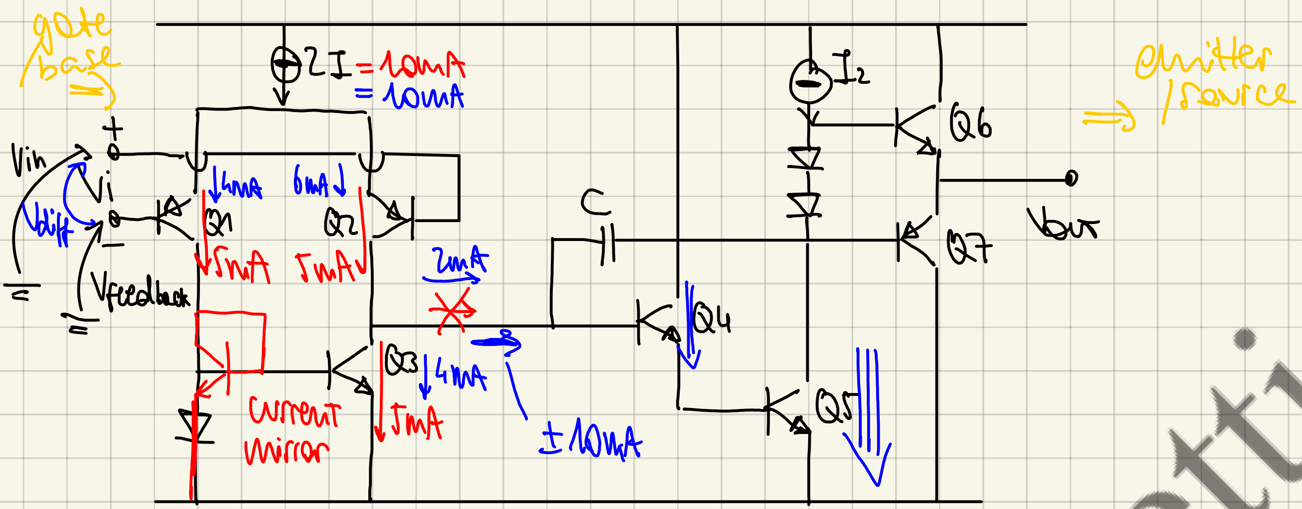
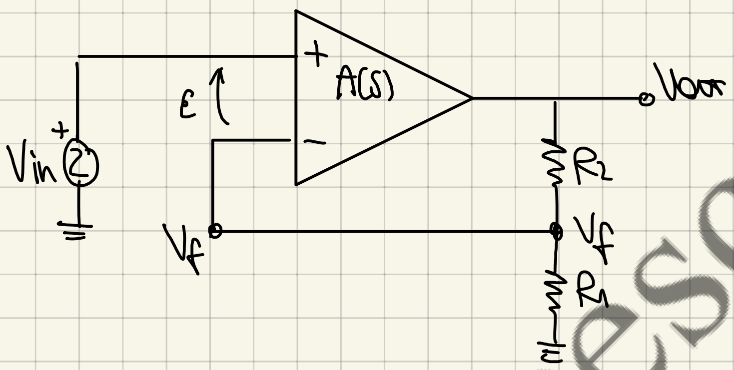
ES07 - CFA

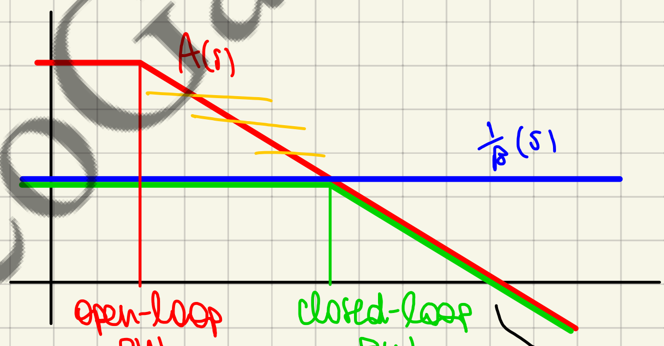
20/10/2021

VOLTAGE MODE OPAMP (VOA - VOLTAGE OPERATING AMPLIFIER)



REQUIREMENTS:
- voltage-driven inputs → high impedance inputs
- voltage output → low impedance output





$V_o = \varepsilon \cdot A_0 = 0 \cdot \infty$

$\varepsilon \to 0$

$\Rightarrow V_f \to V_{in}$ (virtual ground)

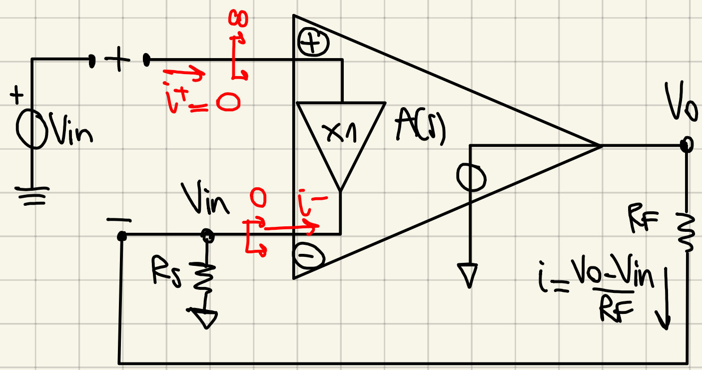
$\Rightarrow V_{out} \frac{R_1}{R_1+R_2} = V_f = V_{in} \Rightarrow \boxed{G_{VOA} = 1 + \frac{R_2}{R_1}}$

Performances

−20dB/dec
constant GBWP

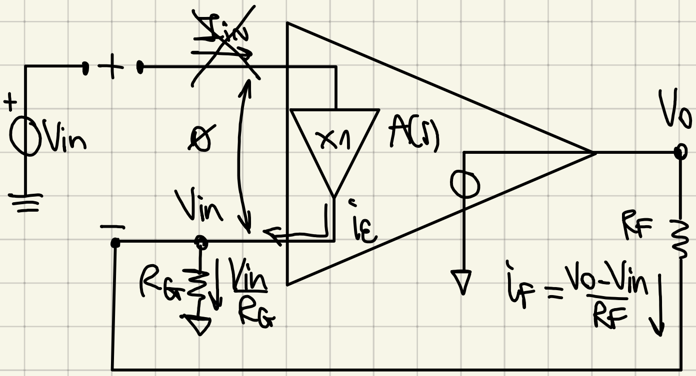
GBWP remains constant
↓
TRADE-OFF GAIN−BW

CURRENT FEEDBACK AMPLIFIER (CFA)



REQUIREMENTS:
- one high-impedance input (hence voltage driven input)
- one low-impedance input (hence input-current, i.e. acting as an output)
- voltage output (proportional to the input current)

CFA GAIN $G_{CFA} = 1 + \frac{R_F}{R_S}$



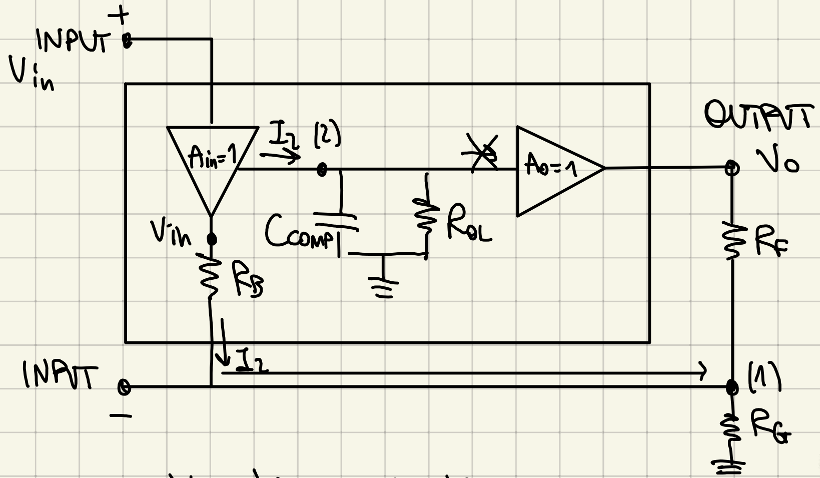
$V_o = i_c \cdot R_{OL} = \emptyset \cdot \infty$

ideally: $i_c \to \emptyset$ → Thanks to FB input buffer can provide current, but it's not needed thanks to feedback

$\Downarrow$

$i_F \to \frac{V_{in}}{R_G}$

$$\frac{V_{in}}{R_G} = \frac{V_o - V_{in}}{R_F} \Rightarrow \frac{V_o}{V_{in}} = \frac{1}{R_G} + \frac{1}{R_F} = \frac{R_F + R_G}{R_G R_F}$$



$$(1) \quad I_2 = \frac{V_{in} - V_1}{R_B} = \frac{V_1 - V_{OUT}}{R_F} + \frac{V_1}{R_G} = V_1\left(\frac{1}{R_G} + \frac{1}{R_F}\right) - \frac{V_{OUT}}{R_F}$$

$$(2) \quad V_2 = I_2\left(\frac{1/sC \cdot R_{OL}}{1/sC + R_{OL}}\right) = I_2 \frac{R_{OL}}{1 + sC_{comp}R_{OL}} = V_o$$

Doing computations we end up w/:

CLOSED LOOP GAIN

$$\frac{V_{OUT}}{V_{in}} = \frac{1 + R_F/R_G}{\left(1 + \frac{R_F + \left(1 + \frac{R_F}{R_G}\right)R_B}{R_{OL} \cdot A_{OUT}}\right)\left[1 + s\frac{\left[R_F + \left(1 + \frac{R_F}{R_G}\right)R_B\right]}{A_{OUT} + \frac{R_F(1 + R_F/R_G)R_B}{R_{OL}}} \cdot C_{comp}\right]}$$

$$f_{POLE} \approx \frac{A_{OUT}}{2\pi\left[R_F + \left(1 + \frac{R_F}{R_G}\right)R_B\right]C_{comp}}$$

- for low gain: $\left(R_B \ll \frac{R_F R_G}{R_F + R_G} = R_F \| R_G\right) \to f_{POLE} \approx \frac{1}{2\pi R_F C_{comp}}$

→ BW depends only on R_F, not on R_G, hence not on gain!

- for high gain (>50) → $GBWP = \frac{A_{OUT}}{2\pi R_B \cdot C_{comp}}$

→ As for VOA, again GBWP stays constant → trade-off GAIN vs BW

By changing R_G we can change the gain of the stage w/o changing its frequency response, i.e. its BW

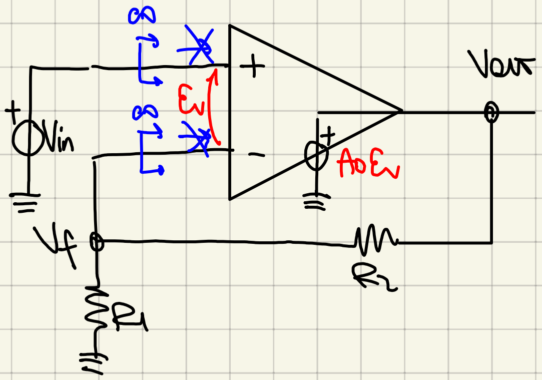
if we want to reach very high gain, since $G = 1 + R_F/R_G$, R_F should be high, while R_G should be small

if $R_S = 10\,\Omega \rightarrow$ set $R_G \cong 10\,R_S$ at least, but under this condition the gain cannot increase too much.

ES07 - CFA (2)
26/10/2021

VOA

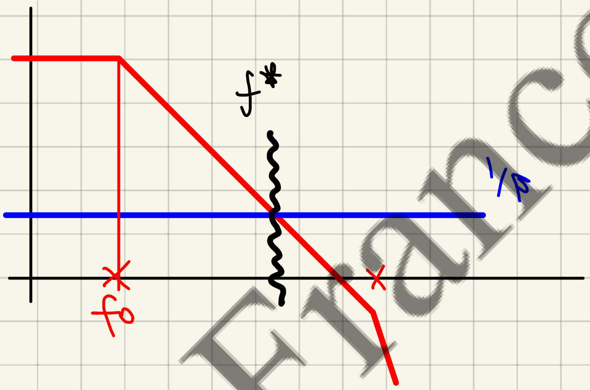
Up to now we have worked w/ VOAs, which is an amplifier w/ high input impedances (ideally ∞)



since $A_0 \rightarrow \infty \Rightarrow G_{loop} = A_0\beta \rightarrow \infty$
$\Rightarrow \varepsilon_v \rightarrow 0$
$\Rightarrow V_f \equiv V_{in}$
$V_{out}\dfrac{R_1}{R_1+R_2} = V_{in}$
$\Rightarrow G = \dfrac{V_{out}}{V_{in}} = 1 + \dfrac{R_2}{R_1}$

What about the freq. response?

$$\beta = \frac{R_G}{R_G + R_F} \Rightarrow \frac{1}{\beta} = 1 + \frac{R_F}{R_G}$$



If we change $G \Rightarrow$ we also change $1/\beta$

$\Rightarrow$ we also change f^*

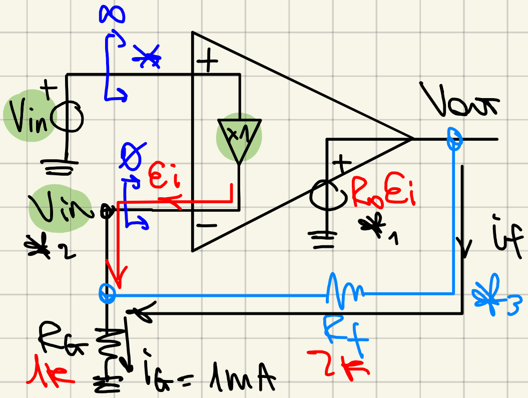
- if $G\uparrow \Rightarrow \frac{1}{\beta}\uparrow \Rightarrow f^*\downarrow$
- if $G\downarrow \Rightarrow \frac{1}{\beta}\downarrow \Rightarrow f^*\uparrow$

$\}$ TRADE-OFF GAIN − BW

for example:
- if $G = 10 \rightarrow f^* = 10\,MHz$
- if $G = 100 \rightarrow f^* = 1\,MHz$

$\}$ $G = \propto f^*$

CFA - CURRENT FEEDBACK AMPLIFIER



*1 $[A_0] = \Omega$ since the error is a current signal so we can call it R_o

+ input : $R_{in}^{+} = \infty \Rightarrow$ VOLTAGE READER

− input : $R_{in}^{-} = \varnothing \Rightarrow$

*2: here we have Vin thanks to the buffer, not due to the feedback, cause the feedback is not acting yet

if $V_{in} = 1V$ and $R_G = 1k \Rightarrow \varepsilon_i = \frac{V_{in}}{R_G} = 1mA = i_G$

$\Rightarrow$ if $R_o = 1M\Omega \Rightarrow V_{out} = 1M\Omega \cdot 1mA = 1kV$

Now we can introduce the FB *3
↓
A current flowing through R_f rises and it tends to go towards the ⊖ pin where the input resistance is nil
↓
if $R_o \to \infty \Rightarrow \varepsilon_i \to \varnothing$ thanks to the FB

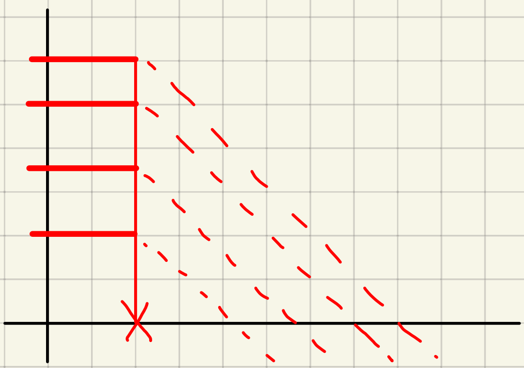
$$\Rightarrow V_{out} = R_o \cdot \varepsilon_i = [\infty]\cdot[0] = 1 + \frac{R_f}{R_G} = 3V$$

Vout does not saturate, but it settles to 3V in such a way that:
⇓
$i_f = \frac{3V - 1V}{2k} = \frac{2V}{2k} = 1mA = i_G$ → the current that the FB can provide is the same current that R_G requires
⇓
$\varepsilon_i = \varnothing$ even if the ⊖ input can provide current since $R_{in}^{-} = \varnothing$

- In the VOA, ε_v could be different to zero, but it isn't, but thanks to the FB $\varepsilon_v \to \varnothing$

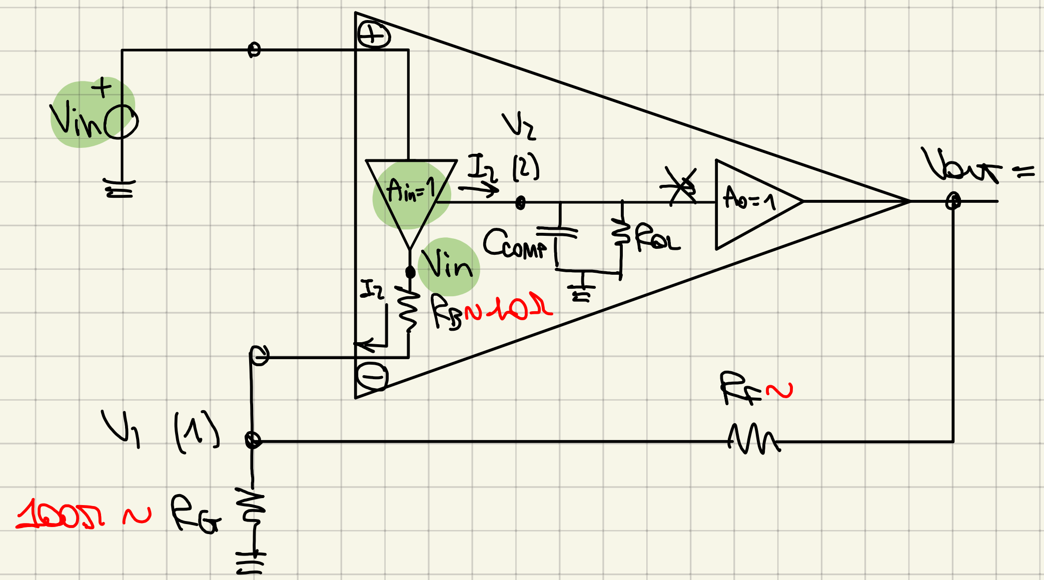
- In the CFA, ε_i could be different to zero, but it isn't, but thanks to the FB $\varepsilon_i \to \varnothing$

$$i_f = i_G \Rightarrow \frac{V_{out} - V_{in}}{R_f} = \frac{V_{in}}{R_G} \Rightarrow G = \frac{V_{out}}{V_{in}} = 1 + \frac{R_f}{R_G}$$



NOTICE (IMP!!): The big ADVANTAGE of the CFA is that you can adjust the gain as you wish but the pole remains at a constant frequency
→ to do so, we must change just R_G

In fact $f_p = \frac{1}{2\pi C R_f}$



(1) $$I_2 = \frac{V_{in} - V_1}{R_B} = \frac{V_1 - V_{OUT}}{R_F} + \frac{V_1}{R_G} = V_1\left(\frac{1}{R_G} + \frac{1}{R_F}\right) - \frac{V_{OUT}}{R_F}$$

(2) $$V_2 = I_2\left(\frac{\frac{1}{sC}\cdot R_{OL}}{\frac{1}{sC} + R_{OL}}\right) = I_2 \frac{R_{OL}}{1 + sC_{comp}R_{OL}} = V_0$$

Doing computations we end up w/:

CLOSED LOOP GAIN $$\frac{V_{OUT}}{V_{in}} = \frac{1 + R_F/R_G}{\left(1 + \frac{R_F + \left(1+\frac{R_F}{R_G}\right)R_B}{R_{OL}\cdot A_{OUT}}\right)\left[1 + s\frac{\left[R_F + \left(1+\frac{R_F}{R_G}\right)R_B\right]}{A_{OUT} + \frac{R_F\left(1+R_F/R_G\right)R_B}{R_{OL}}}\cdot C_{comp}\right]}$$

CFA BANDWIDTH

$$f_{ROLL} \approx \frac{A_{OUT}}{2\pi\left[R_F + \left(1+\frac{R_F}{R_G}\right)R_B\right]C_{comp}}$$

- for low gain: $\left(R_B \ll \frac{R_F R_G}{R_F + R_G} = R_F \| R_G\right) \rightarrow f_{ROLL} \approx \frac{1}{2\pi R_F C_{comp}}$

→ BW depends only on R_F, not on R_G, hence not on gain!

- for high gain (>50) → $GBWP = \frac{A_{OUT}}{2\pi R_B \cdot C_{comp}}$

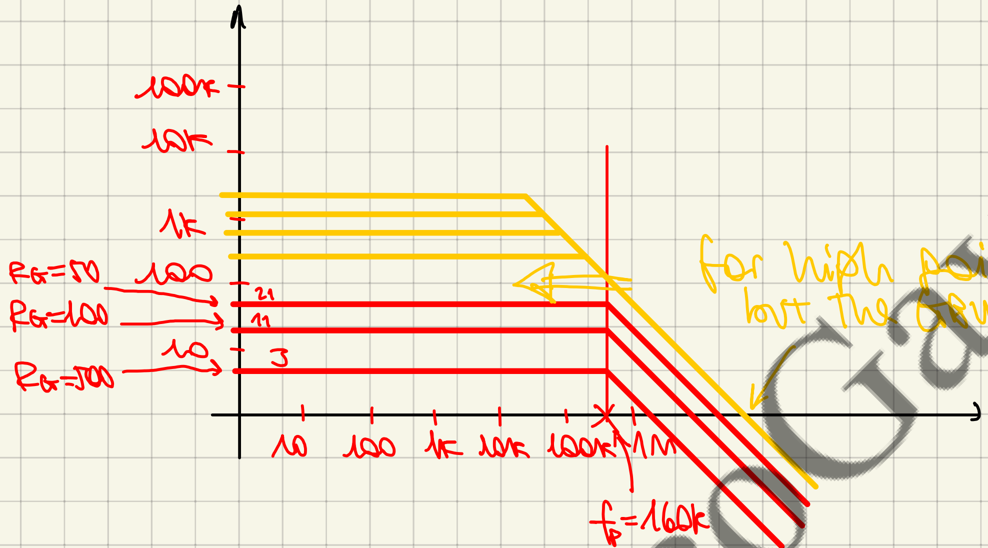
→ As for VDA, again GBWP stays constant → trade-off GAIN vs BW let's study the FB in a smarter way:

- open loop: $pole_{OL} \approx \frac{-1}{C_{comp} R_{OL}}$ $\qquad G_{loop} \approx -R_{OL} \cdot A_{OUT} \frac{1}{R_B + R_F} \approx -\frac{R_{OL}}{R_F}$

- closed loop: $pole_{CL} = pole_{OL}(1 - G_{loop}) \approx \frac{-1}{C_{comp} R_{OL}} \cdot \frac{R_{OL}}{R_F} = \frac{-1}{C_{comp} \cdot R_F}$

we set $C_{comp} = 1pF$ $R_{OL} = 1M\Omega$ $R_F = 1k\Omega$ $R_B = 10\Omega$

$f_p = \frac{1}{2\pi C_{comp} R_F} \simeq 160kHz$



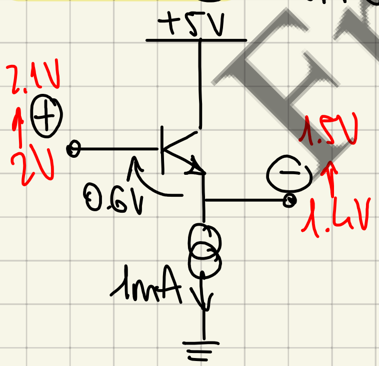
if $R_G = 1\Omega \Rightarrow f_p$ doesn't stay @ 160kHz anymore

Why? Bcz if R_G becomes comparable to R_B, the gain increases too much and f_p is no more constant, but the GBWP becomes constant

REAL ARCHITECTURE OF THE CFA

STEP 1: we need to design a BUFFER

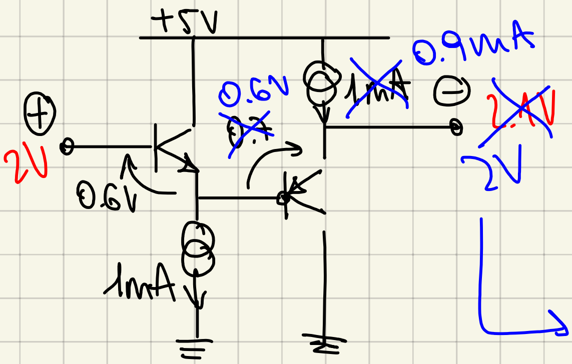
SOLUTION ① BUFFER = EMITTER FOLLOWER



⇒ in this way we loose 0.6V ⇒ There's 0.6V of offset (600mV)

ISSUE!

SOLUTION ②



⇒ There's still a 100mV offset

⇓

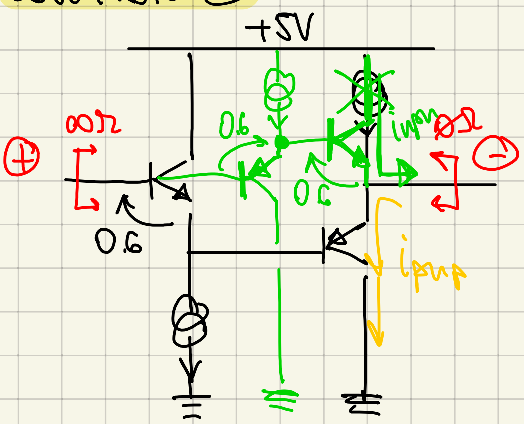
* we can erase it by changing the current

NOW IT'S PERFECT!! (OS=0)

There's still an issue → in this way we cannot provide more than 0.9 mA to the output

→ if the output drinks more than 0.9 mA ⇒ the pnp transistor interdicts (goes off!) ☹

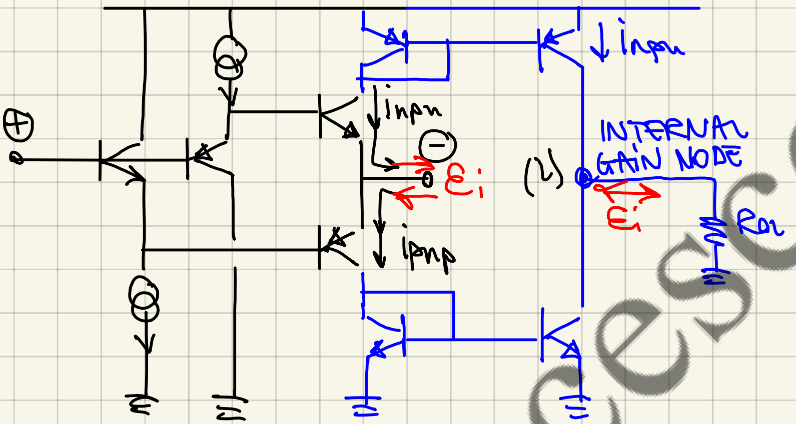
SOLUTION ③



⇒ This is a symmetric voltage buffer

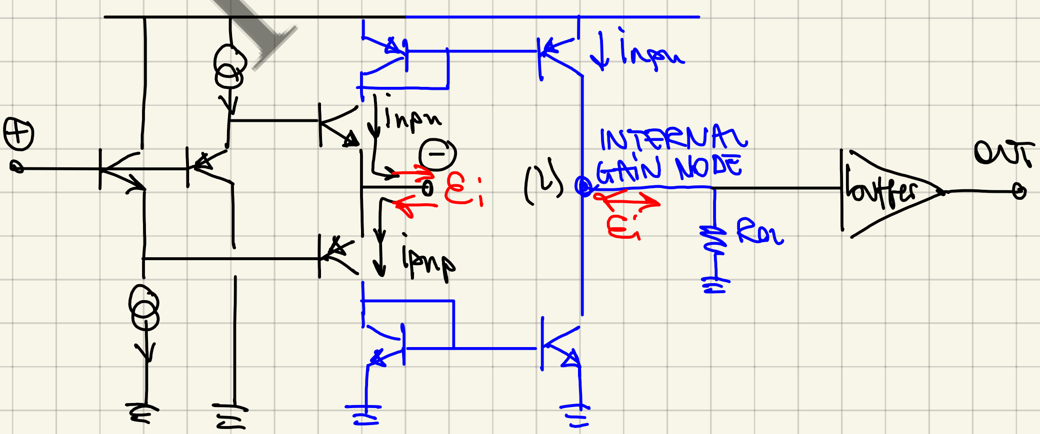
STEP 2: we need to make a copy of i_{pnp} to feed into R_{OL}

SOLUTION ①: we use a current mirror

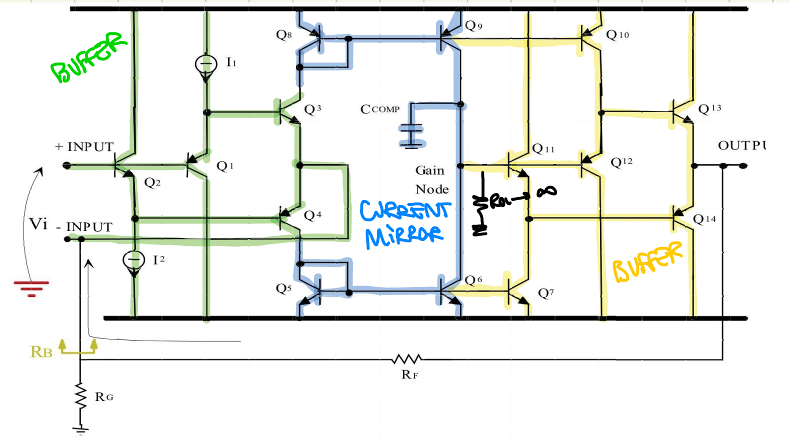


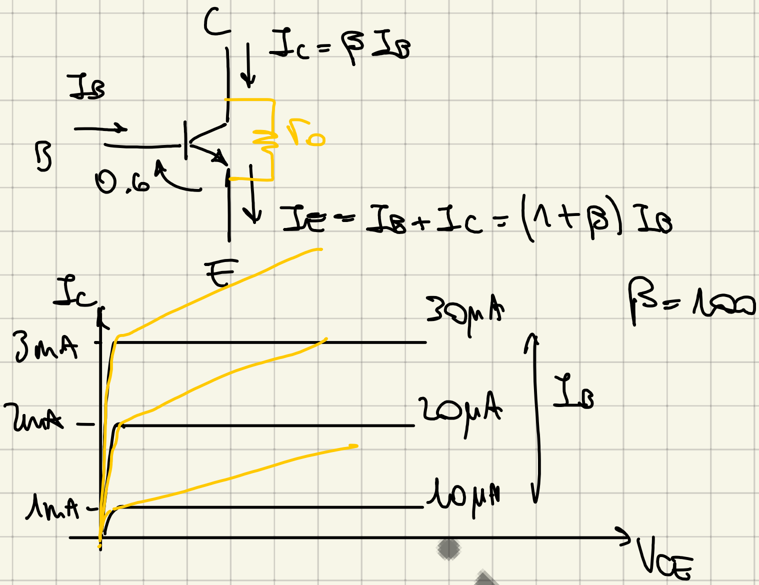
STEP 3: we want to provide the voltage across R_{OL} to the output w/o drinking current from node (2)

How? Again, using a buffer! (which can be the same of the input one)



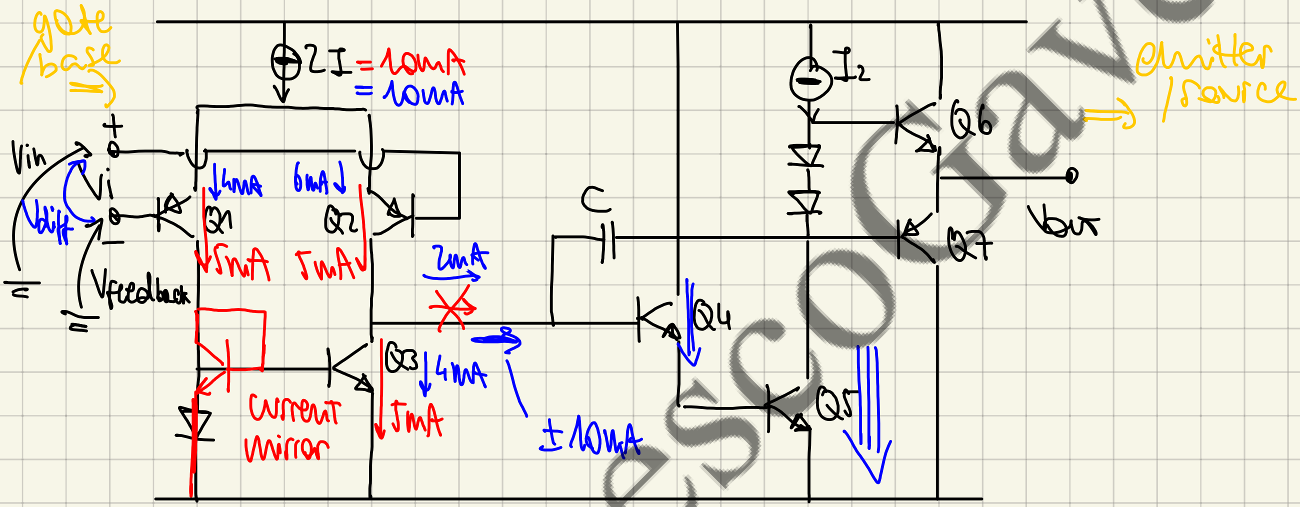
In conclusion we have:





The CFA is much better than the VOA even in terms of SLEW RATE:

VOA SR

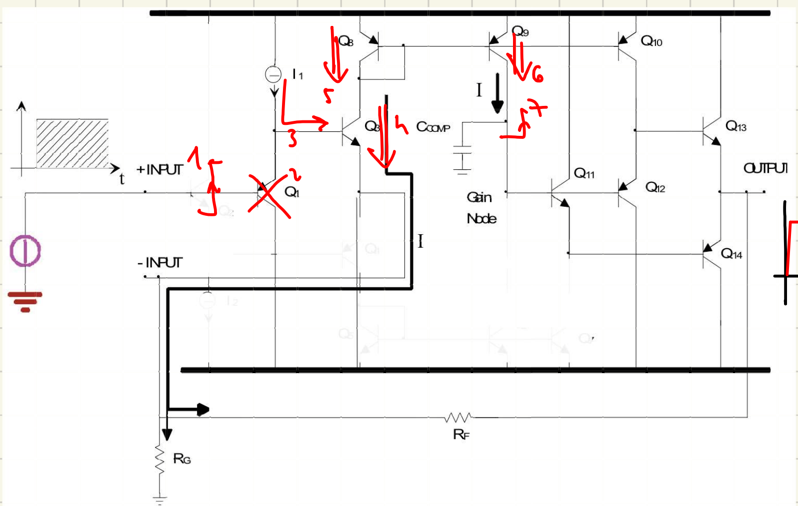


$$SR = \left.\frac{\Delta V_{out}}{\Delta t}\right|_{MAX} = \frac{I}{C}$$ and since both I and C are fixed by the manufacturer we cannot change the SR of a VOA

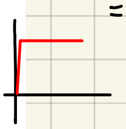
$\hookrightarrow \sim 10 \div 100 \, V/\mu s$

CFA: TRANSIENT RESPONSE → SLEW RATE

Now, let's imagine to apply a step at the input of the CFA:



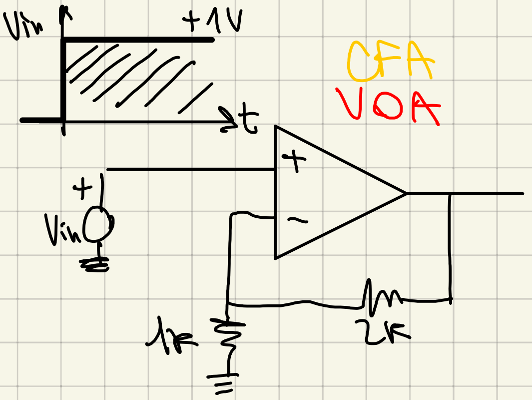
$$SR = \left.\frac{dV}{dt}\right|_{MAX} = \frac{I_{MAX}}{C_{comp}} \approx \frac{V_{step}/R_F}{C_{comp}} =$$

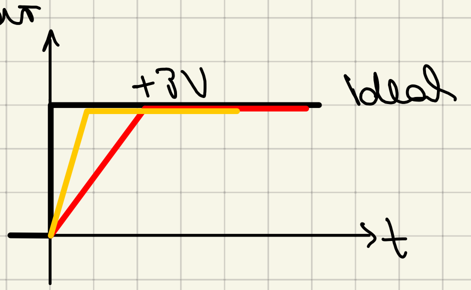


$$= \frac{1mA}{1pF} = \frac{50kV}{\mu s}$$

⇒ In a CF, in principle, the SR can be much faster than in a VOA

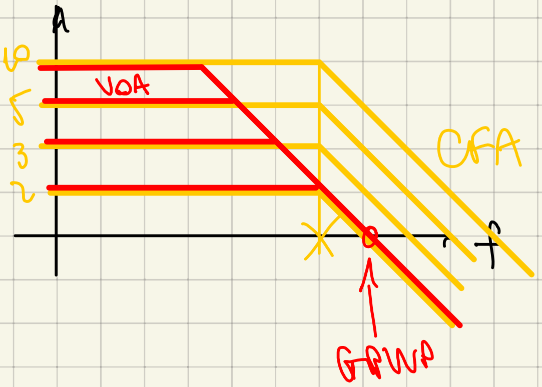
COMPARISON:





$SR_{CFA} \sim 1k \div 100\ kV/\mu s$

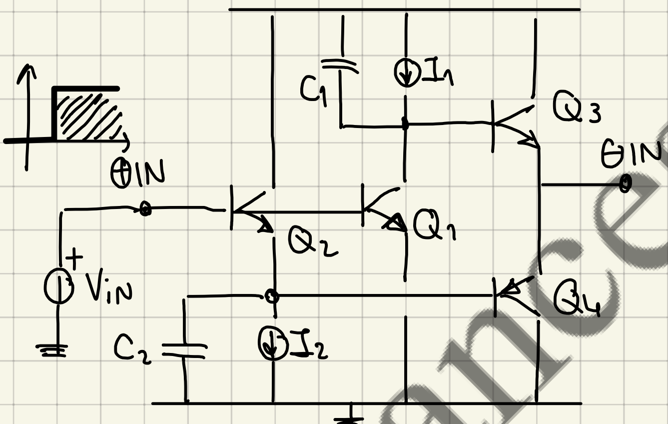
$SR_{VOA} \sim 10 \div 100\ V/\mu s$

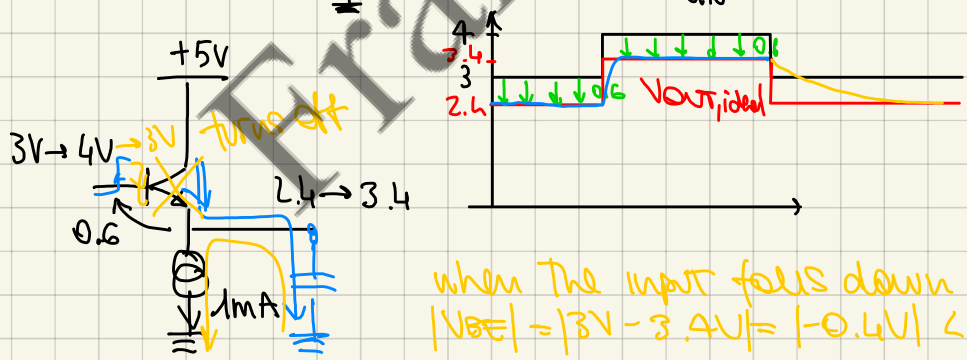


At a freq. stand point:

- CFA → the pole remains constant (for a low gain
- VOA → the GBWP stays constant

Unfortunately there are parasitisms that slow down also the CFA:

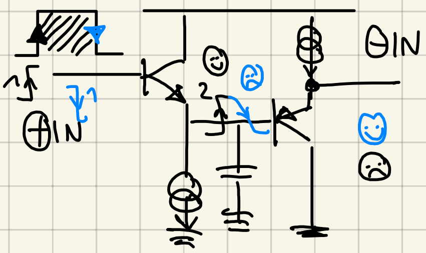




when the input falls down from 4V to 3V
$|V_{BE}| = |3V - 3.4V| = |-0.4V| < 0.6V$ ⇒ the transistor turns off

⇒ The capacitor can be discharged only through the current generator

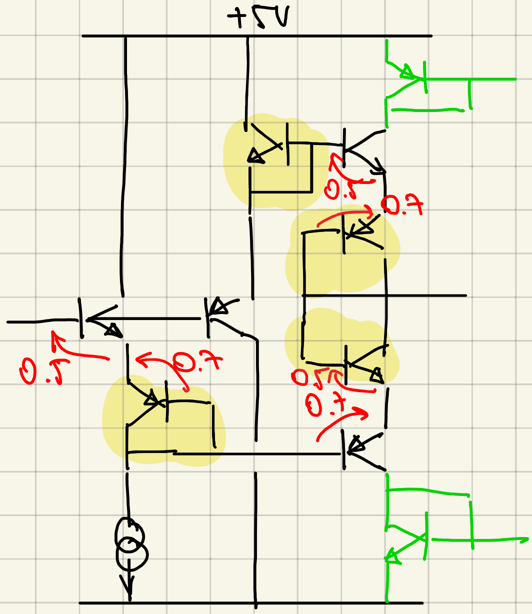
let's add another transistor:
↓
a pnp transistor which is a slow transistor, so it's happy during the slow falling down



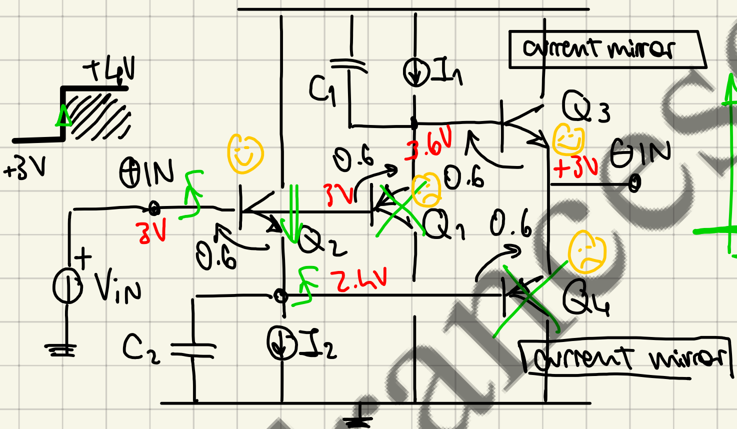
Due to parasitisms the CFA is not as ideal as we supposed to be
↓
The parasitisms slow down both the pull-up and the pull-down

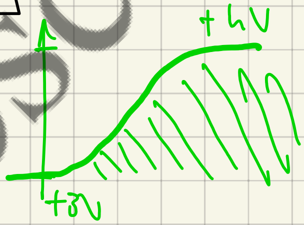
How can we solve it?
We have to introduce some improvements:

① Introduce transdiodes:

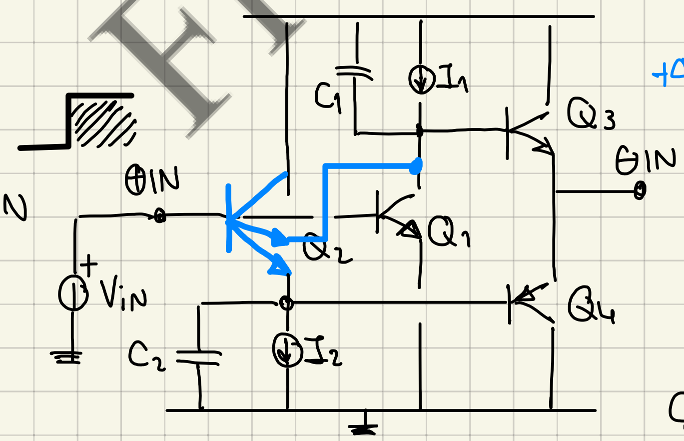


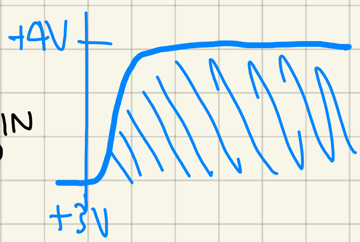
② Use multi-emitter transistors





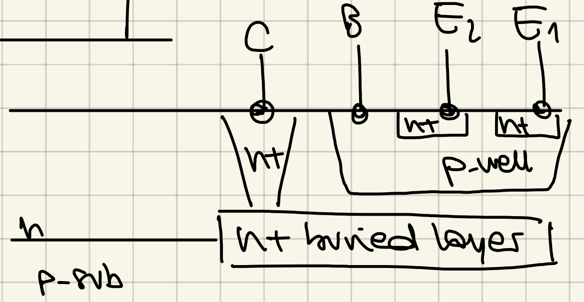
solution

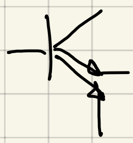




The output will speed up

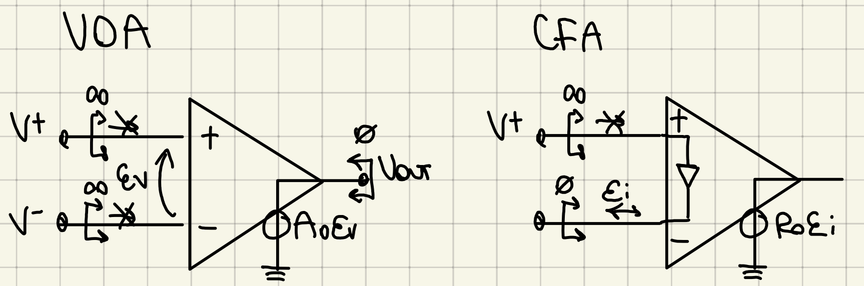
What is a multi-emitter transistor??



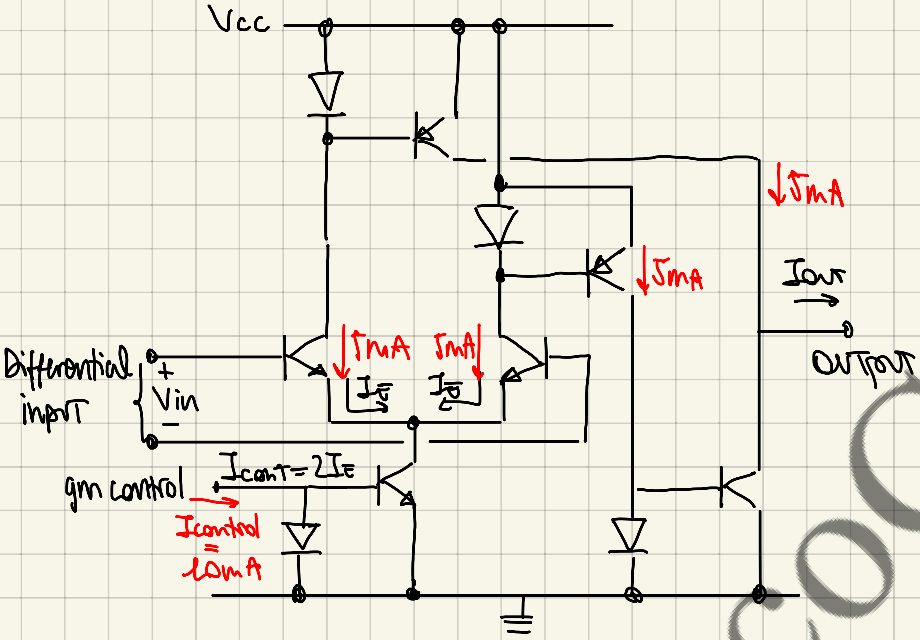


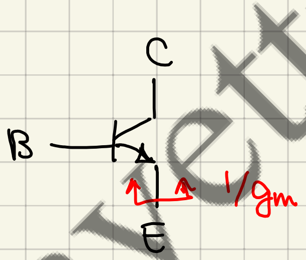
ES08 - OTA and ISO AMPLIFIERS

26/10/2021



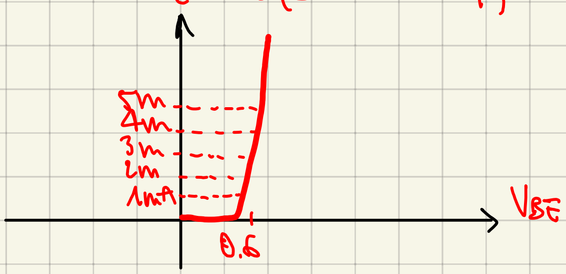
OTA : OPERATIONAL TRANSCONDUCTANCE AMPLIFIER





TRANSCONDUCTANCE $\frac{dI_C}{dV_{BE}} = g_m = \frac{I_C}{KT/q}$

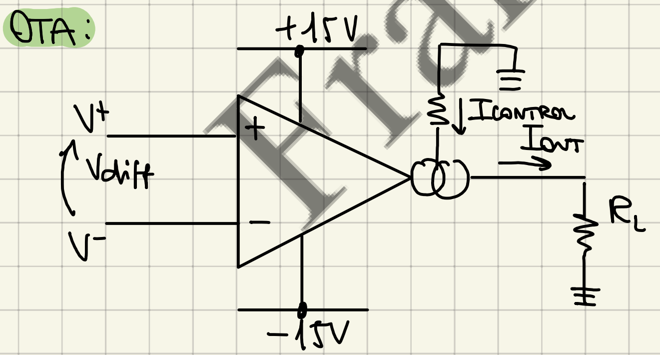
$I_C = I_S\left(e^{V_{BE}/KT/q} - 1\right)$



OTA GAIN: $\boxed{g_m = \frac{dI_{OUT}}{dV_{diff}} = \frac{dI_{OUT}}{d(V^+ - V^-)} = \frac{I_{control}}{KT/q}}$ [A/V]

where I_E is the current we force in the input stage

OTA:



$i_{OUT} = G_m \cdot V_{diff} = \frac{I_{control}}{KT/q} \cdot V_{diff} = \frac{I_{control}\, V_{diff}}{25mV}$

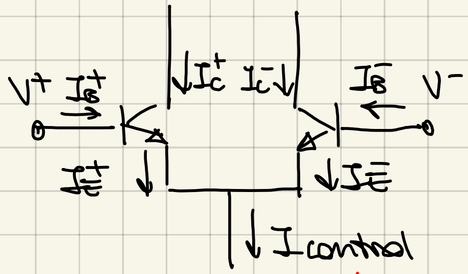
$\frac{KT}{q}(@RT) = 25mV$

$$\boxed{G_m = \frac{I_{OUT}}{V_{diff}} = \frac{I_{control}}{KT/q}}$$

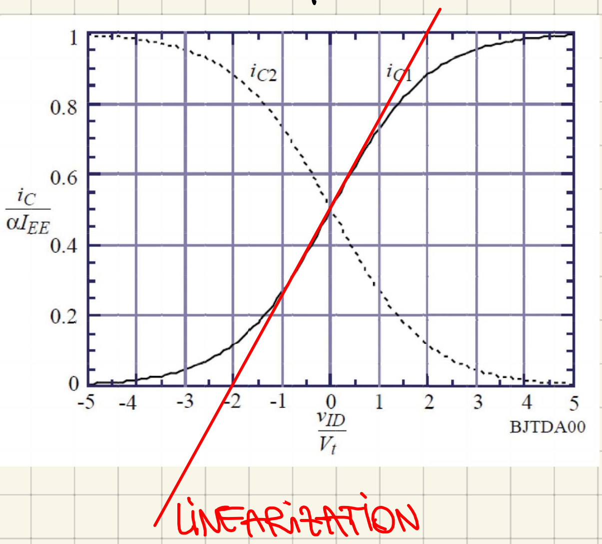
ADVANTAGES:
- all low-impedance node inside the OTA
  ↳ This means that any parasitism will introduce very low spurious time constants $\tau = RC$ is low since R is low
- wide BW (low τ → high freq poles)
- $f_{POLE} = 1/2\pi C_L R_L$

DISADVANTAGES

- no infinite gain
- no virtual ground
- it is used open loop (but Gloop is too poor)



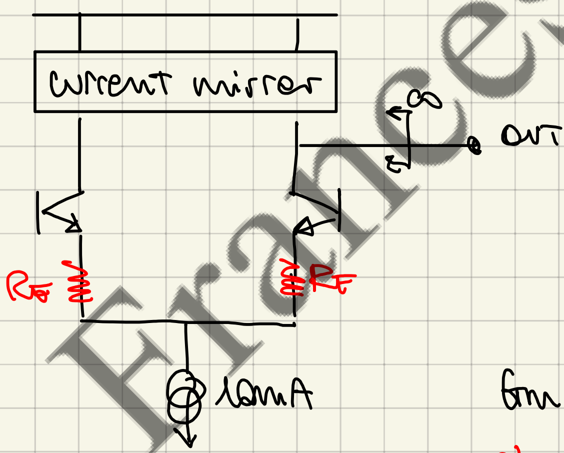
$$\boxed{I_{OUT} = I_C^+ - I_C^- = I_{control} \cdot \tanh\left(\frac{V_{diff}}{2kT/q}\right)}$$

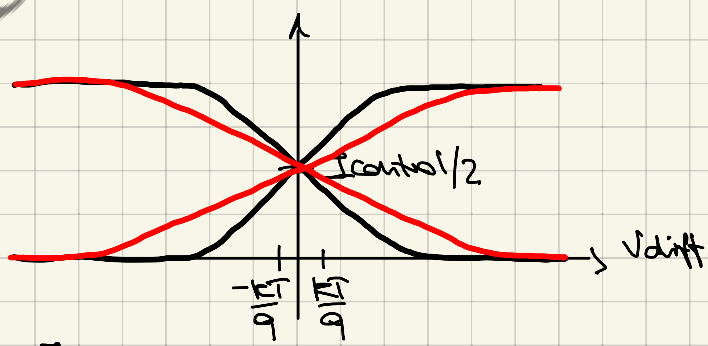


If $V_{diff} = 0 \Rightarrow I_C^+ = I_C^-$

up to $|V_{diff}| \leq 2U \Rightarrow$ linear stage

How can we increase the linearity of this stage?





$G_m = I_{control}/V_{th} = g_m$

$$G_m = \frac{1}{1/g_m + R_E}$$

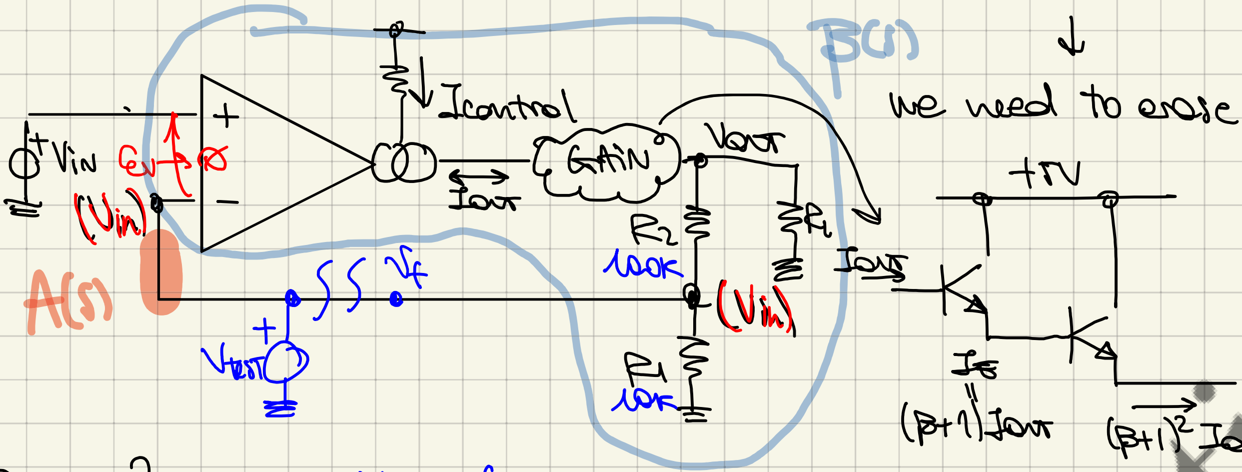
Another way to improve linearity is adding diodes into the configuration

EXAMPLE OF OTA APPLICATION: VOLTAGE-CONTROLLED LOW-PASS FILTER

I want to use the OTA in a FB config. but we can't b/c G_{loop} is too poor

↓

we need to erase it, to do so



$G_{loop} = ?$ $\quad I_{OUT} = V_{test}\, G_m$

$$V_{OUT} = V_{TEST}\, G_m (\beta+1)^2 \cdot \left[R_L \| (R_1+R_2) \right]$$

$$V_f = V_{OUT} \frac{R_1}{R_1+R_2} = V_{TEST}\, G_m (\beta+1)^2 \cdot \left[R_L \| (R_1+R_2) \right] \frac{R_1}{R_1+R_2}$$

$$G_{loop} = G_m (\beta+1)^2 \cdot \left[R_L \| (R_1+R_2) \right] \frac{R_1}{R_1+R_2} = \frac{I_{control}}{KT/q} (10'000)\ 1k\Omega\ \frac{10K}{110K} >>> 1$$

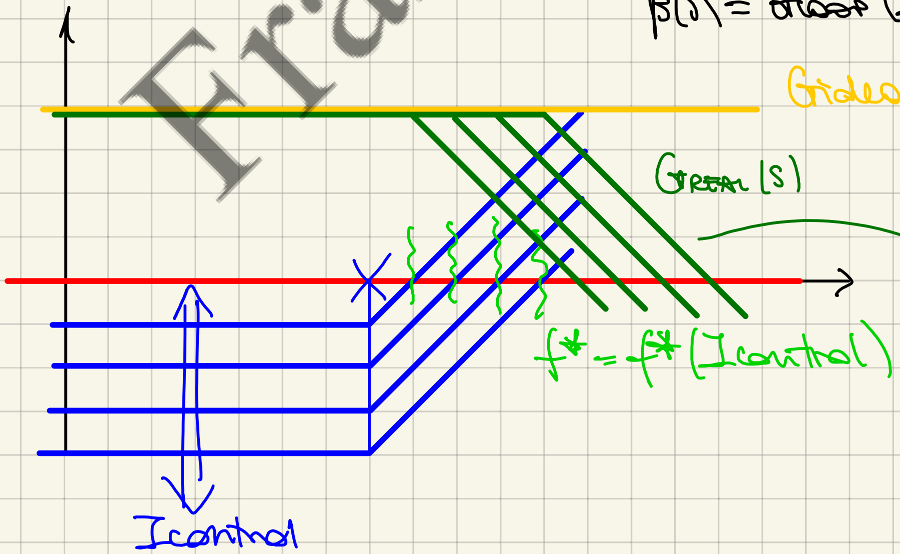
$G_{loop} = G_{loop}(I_{control})$

if G_{loop} is very high $\Rightarrow \varepsilon_v \to \emptyset \Rightarrow V_{OUT} \frac{R_1}{R_1+R_2} = V_{IN}$

$$\Rightarrow G = \frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_2}{R_1}$$

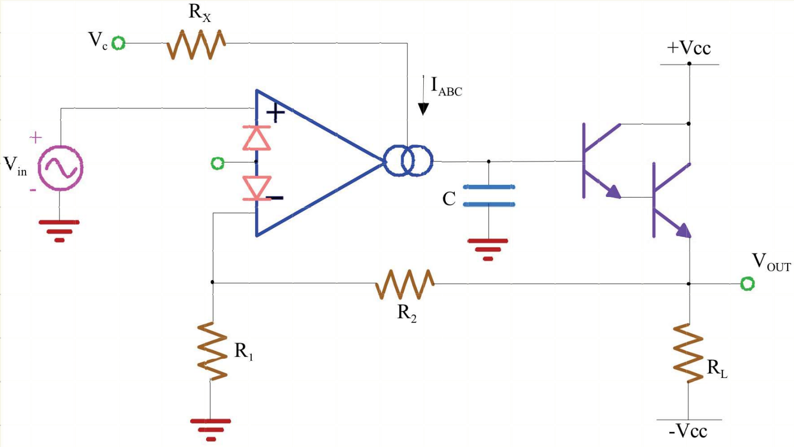
Let's study the Bode diagram:

$\beta(s) = G_{loop}(I_{control})$



The gain is fixed but it's pole changes as function of Icontrol

$pole_{REAL}(I_{control})$



$f_{POLE,REAL} = f_{POLE,R}(I_{control}) = f_{P,R}(V_c)$

since $I_{control} = I_{ABC} = V_c / R_X$

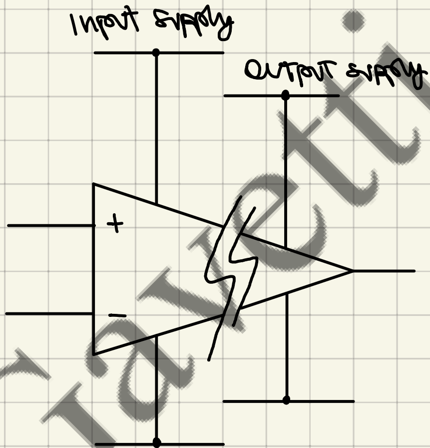
$\Downarrow$

That's why this configuration is called VOLTAGE-CONTROLLED LPF

ES08 - OTA and ISO AMPLIFIERS (2)

27/10/2021

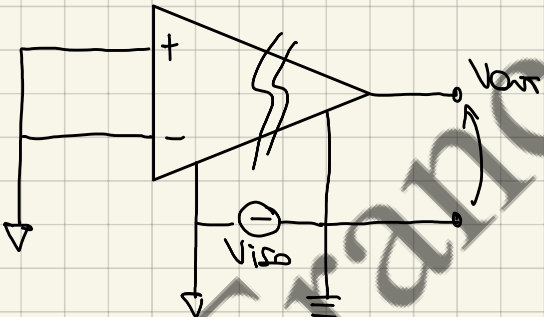
ISO - ISOLATION AMPLIFIER



REQUIREMENTS:
- isolated input and output stages
- double power supply
- galvanic isolation

In this way, if an issue happens @ the output port it doesn't have any impact on the input and vice versa

figure of merit → ISOLATION MODE REJECTION RATIO



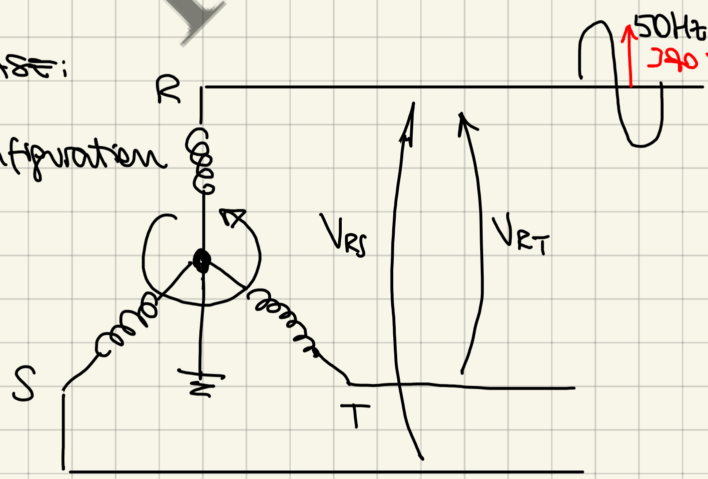
$$\boxed{IMRR = 20\log\left(\frac{V_{iso}}{V_{OUT}/G}\right)}$$

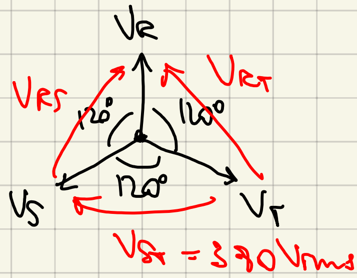
↓ it gives an idea on how much the isolation is able to cancel any possible voltage difference btw the two power supplies

EXAMPLE: ISOLATED POWERLINE MOTOR

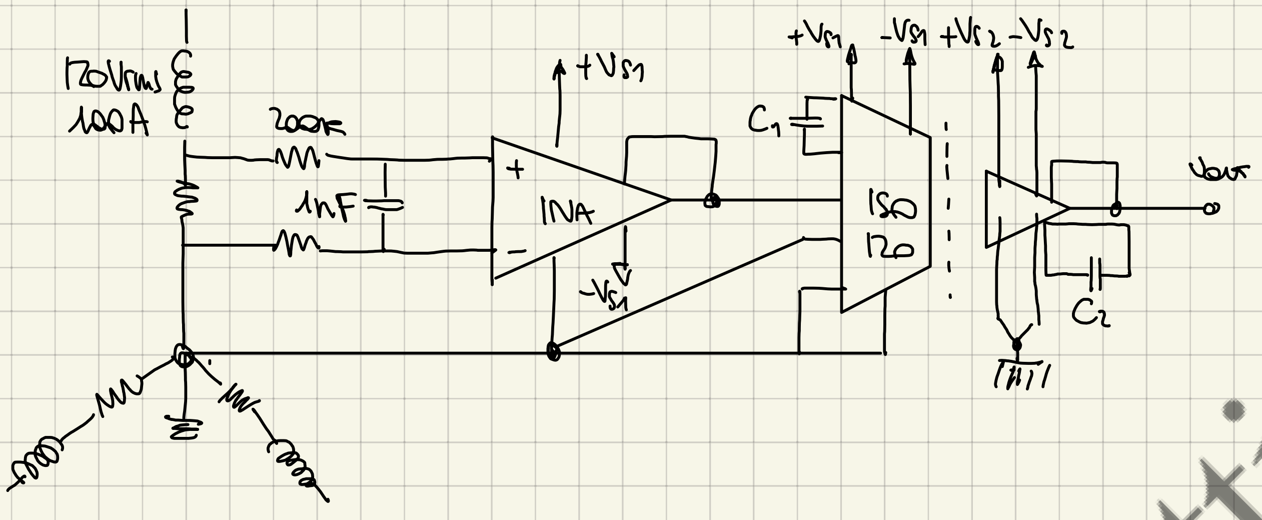
TRIPHASE:

"Y" configuration

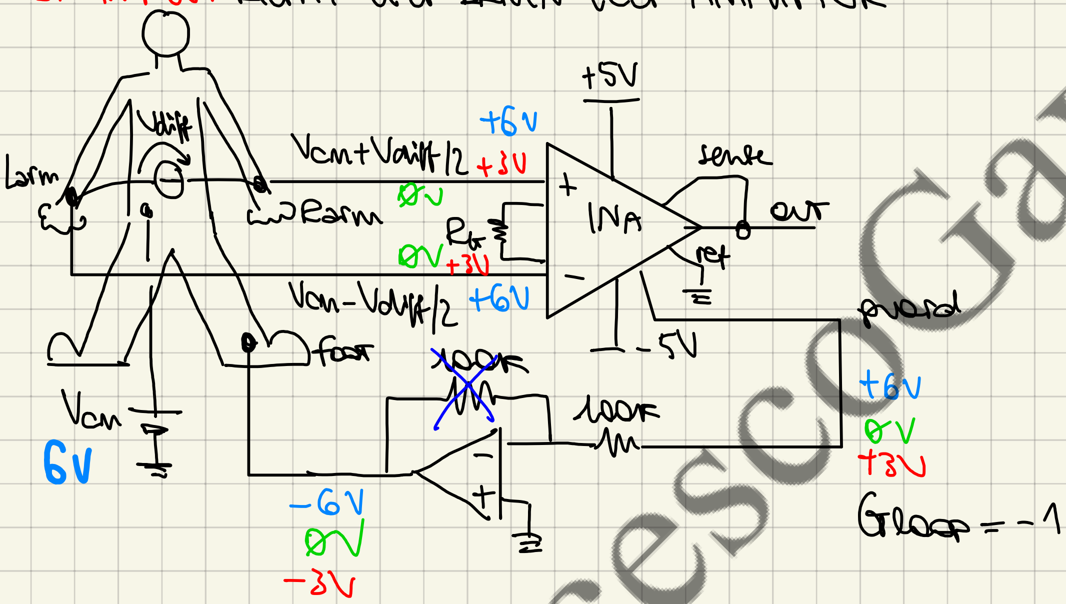




If we want to measure what's the current absorbtion we should measure what's the current flow through each winding



EXAMPLE: RIGHT-LEG DRIVEN ECG AMPLIFIER



Gloop = -1

In this circuit we've introduced negative FB to reject the common mode signal but Gloop is too poor

↓

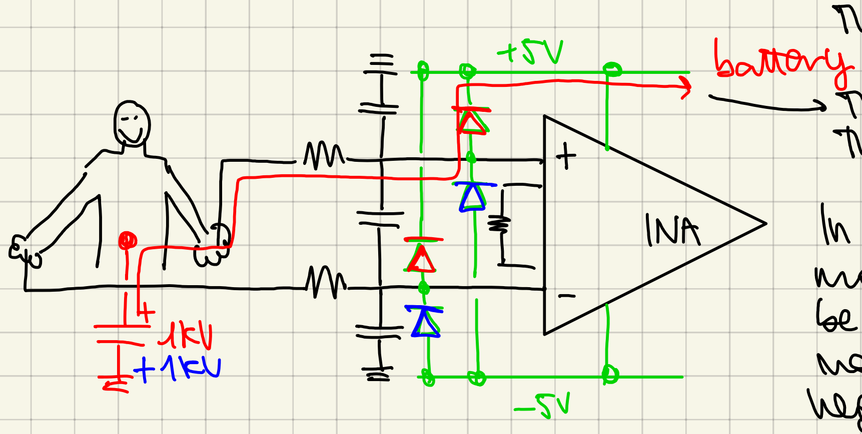
we have to increase it

↓

we can open the FB of the 2nd OpAmp * ⇒ Gloop = - Ao = - 10^5

↓

in this way we are able to counterbalance the CM of the patient body



→ This current will completely discharge the charge accumulated by the body

In any case the voltage that will be measured at the input of the INA will be very close to the PS (0.6V above the most positive PS, 0.6V below the most negative PS)

ISO: TECHNIQUES

We told that in principle IMRR should be infinite but is not.
In real circuits it's very high, up to 160 dB (10^8)

ISSUE OF THE ISO: The rejection is very strong JUST for DC values and it decreases a lot for increasing frequencies

$\rightarrow$ @ 1 MHz IMRR $\simeq$ 60 dB $= 10^3$

Also TRANSIENT IMMUNITY could be an issue $\rightarrow$ if the two power supplies change very quickly in a really short time, the output of the ISO could show some glitches, fluctuations and some spikes that could destroy the following electronics

TI < 1000 V/ms (iso122)

There are mainly 3 techniques to properly isolate the input from the output

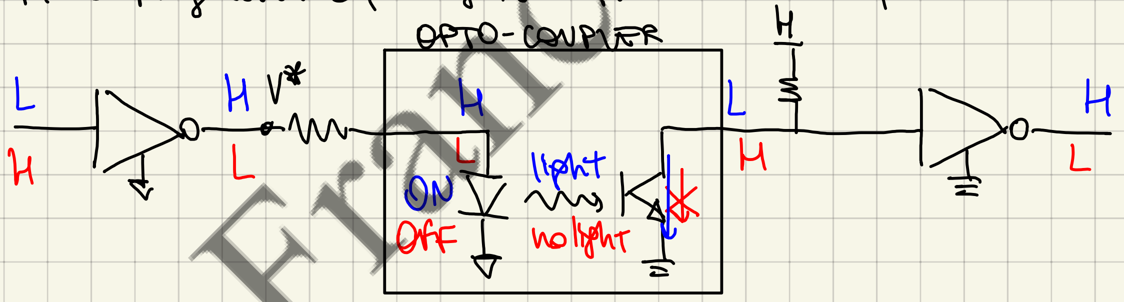
① OPTICAL COUPLING

An OPTO COUPLER is an IC that contains inside one LED and one transistor

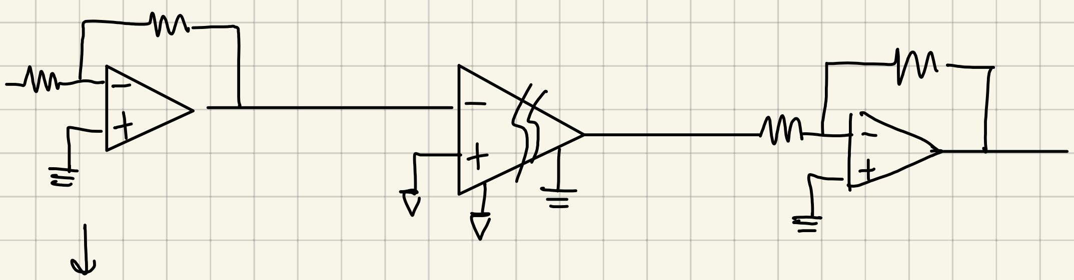
- if LED is OFF (it doesn't provide current)
  $\rightarrow$ the transistor has no I_B $\Rightarrow I_C = \beta I_B = 0$

- if LED is ON (the LED emits internally some light)
  $\rightarrow$ we have injection of carriers into the base of the transistor
  $\rightarrow I_B \neq 0 \rightarrow I_C \neq 0$

Opto coupling scheme for digital information transfer:



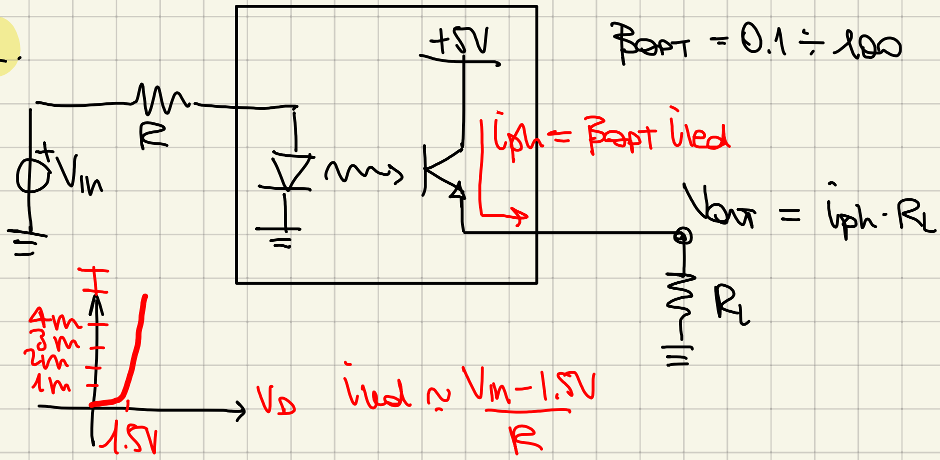
Optical coupling scheme for analog information transfer:



$\downarrow$

This is the final solution, but let's reach this final solution step by step

1.



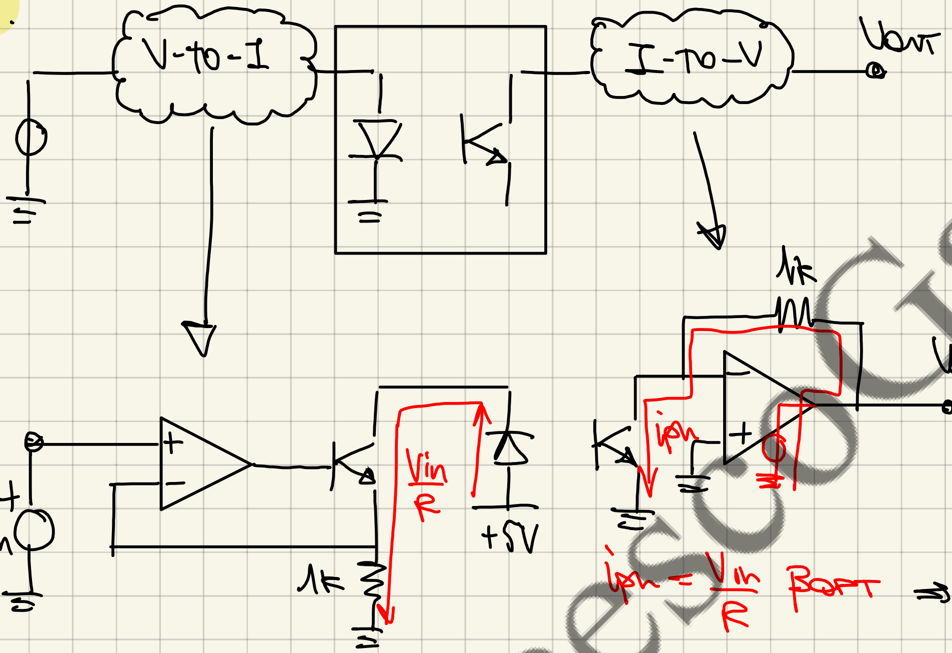
$\beta_{opt} = 0.1 \div 100 \Rightarrow$ it drifts/changes a lot (very high degree of mismatch)

$V_{out} = i_{ph} \cdot R_L$

$i_{led} \approx \frac{V_{in} - 1.5V}{R}$

Unfortunately this is not a good iso amplifier bc the relationship btw input and output is not linear

2.



$V_{out} = i_{ph} \cdot 1k\Omega$

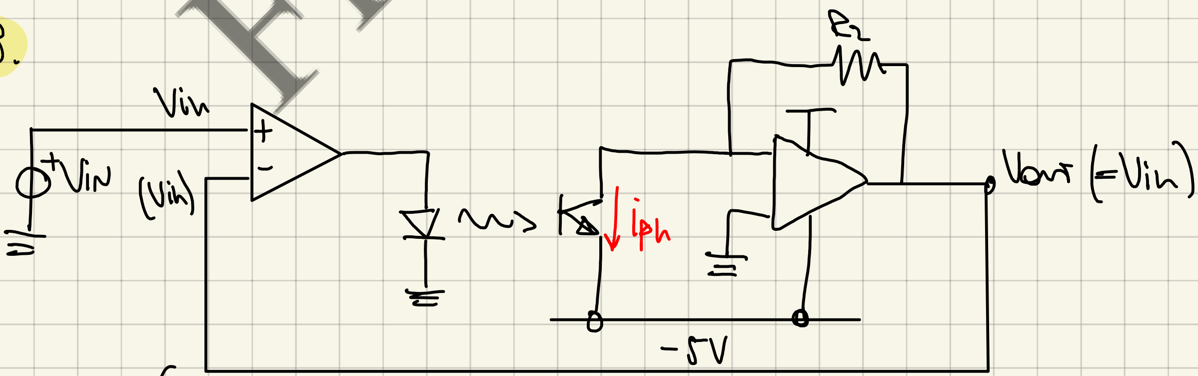
$i_{ph} = \frac{V_{in}}{R} \beta_{opt} \Rightarrow V_{out} = \frac{V_{in}}{1k} \beta_{opt}\, 1k$

Since β_{opt} drifts a lot, the opto coupler is not a good component, we can exploit its non ideality trying to compensate its non ideality

↓

let's use two opto-couplers which try to compensate each other

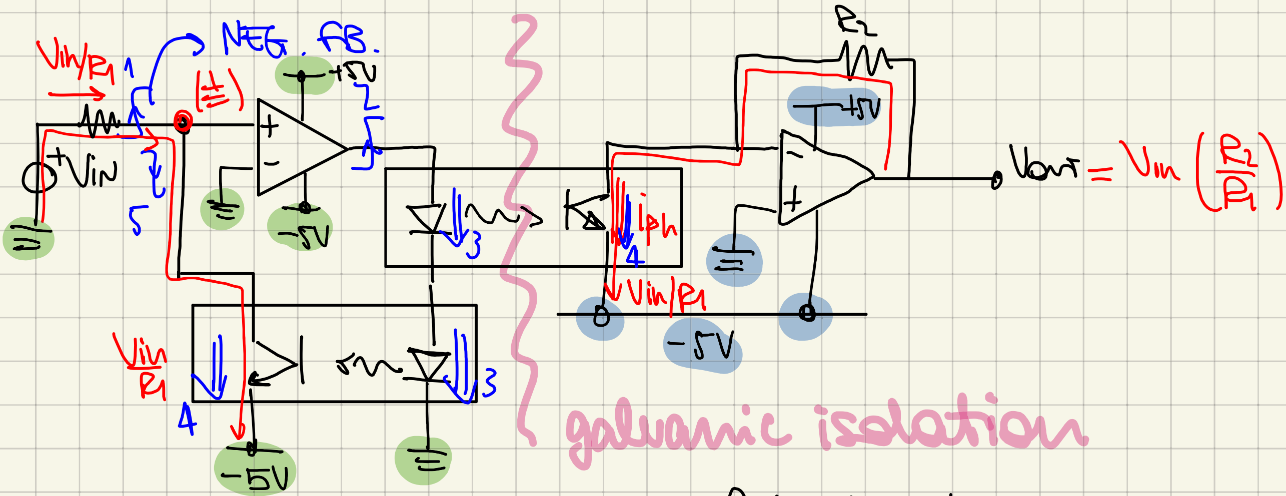
3.



↓

we don't want a wire that connects the input to the output, bc we want galvanic isolation

4.



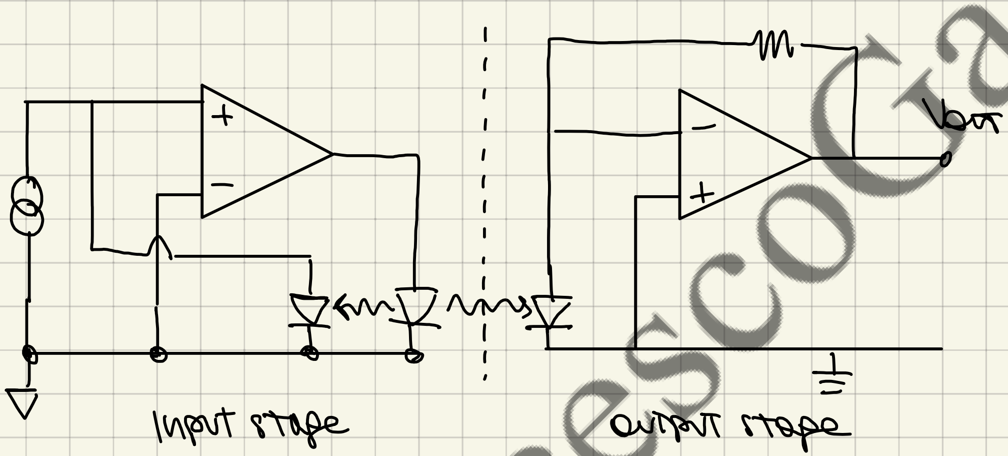
Let's place a 2nd opto-coupler identical to the 1st one

$G = R_2/R_1$ ☺

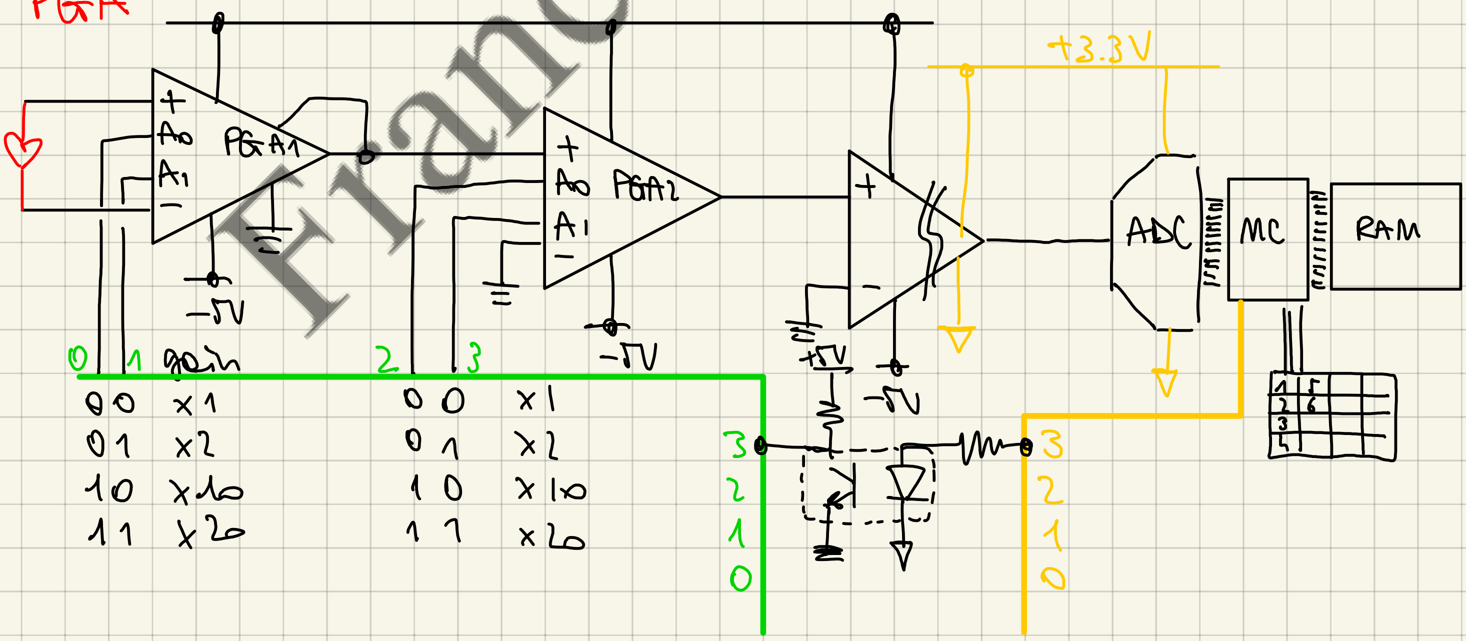
⊕

galvanic isolation ☺

EXAMPLE:

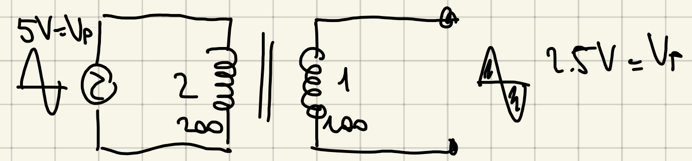


PGA

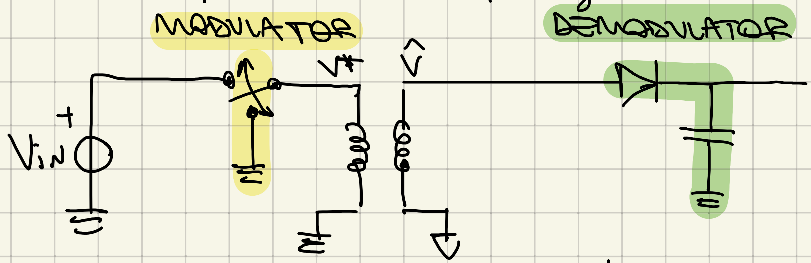


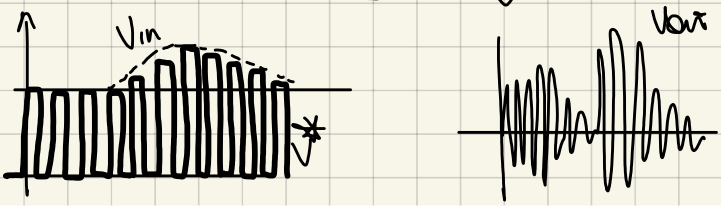
② MAGNETIC COUPLING

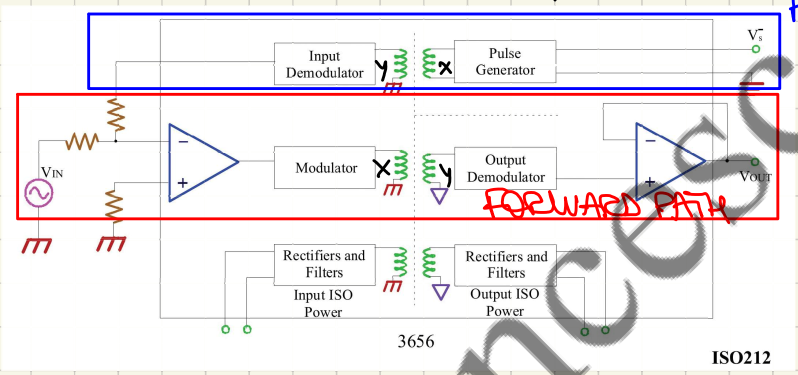
Transformer:



ISSUE: we cannot apply a DC signal
↓
if we want to apply a DC value we need a kind of MODULATION which transforms the DC value into a variable signal, then it will provide a variable OUTPUT which will finally be reconverted into a DC value







FEEDBACK PATH → it sense the output signal and brings it back to the input through another modulator and another demodulator
↓
the input demod of the FB path is identical to the output demod. of the forward path

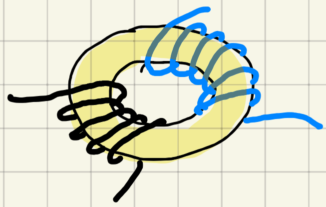
Even the transformers must be identical

If these conditions are verified the FB path and the forth one compensate each other and eventually any mismatch will be cancelled

Obviously we need two dedicated PSs: one for the input stage and another for the output one
↓
to spare components we can use another transformer How? IDK

How transformers are made?

A typical transformer could be based on a TOROID



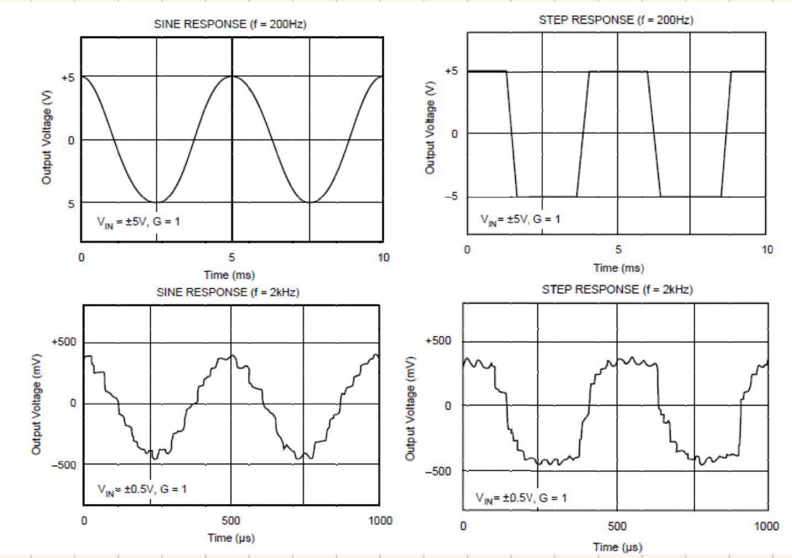
PROBLEM OF THE ISOLATION AMPLIFIER BASED ON MAGNETIC COUPLING

↓

we need modulators and demodulators

↓

This means that we cause a fluctuation inside the chip, a RIPPLE



In case of large signal it doesn't appear

But in case of small signals we have residual ripple at the output

↓

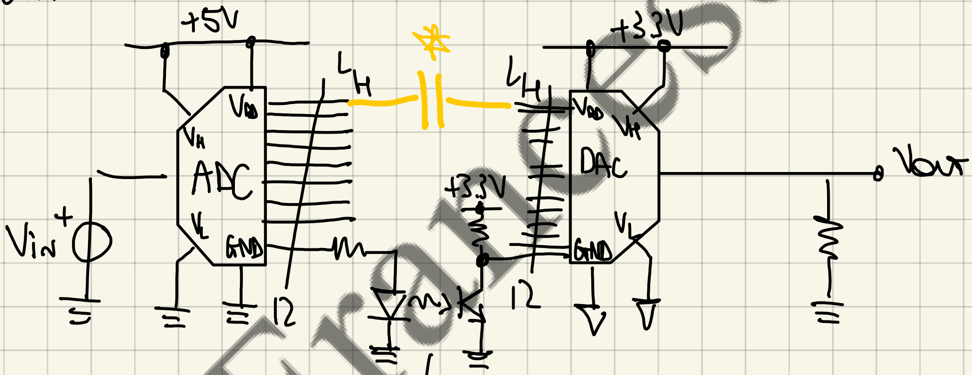
fluctuations in the zero crossing could badly impact the performance of the circuit

→ we should perform a kind of filtering at the output → LPF

③ CAPACITIVE COUPLING

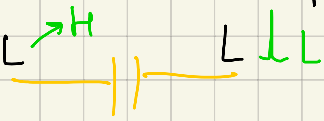
we can transfer information by charging and discharging a capacitor

To do so we can feed an ADC whose output is a digital bus then this digital bus can be connected to another DAC to bring back the signal in the analog domain



for each bit we have an opto-coupler like this

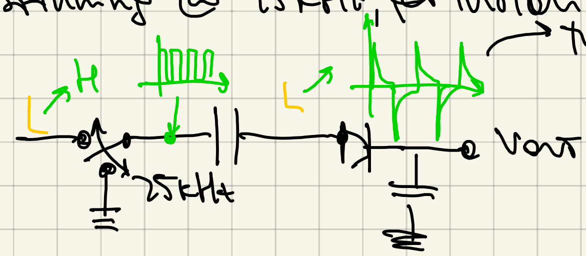
To connect the two pins we can also use a capacitor *

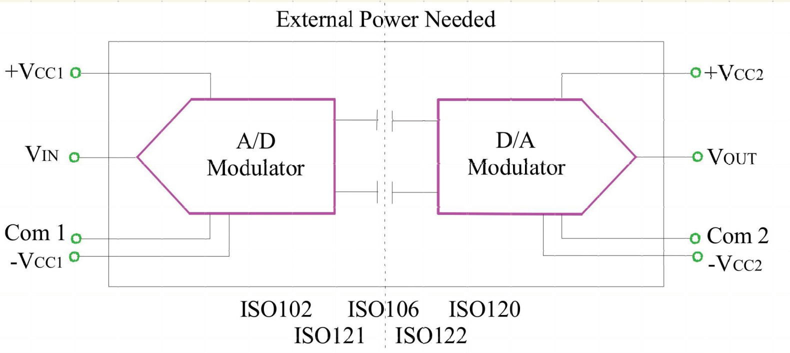


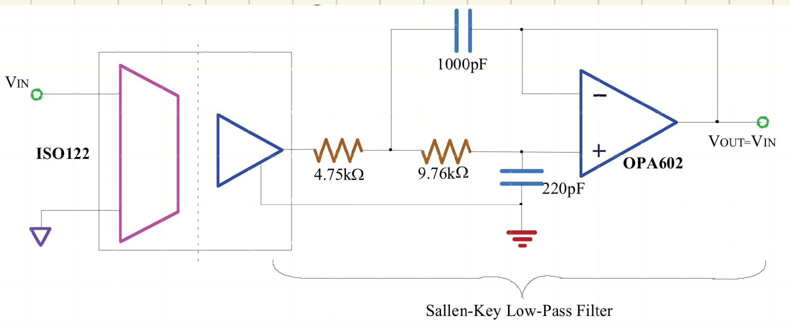
if we apply an high level in input we'll have a glitch but then the output will remain low bcs there's no DC information that can pass through the capacitor

→ To pass DC values we need a commutator at the input which keeps spinning @ 25kHz, for instance

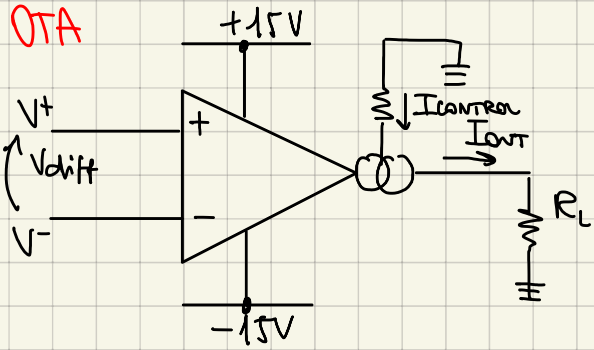
typical HPFing action of a capacitor







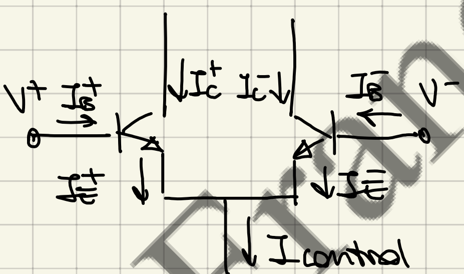
OTA



$$i_{out} = G_m(I_{control})\, V_{diff} = \frac{I_{control}}{KT/q} \cdot V_{diff} = \frac{I_{control}\, V_{diff}}{25mV}$$

The output current is linearly related to V_{diff} in this equation but we assumed the transconductance to be linear

But we have seen that the transconductance is not really linear



By varying V_{diff}, the current flowing through one branch, or the other, goes as the hyperbolic tangent of V_{diff}

$$I_{out} = I_C^+ - I_C^- = I_{control} \tanh\left(\frac{V_{diff}}{2KT/q}\right)$$

↓

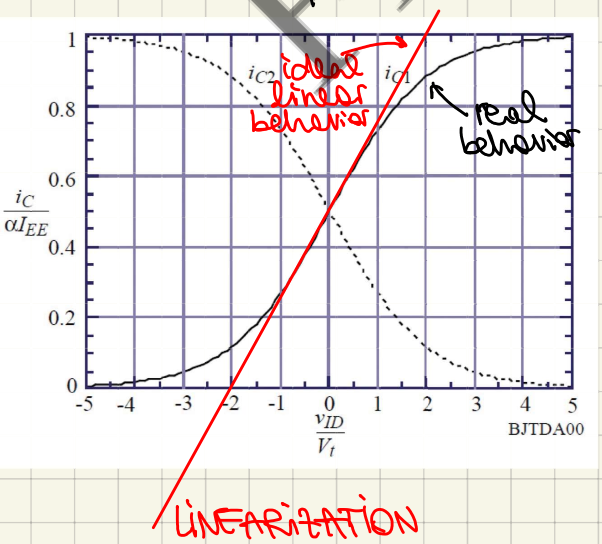
So it's not true we can apply whatever V_{diff} we want

↓

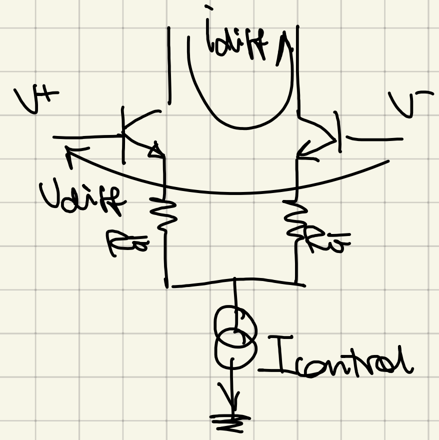
The OTA works linearly if V_{diff} is in the range of $V_{th} = KT/q$

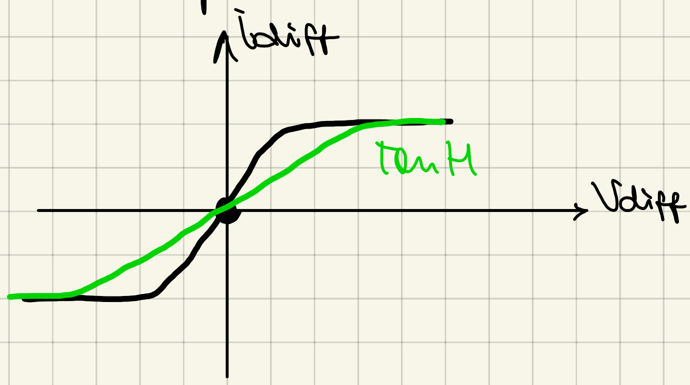
if $\boxed{|V_{diff}| \le V_{th}}$ ⇒ LINEAR → $-25mV \le V_{diff} \le +25mV$

if $V_{diff} \sim 2V_{th}$ → the linear behavior differs a lot from the real one



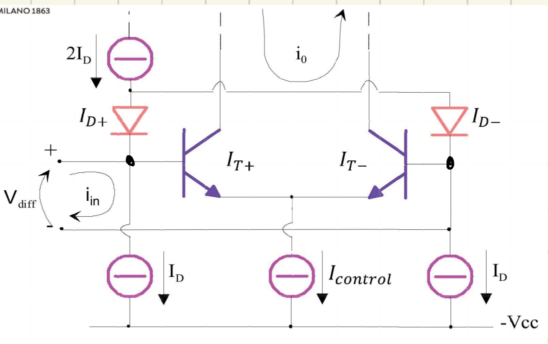
In order to improve the linearity we can add 2 resistors in series to the input transistors





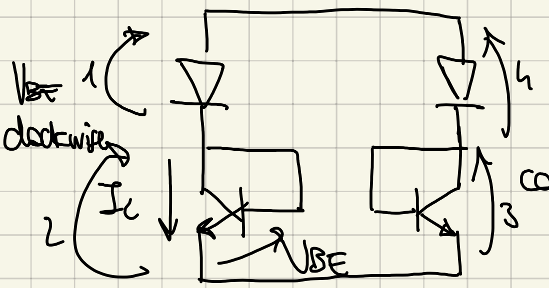
$$\frac{I_{control}}{2} \cdot R_E = 2V$$

Another better solution relies on a loop made by diodes



if the stage is balanced ($V_{diff}=0$) $\Rightarrow i_{in}=0$

if $V_{diff} \neq 0 \Rightarrow i_{in} \neq 0$



VBE counterclockwise

$$I_C = I_d \cdot e^{V_{BE}/kT/q}$$

$$\Rightarrow V_{BE} = \frac{kT}{q}\ln\left(\frac{I_C}{I_d}\right)$$

$$V_{BE1} + V_{BE2} = V_{BE3} + V_{BE4}$$

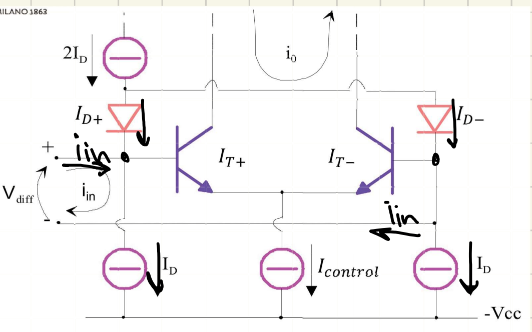
$$\cancel{\frac{kT}{q}}\ln\left(\frac{I_{C1}}{\cancel{I_d}}\right) + \cancel{\frac{kT}{q}}\ln\left(\frac{I_{C2}}{\cancel{I_d}}\right) = \cancel{\frac{kT}{q}}\ln\left(\frac{I_{C3}}{\cancel{I_d}}\right) + \cancel{\frac{kT}{q}}\left(\frac{I_{C4}}{\cancel{I_d}}\right)$$

$$\sum_{i\ \text{clockwise}} \ln(I_{Ci}) = \sum_{j\ \text{counterclockwise}} \ln(I_{Cj}) \Rightarrow \ln(I_{C1} I_{C2}) = \ln(I_{C3} \cdot I_{C4})$$

$$\Downarrow$$

TRANSLINEAR PRINCIPLE

$$\boxed{\prod_{i\ \text{clockwise}} I_{Ci} = \prod_{j\ \text{counterclockwise}} I_{Cj}}$$



$$\Rightarrow I_D^+ \cdot I_T^+ = I_T^- \cdot I_D^-$$

Then let's look at the nodes:

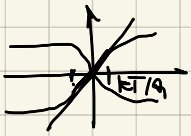
$I_D^+ = I_D - I_{IN}$ $\quad$ $I_D^- = I_D + I_{IN}$

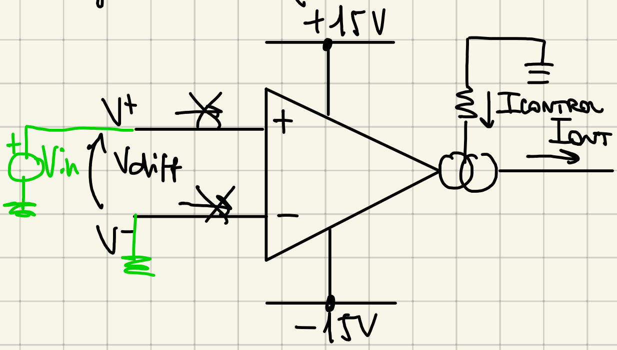
$I_{IN} = I_D - I_D^+ = -(I_D - I_D^-) \Rightarrow I_{IN} = I_D^- - I_D^+$

$I_{OUT} = I_T^+ - I_{CONTROL} = I_{control} - I_T^-$

$I_{OUT} = I_C^+ - I_C^- = \dfrac{I_{control}}{I_d} \cdot I_{in} \Rightarrow$ LINEAR!!!

Everytime we find an OTA like this:

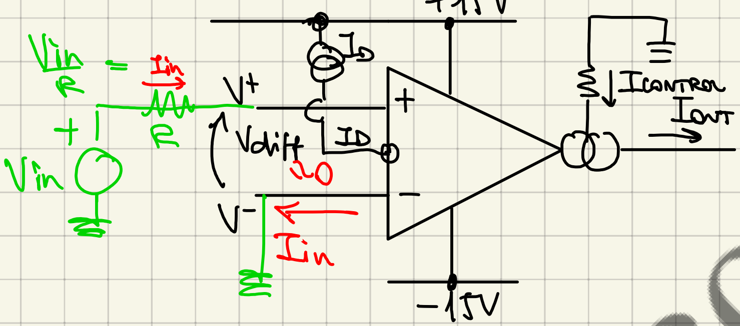




For $|V_{diff}| < V_{th} = kT/q$

$i_{out} = G_m(I_{control})\, V_{diff} = \dfrac{I_{control}}{kT/q} V_{diff}$

Instead, if we have an OTA like this:

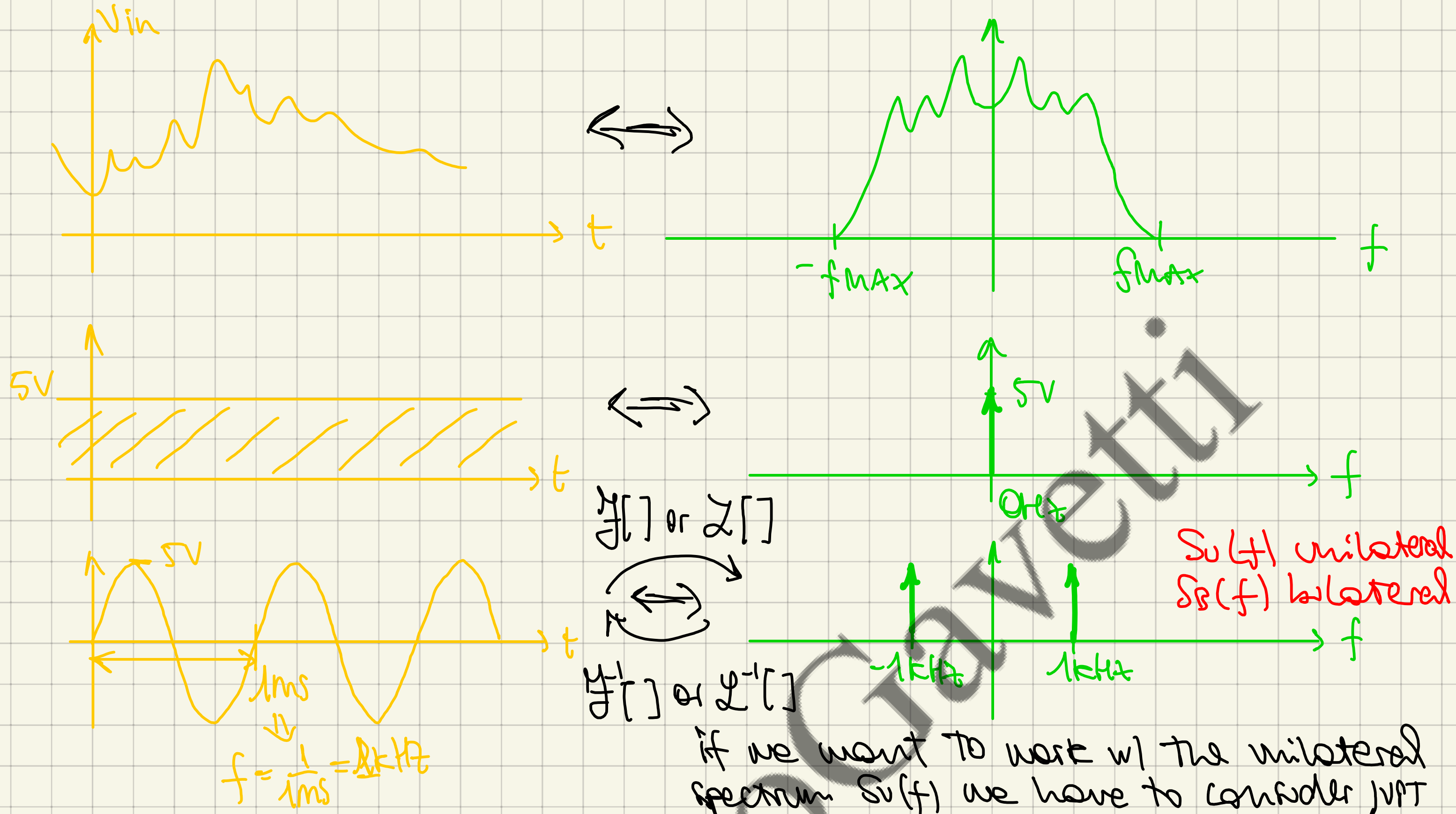


$i_{out} = I_{in} \dfrac{I_{control}}{I_{diode}}$

$= \dfrac{V_{in}}{R} \dfrac{I_{control}}{I_{diode}}$

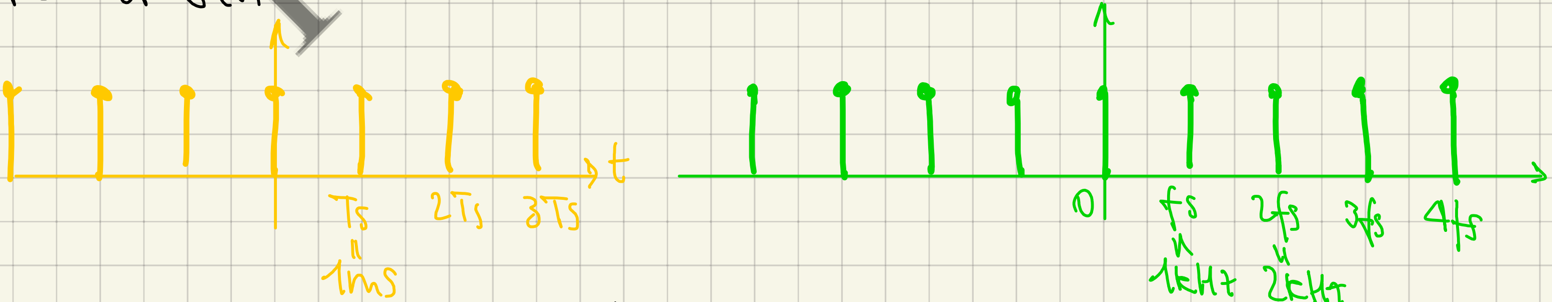
ES 09 - SAMPLING
02/11/2021

INTRODUCTION TO SAMPLING



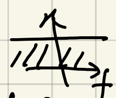
if we want to work w/ the unilateral spectrum $S_U(f)$ we have to consider just one delta @ 1kHz w/ 5V amplitude.
If we want to work w/ the bilateral spectrum instead, we have to consider two sinusoids respectively @ 1kHz and -1kHz w/ amplitude $5/2 = 2.5V$

train of $\delta(t)$



Notice : if $T_S = 1\mu s \Rightarrow f_S = 1MHz$

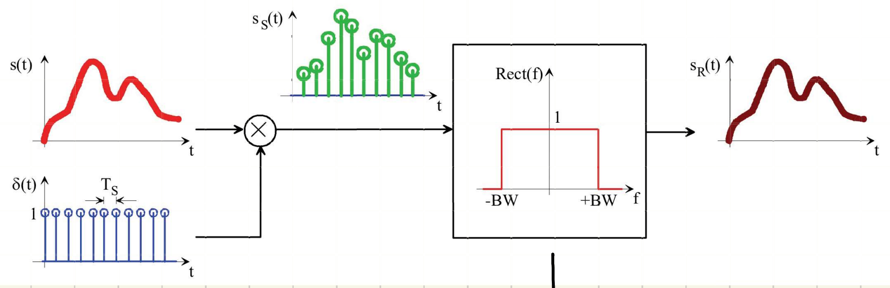
⇒ if in the time domain the deltas are very close to each other, they will be very spread in frequency (and vice versa)



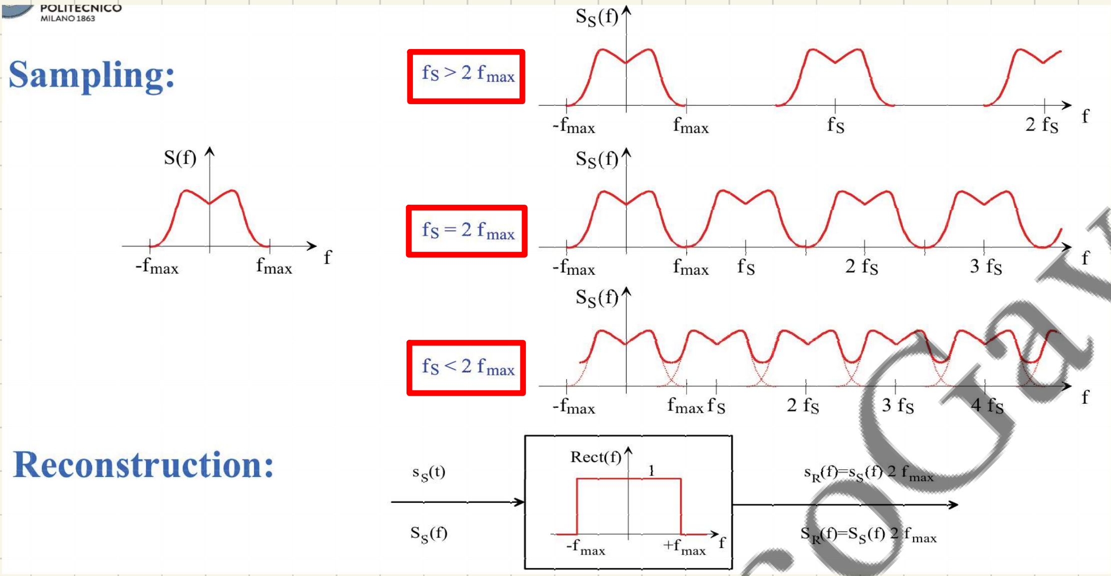
if $T_S \to \infty \Rightarrow f_S = \frac{1}{T_S} \to 0 \Rightarrow$ will have a constant distribution in freq.

SAMPLING THEORY

"if a ***function s(t)*** *has no frequency components above BW Hz, then it is fully* ***determined*** *by its* ***values****, sampled every* $T_S = {}^1/_{2BW}$"



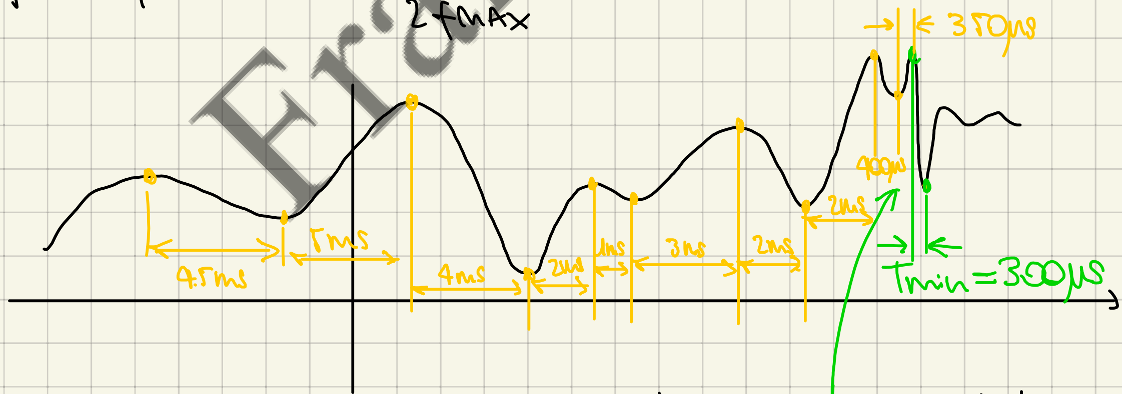
The filter will kill all replicas



$f_S > 2f_{max} \equiv T_S < \frac{1}{2f_{max}}$: OVERSAMPLING

$f_S = 2f_{max} \equiv T_S = \frac{1}{2f_{max}}$: SHANNON

$f_S < 2f_{max} \equiv T_S > \frac{1}{2f_{max}}$: UNDERSAMPLING → ISSUE : ALIASING



$T_{min} = 300\mu s \Rightarrow f_{max} = \frac{1}{T_{min}} = 3.3kHz$

we have to place at least 2 samples w/in T_{min} (so w/in f_{max})

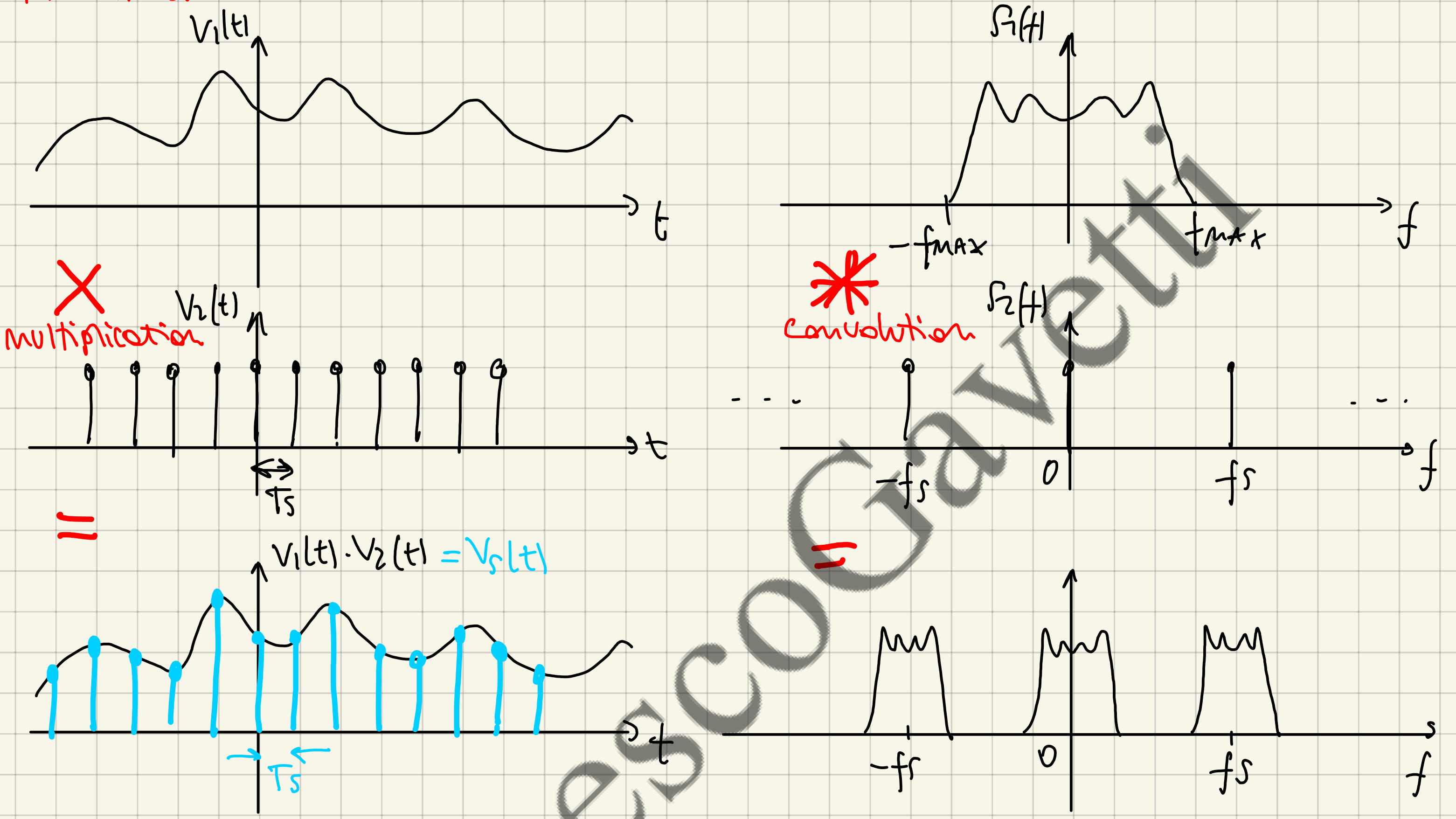
$\Rightarrow \boxed{T_S \le \frac{T_{min}}{2}}$ SAMPLING TIME

$\boxed{f_S \ge 2f_{max} = \frac{2}{T_{min}}}$ SAMPLING FREQUENCY if $f > \frac{f_s}{2}$ $(f_s < 2f)$ $\Rightarrow$ ALIASING

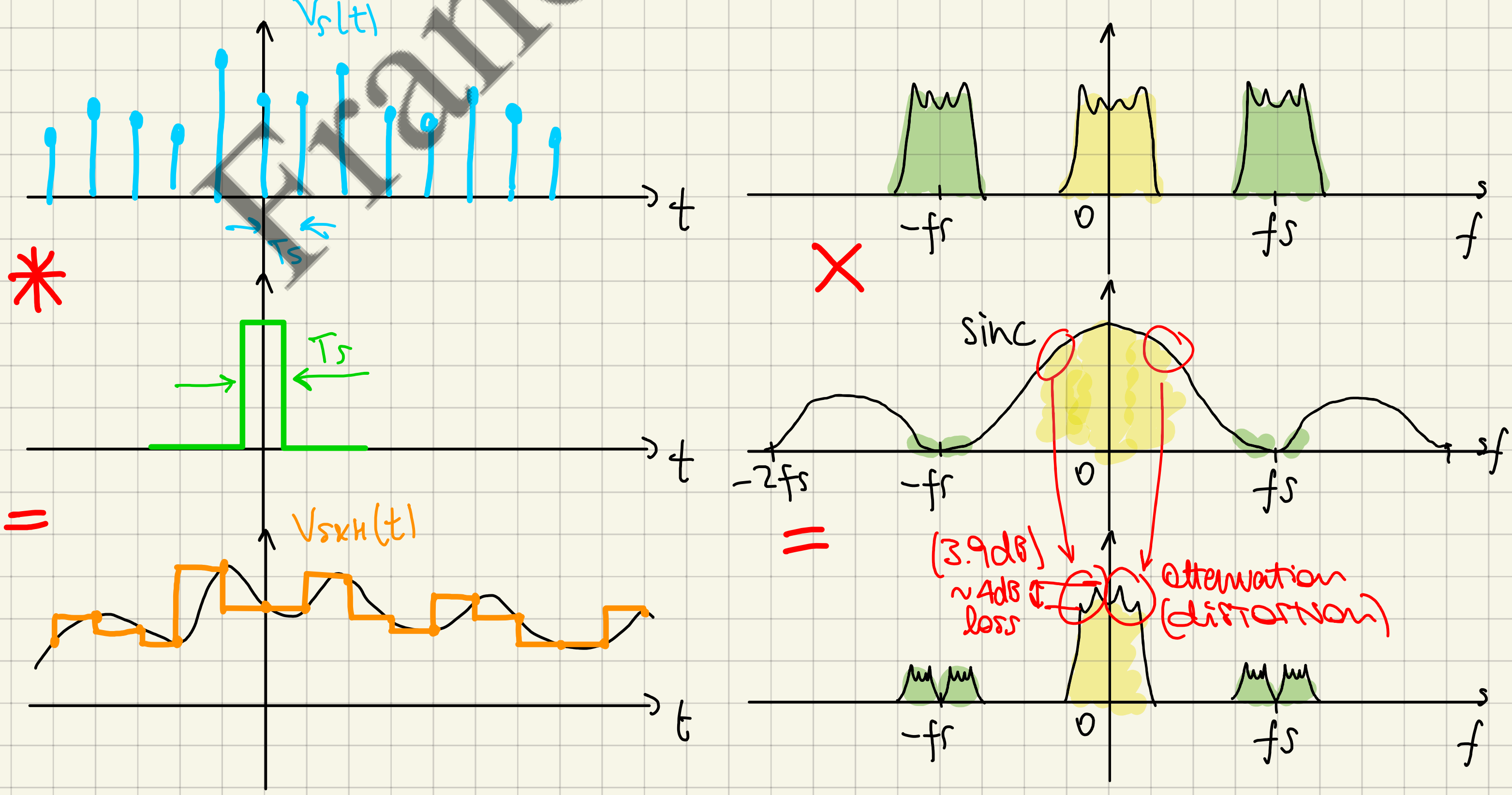
AVOID!!

Beware: There are also spurious frequency components and if $f_{spur} > f_s/2$ it will corrupt our signal

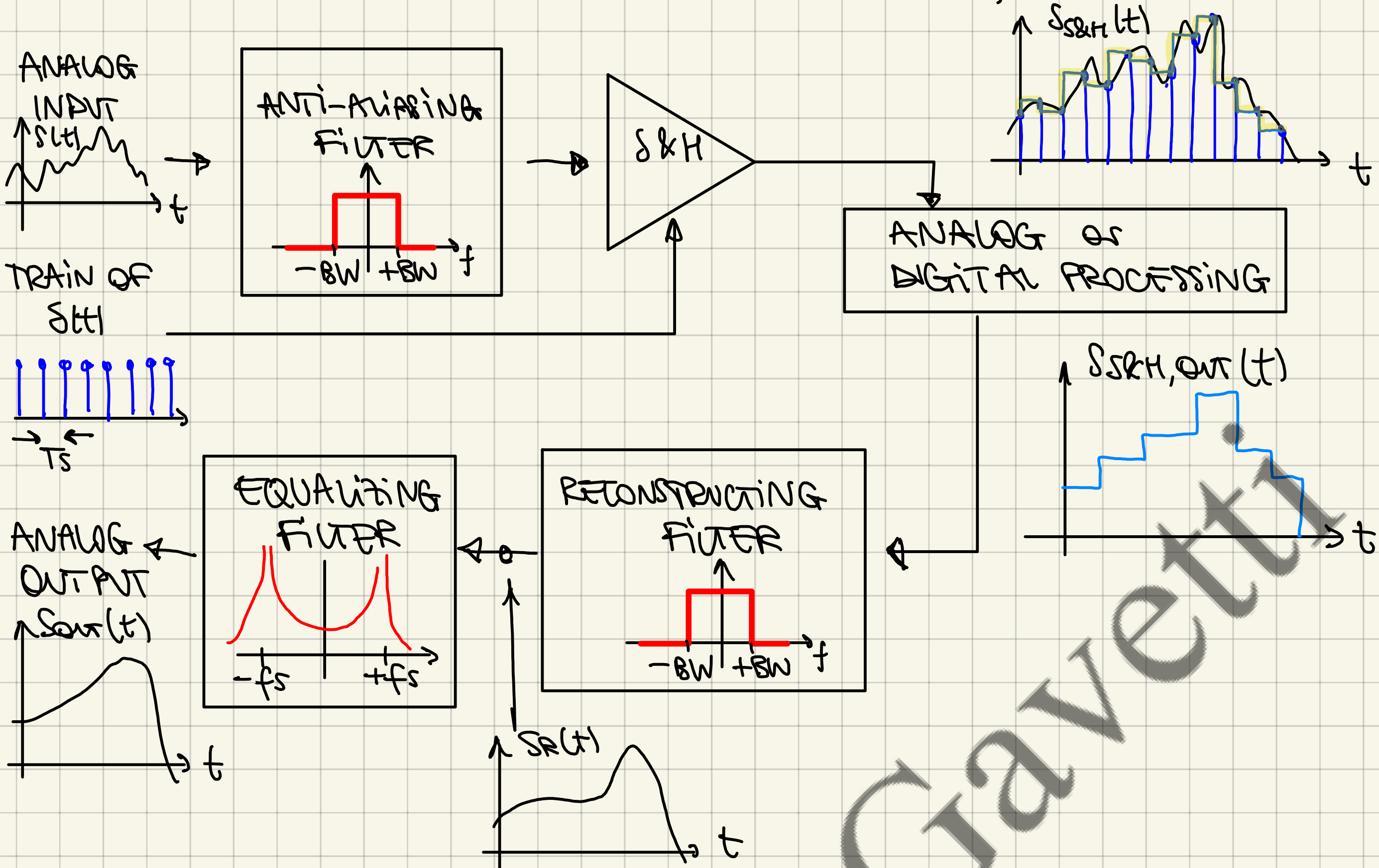
SAMPLING



S&H - SAMPLING AND HOLD

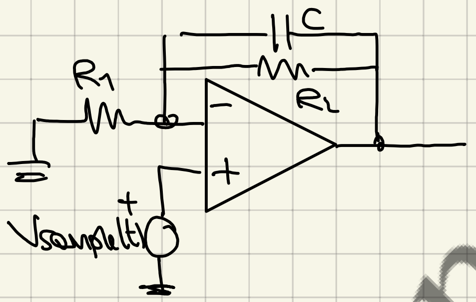


To recover the 4dB loss (3.9dB) (due to the sinc) we have to use an EQUALIZATION FILTER (to compensate the distortion)

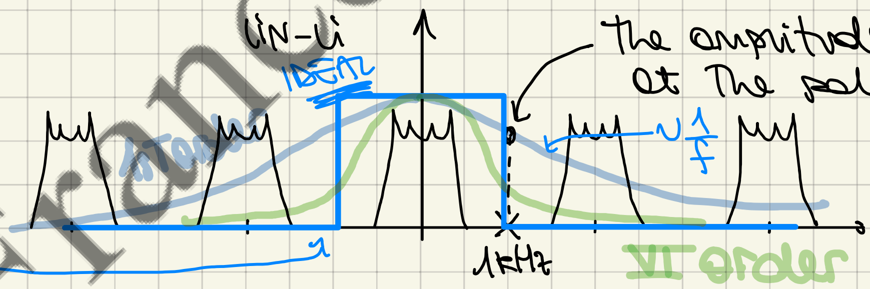


ANALOG FILTER

We have to kill all the replicas

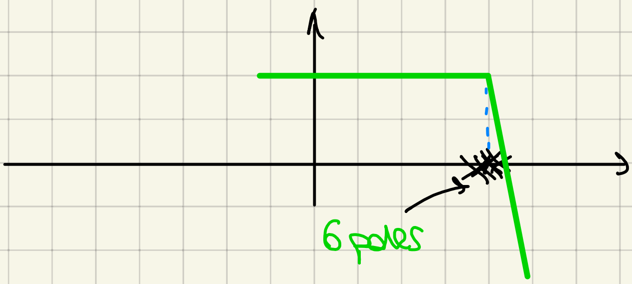


$$\text{pole} = \frac{1}{2\pi C R_2}$$



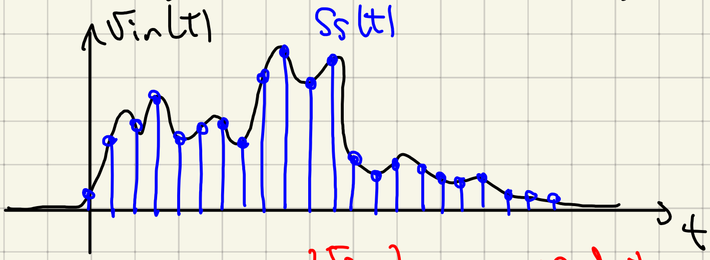
The amplitude of the filter at the pole freq (1kHz) is $1/\sqrt{2}$ compared to the max height

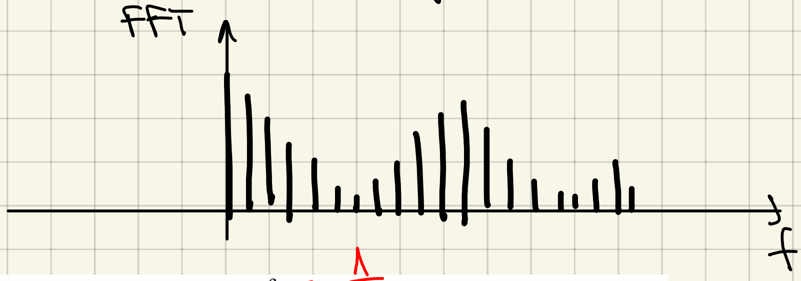
A LPF like that in the lin-lin plot is not a 1st order LPF!!!

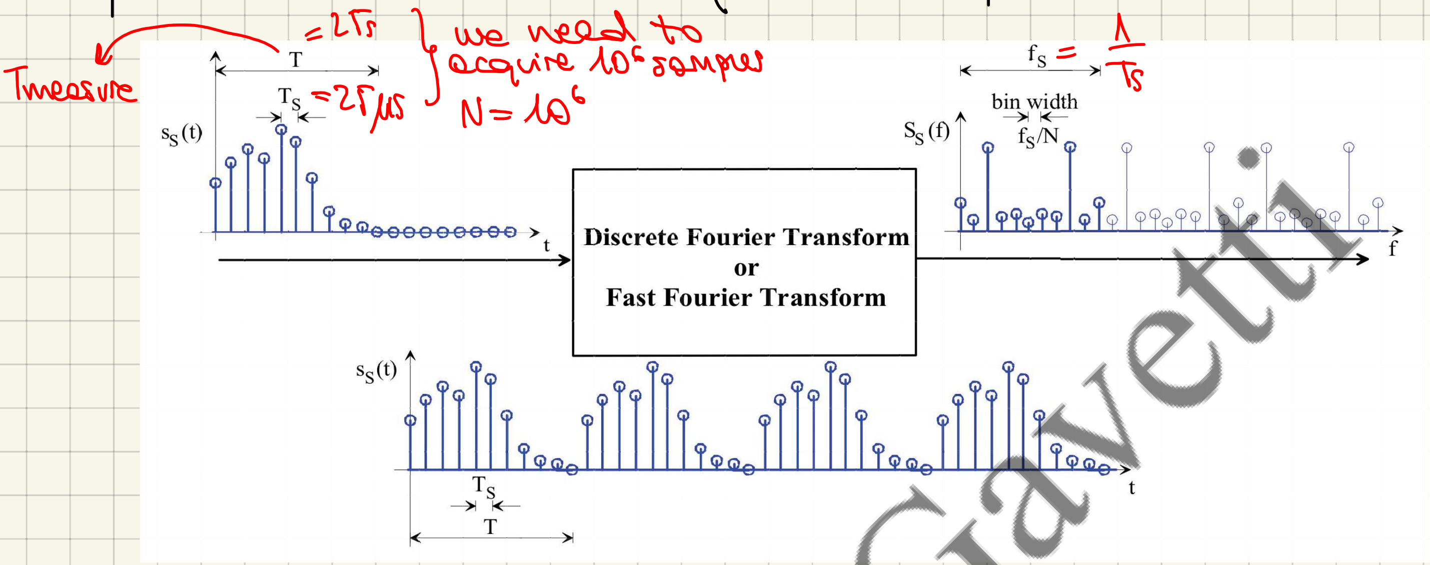


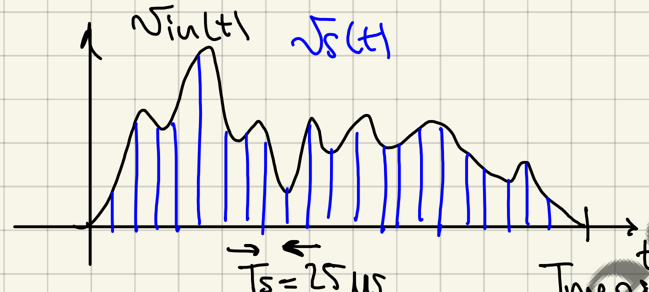
HOW TO COMPUTE THE SPECTRUM (FFT)

A FFT (FAST FOURIER TRANSFORM) is a mathematical algorithm where:



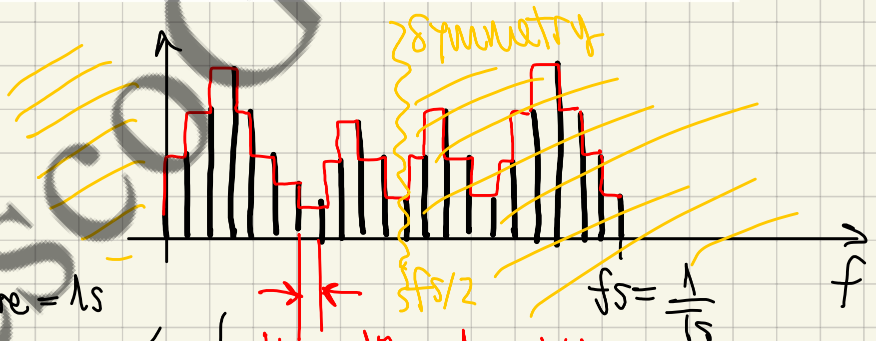






$T_s = 25\mu s$ $\qquad$ $T_{measure} = 1s$

$$N_{samples} = \frac{T_{measure}}{T_s} = \frac{1s}{25\mu s} = 40KSps$$



$$\text{bin-width} = \frac{1}{T_{measure}} = 1Hz$$
$$\text{bin-width} = f_s/N$$

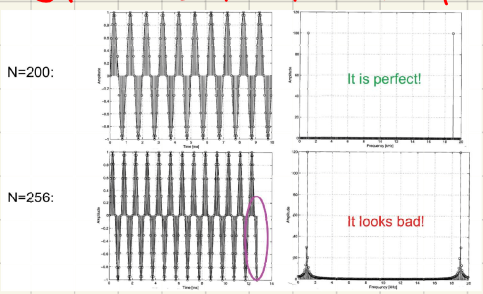
We have N samples also in the freq domain (btw 0 and f_s)

$$N = \frac{f_s}{\text{bin-width}} = \frac{1/T_s}{1/T_{meas}} = \frac{T_{meas}}{T_s} = 40KSps$$

Thanks to the symmetry we can stop to $f_s/2$!!

NOTICE: The FFT algorithm consider the input signal periodic, hence the FFT spectrum is for the periodic version of our input signal and not the original one!!

ERRORS IN THE FFT



W/ N=200 we have a perfect connection btw the beginning and the ending of the sequence while w/ N=256 samples (or a number different from an integer number of f_s/f_{IN}), the periodic sequence has an abrupt connection

Beware: The problem is that these truncations are not predictable when we have an unknown input signal

↓

"SOLUTION": to alleviate this problem, we can smooth the values of the sequence at the ending of the interval before applying the FFT algorithm

↓

WINDOWING BEFORE FFT { To do so we can use the WINDOWING TECHNIQUE according to which the samples' weight becomes less and less as we approach the extreme of the interval

↓

The windowing technique alleviates the problem of the connection and therefore improves the accuracy of the spectrum computed by the FFT

Types of windows:

- RECTANGULAR → not good, it doesn't smooth out external samples at all! ☹
(window(n) = 1)
- TRIANGULAR → $\text{window}(n) = 1 - \frac{|n|}{M}$ for $-M \leq n \leq M$ and $\emptyset$ elsewhere
- HAMMING → ...
- BLACKMAN → ....