_{by\text{-}pass}(100k \parallel 100k)} = \frac{1}{2\pi C_{by\text{-}pass} 50k} \ll f_{ripple}$$

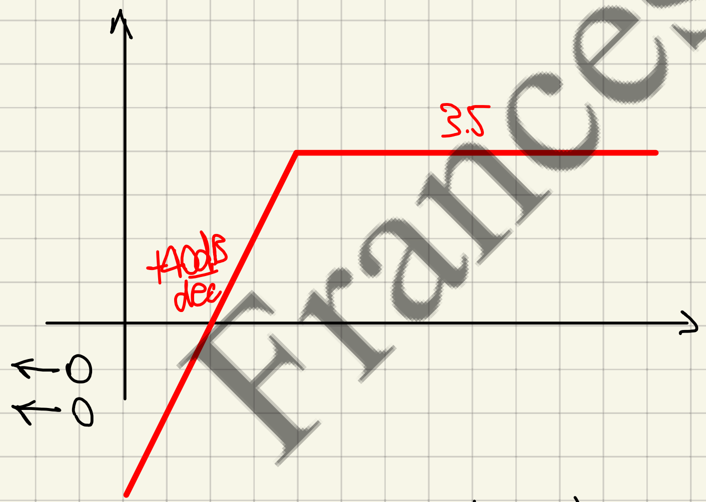
for instance let's choose $f_{P_{by\text{-}pass}} = 1Hz \rightarrow C_{by\text{-}pass} \gg \frac{1}{2\pi 50k \cdot 100Hz} \approx$

NON-INVERTING CONFIGURATION



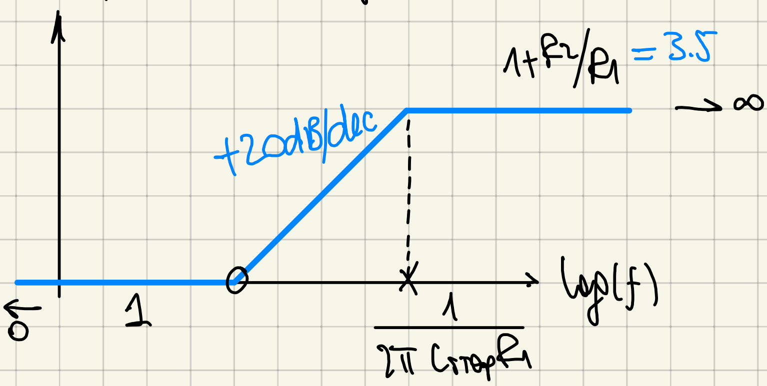
$$G(\infty) = 1 + \frac{R_2}{R_1} = 11$$

Both Cin and Cout, since they belong to the signal path, introduce a zero @ the origin (@ f=0)



HPF action

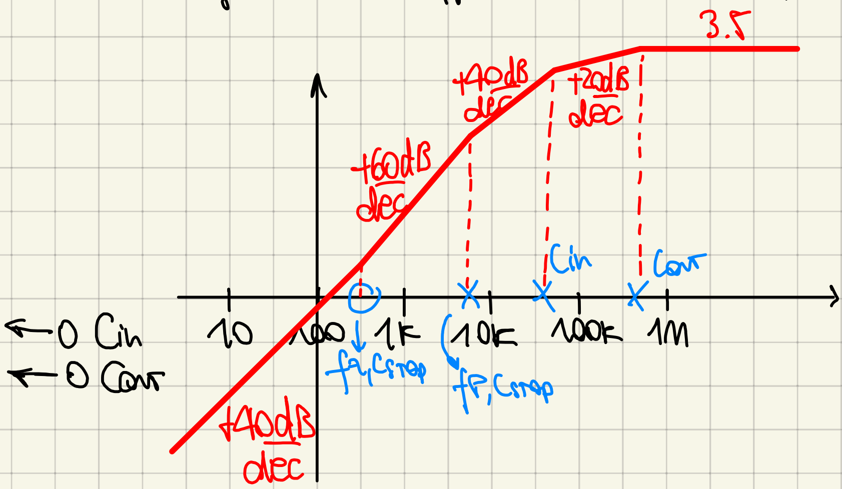
Cstop effect: (studying just Cstop)



+20dB/dec $\quad \frac{f_1}{A_1} = \frac{f_2}{A_2}$

$$\frac{f_{zero}}{1} = \frac{f_{pole}}{3.5}$$

let's sum together the effects of all the capacitors:



$$f_{p,C_{in}} = \frac{1}{2\pi C_{in}(100k \parallel 100k)} =$$

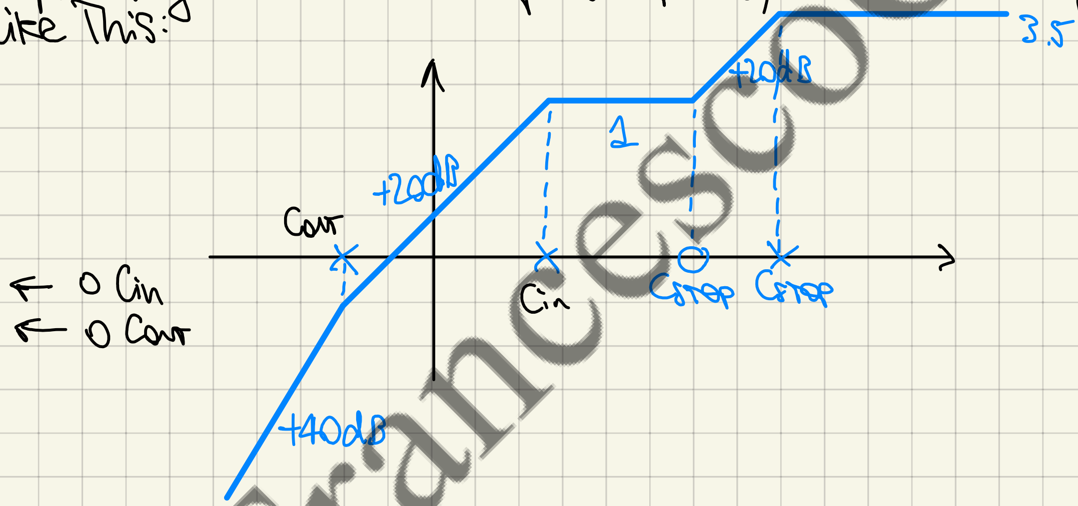
$$f_{p,C_{out}} = \frac{1}{2\pi C_{out} R_L} =$$

$$f_{p,C_{stop}} = \frac{1}{2\pi C_{stop} R_1} =$$

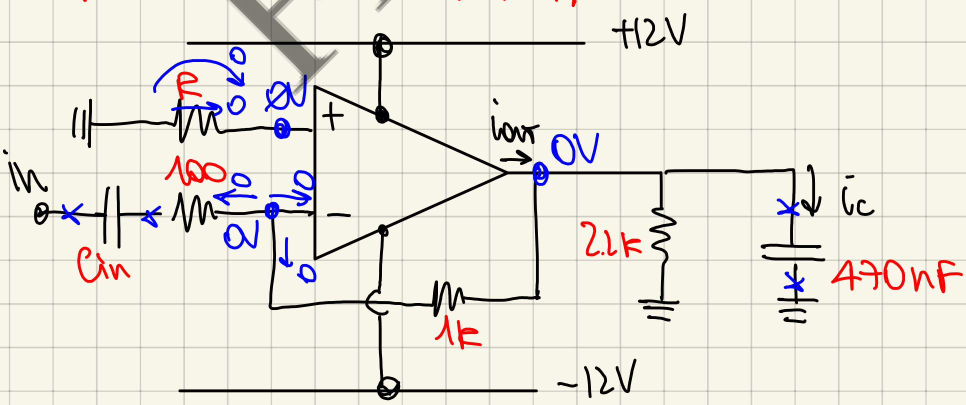
$$G(\infty) = 1 + \frac{R_2}{R_1} = 3.5$$

$$G(0) = 1$$

Depending on the values of the poles, the Bode diagram could be also like this:

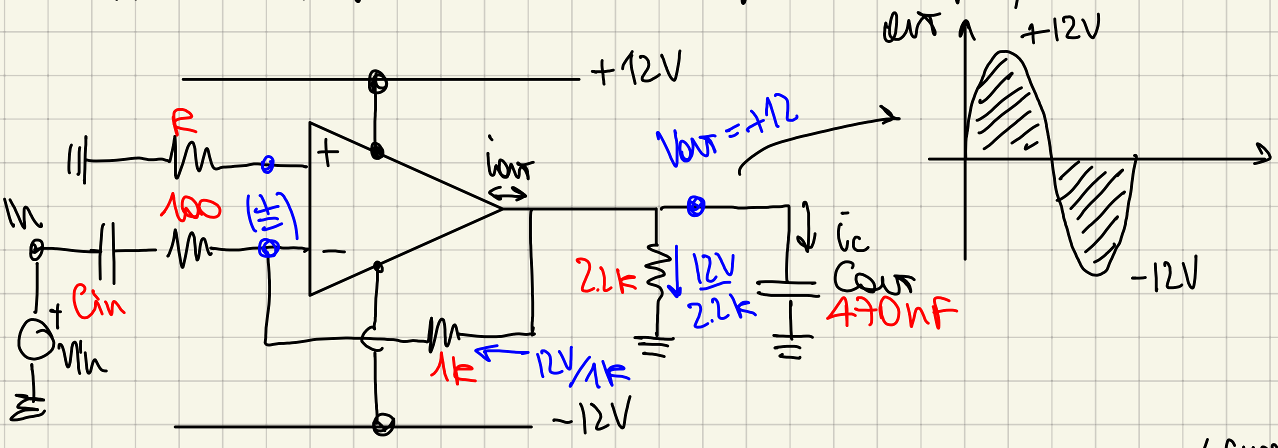


OUTPUT DRIVING CAPABILITY



@DC → $V_{out} = 0V$ → $G(0) = 0$

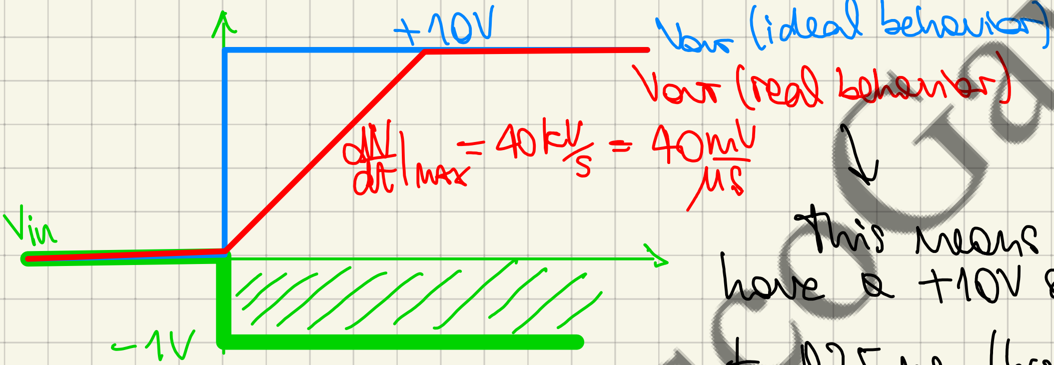
→ no current flows in the circuit → no PC (power consumption) let's suppose to apply the max possible signal in input, @ the output will have:



Resistive (DC) component: $I_{OUT,MAX} = \pm\left(\frac{12V}{1k} + \frac{12V}{2.2k}\right) = \pm 18mA$ (current that the OpAmp is able to provide w/o Cout)

Now if we connect Cout

$$i_c = C\frac{dV_c}{dt} \rightarrow \left.\frac{dV_{OUT}}{dt}\right|_{MAX} = \frac{I_{OUT,MAX}}{C_{OUT}} = \frac{18mA}{470nF} \simeq 40\frac{mV}{\mu s} = 40\frac{V}{ms}$$



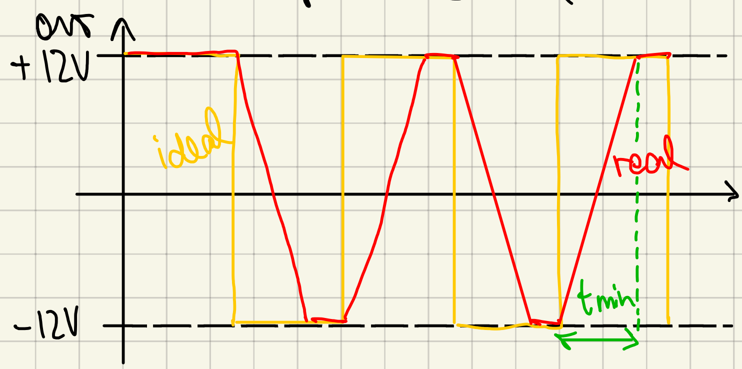
This means that if we want to have a +10V swing, it will take $t = 0.25\ ms$ $\left(bcz \left.\frac{dV_{OUT}}{dt}\right|_{MAX} = 40V/ms\right)$

CONCLUSION: don't place any capacitor @ the output of the OpAmp, bcz if we place a capacitor we'll cause the OpAmp to provide more current, so we'll run the risk to slow down the transition of the stage

- $I_{OUT,MAX}$ limits also the voltage swing, causing a saturation of the slope of the output swing

ex: $I_{OUT,MAX} = \pm 20mA \rightarrow \left.\frac{dV_{OUT}}{dt}\right|_{MAX} = \frac{I_{OUT,MAX}}{C_{OUT}} \simeq \frac{20mA}{0.47\mu F} = 40\frac{mV}{\mu s}$

Now let's suppose that we want that Vout goes up and down bcz we want to transmit a digital signal (and the PS is ±12V)



Which is the max freq. of this digital transmission?

$$\left.\frac{dV}{dt}\right|_{MAX} = \frac{I_{OUT,MAX}}{C_{OUT}} = \frac{\Delta V}{\Delta t} = \frac{24V}{t_{min}}$$

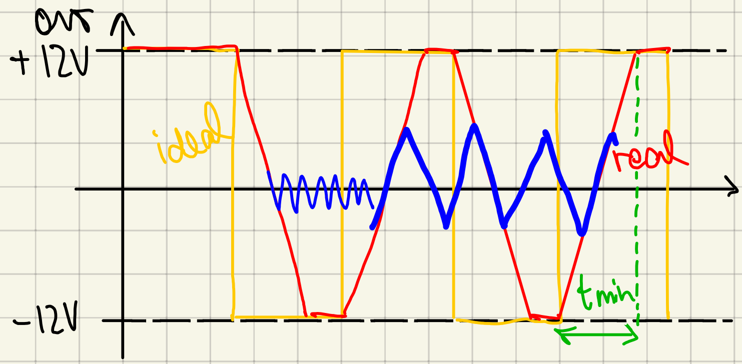
$$t_{min} = \frac{24V \cdot C_{OUT}}{I_{OUT,MAX}}$$

$\rightarrow f_{MAX} = \frac{1}{2 t_{min}} = \frac{I_{OUT,MAX}}{2 \cdot C_{OUT}\, \Delta V_{OUT,MAX}} = 833\,Hz$ (max freq. for digital application)

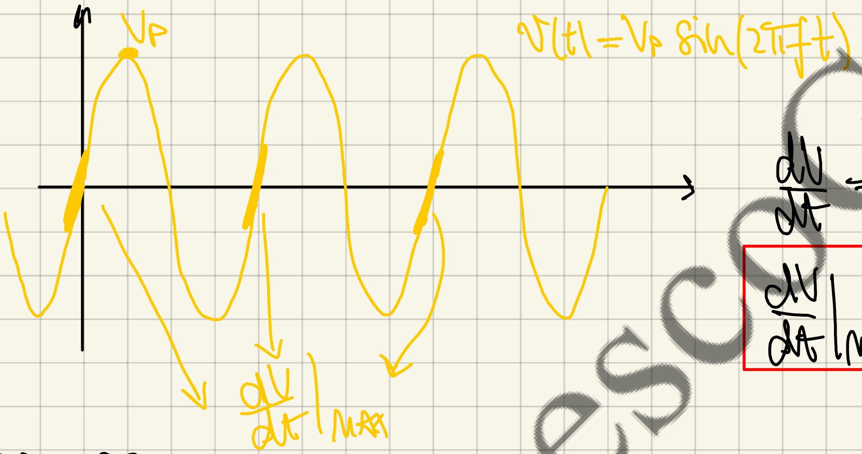
$\Delta V_{OUT,MAX} \rightarrow 24V$

What happens if we apply higher freqs.? * → higher freq = faster circuit

BUT The output swing will be very much reduced!! ☹



Finally, if we want to apply a sinusoid in input and that the output follows this sinusoid
↓
we have to compute $dV/dt|_{max}$ of the sinusoid



$$\frac{dV}{dt} = V_P\, 2\pi f \cdot \cos(2\pi f t)$$

max cos() value

$$\boxed{\frac{dV}{dt}\Big|_{MAX} = V_{P,MAX} \cdot 2\pi f_{MAX} \cdot 1}$$

TRADE-OFF:

This equation tells us again that if we want to have an high value for $V_{P,MAX}$, f_{MAX} should be low, and vice versa if we want to have an high f_{MAX}, $V_{P,MAX}$ should be low.

Which is the max freq. in order to reach the max amplitude @ the output of the OpAmp?

The maximum amplitude if of course the PS (+12V) and the maximum freq. that allows us to reach the PS @ the output is called FULL POWER BANDWIDTH (FPBW):

$$\frac{dV_{OUT}}{dt}\Big|_{MAX} = 2\pi f_{MAX} V_{P,MAX} = \frac{I_{OUT,MAX}}{C_{OUT}}$$

$$\rightarrow \boxed{FPBW = \frac{I_{OUT,MAX}}{2\pi \cdot C_{OUT} \cdot \underbrace{V_{OUT,MAX}}_{V_{alim}\ (=+12V)}}}$$

→ The FPBW represents the maximum freq. of a sinusoidal input that causes a sinusoidal output w/ the maximum amplitude!!

Again if our OpAmp has a FPBW = 1kHz this does not mean that it cannot operate @ an higher freq., but it means that if it operates @ an higher freq. the output cannot reach the FS, but it'll be reduced!

Finally, an OpAmp has not just a limitation due to $I_{OUT,MAX}$, but it also has an intrinsic limitation due to the SLEW RATE

The SLEW RATE of an OpAmp is the maximum output speed

$$\boxed{SR = \left.\frac{dV_{OUT}}{dt}\right|_{MAX}} = \frac{I_{OUT,MAX}}{C_{OUT}}$$

$$\Rightarrow \boxed{f_{MAX,\,slewrate} = \frac{SR}{\Delta V_{OUT,MAX}}}$$

$$\boxed{FPBW = \frac{SR}{2\pi V_{OUT,MAX}}}$$

CONCLUSION: Given an OpAmp the output can move w/ a maximum slope dV/dt set by either the SR or $I_{OUT,MAX}$.

- if there is no C_{OUT} → there is no limitation due to $I_{OUT,MAX}$
  → the limitation is given by SR
- if there is C_{OUT} → the limitation is given by the most limiting factor
  - if $\frac{I_{OUT,MAX}}{C_{OUT}} < SR \Rightarrow \left.\frac{dV_{OUT}}{dt}\right|_{MAX} = \frac{I_{OUT,MAX}}{C_{OUT}}$
  - if $SR < \frac{I_{OUT,MAX}}{C_{OUT}} \Rightarrow \left.\frac{dV_{OUT}}{dt}\right|_{MAX} = SR$