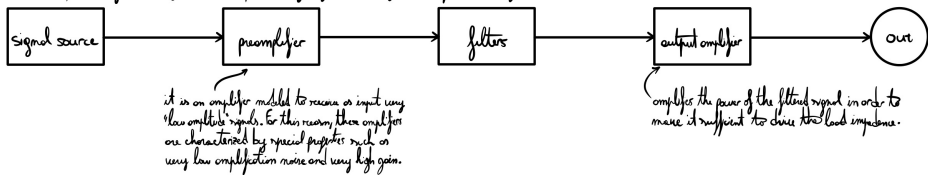


Audio Processing Chain

The audio processing chain of a audio-processing system is often composed as follows:



From now on, we will study one-by-one all the components of the audio processing chain.

Signal sources

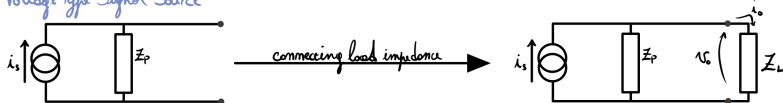
An audio signal source can be represented by one of the two following models, independently from their internal complexity.

Model 1: Voltage Type Signal Source



The ideal case is when $Z_s = 0$ and the voltage applied on the load impedance is independent from the load impedance itself (if we consider an ideal voltage generator). In reality, the condition $Z_s = 0$ is not possible, so we need to carefully choose Z_L .

Model 2: Voltage Type Signal Source



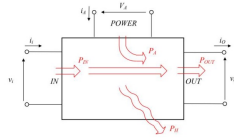
In this case the ideal case occurs when $Z_p = 0$ and all the generated current flows into Z_L . Also in this case, this configuration cannot exist in the real world.

Note: The two models are equivalent if $i_s = \frac{v_s}{Z_s}$ and $Z_s = Z_p$.

Macromodel of the Amplifier

The macromodel is an electrical model valid for a wide range of amplifiers. We study it before analyzing the specific types of amplifiers. Let's analyze the macromodel using various levels of abstraction.

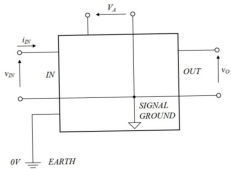
● **Macromodel - level of abstraction 0** → in this case we see the model as a black-box component having two input and two output pins, plus two pins for the power supply. It's possible to define the following quantities:



- **Power**
 - Input power → $P_i = i_i \cdot V_i$
 - Output power → $P_o = i_o \cdot V_o$
 - Power consumption → $P_h = i_{in} \cdot V_{in}$
- **Gain**
 - Voltage gain → $G_v = V_o/V_i$
 - Current gain → $G_i = i_o/i_i$
 - Power gain → $G_p = P_o/P_i = \frac{V_o \cdot i_o}{V_i \cdot i_i} = G_v \cdot G_i$
- **Efficiency** $\epsilon = \frac{P_{out}}{P_h}$ ← it's always < 1

for any amplifier, the power balance is $P_i + P_h = P_o + P_{dis}$. This should be > 1 .

● **Macromodel - level of abstraction 1** → let's consider the ground/earth potential. All the potentials in an amplifier are expressed in relation to the ground or Earth:

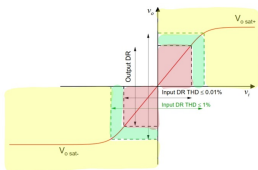


- **Earth:** it is the 0V potential, the potential of the Earth.
- **Ground:** it is an arbitrary potential used as reference point.

to be sure that no user can be electrically injured, it is important that every component and that can be touched is connected to Earth.

Sometimes the multiple connections of different modules to the Earth can cause a "ground loop" that generates disturbances in audio signal. To avoid that, all the modules are connected to a single cable and that cable goes to Earth.

• **Gain and its dependence from amplitude** → at this level of abstraction, we can talk about in/out characteristics. Ideally, when we feed an amplifier with a sinusoidal $A \cdot \sin(2\pi f t)$, the output we expect is $G \cdot A \cdot \sin(2\pi f t)$. Actually, it isn't always in this way. Real amplifiers have a non ideal input/output characteristic:



- **Linear region:** in this range the amplifier behaves as if was ideal, so $V_{out} = G \cdot V_{in}$. (THD $< 0,01\%$)
- **Non-linear region:** the still depends from V_{in} but not in a linear way. (THD $> 1\%$)
- **Saturation region:** that doesn't depend on V_{in} , but only on the saturation voltage.

Moreover, we are considering G in the linear zone not depending by other variables such as V_i , but in reality it is, and it is composed by several components as showed below:

$$G = G_0 + G_1 \cdot V_i + G_2 \cdot V_i^2 + \dots + G_m \cdot V_i^m$$

in which $G_0, \dots, G_m$ are constants and $G_{m+1} < G_m$.

● **Effect on the signal spectrum due to non-linearity** → the non-linear gain function has important effects on the amplified signal. Let's consider a simplified case in which $G = G_0 + G_2 \cdot V_i^2$. Two types of phenomena verify:

• **Harmonic Distortion** → let's see what happens if we feed an amplifier with a sinusoidal $V_i = A \cdot \sin(2\pi f t)$:

$$V_o = V_A \cdot G = V_A \cdot (G_0 + G_2 \cdot V_i^2) = G_0 \cdot V_A + G_2 \cdot V_A \cdot V_i^2 = G_0 \cdot A \cdot \sin(2\pi f t) + G_2 \cdot A^3 \cdot \sin^2(2\pi f t) = G_0 \cdot A \cdot \sin(2\pi f t) + \frac{G_2 \cdot A^3}{2} \cdot (1 - \cos(2\pi f t))$$

$$\text{Remember: } \sin^2(x) = \frac{1 - \cos(2x)}{2}$$

As we can notice, an additional component ($\frac{G_2 \cdot A^3}{2} \cdot (1 - \cos(2\pi f t))$) that wasn't in the original signal has been generated. If we consider a more complex gain function, even more additional harmonic components are generated. The effect of these components is the harmonic distortion.

• **Intermodulation Distortion** → let's now consider an input signal composed by two sinusoids $V_i = A_1 \cdot \sin(2\pi f_1 t) + A_2 \cdot \sin(2\pi f_2 t)$:

$$V_o = V_A \cdot G = V_A \cdot (G_0 + G_2 \cdot V_i^2) = G_0 \cdot V_A + G_2 \cdot V_A \cdot (A_1 \cdot \sin(2\pi f_1 t) + A_2 \cdot \sin(2\pi f_2 t))^2 = G_0 \cdot V_A + G_2 \cdot V_A \cdot (A_1^2 \cdot \sin^2(2\pi f_1 t) + A_2^2 \cdot \sin^2(2\pi f_2 t) + 2 \cdot A_1 \cdot A_2 \cdot \sin(2\pi f_1 t) \cdot \sin(2\pi f_2 t))$$

In this case, additionally to the harmonic distortion component, we are generating another component ($2 \cdot A_1 \cdot A_2 \cdot \sin(2\pi f_1 t) \cdot \sin(2\pi f_2 t)$) called "intermodulation distortion".

• **Measuring the distortion: Total Harmonic Distortion** → to measure the distortion of an amplifier in a simpler way it is often used the total harmonic distortion:

- Consider the distorted output $V_{out} = A_0 \cdot \sin(\omega t) + A_2 \cdot \sin(2\omega t) + \dots + A_m \cdot \sin(m\omega t)$.
- We know that in case of sinusoidal signals, the power can be calculated as $P = \frac{A^2}{2R_L}$, so in this case we have: $P_{out} = \frac{A_0^2 + A_2^2 + \dots + A_m^2}{2R_L}$.
- We want to evidence P_0 (power of the main component), so: $P_{out} = \frac{A_0^2}{2R_L} \cdot (1 + D_2^2 + \dots + D_m^2)$
- We rename the A_i/A_0 components: $P_{out} = P_0^2 \cdot (1 + D_2^2 + \dots + D_m^2)$

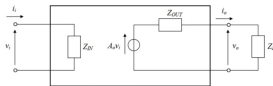
• The Total Harmonic Distortion is calculated as $THD = \sqrt{D_2^2 + \dots + D_n^2}$.

The input and output ranges of an amplifier can be defined based on the THD; for example we could require to be in "linear region" if $THD < 0.01\%$.

• **Gain and its dependence from frequency** → the gain function is also dependent from the frequency of the input signal: $V_{out} = G(j\omega) \cdot V_{in}$.
To understand that, we must pass to the next level of depend of the amplifier macromodel.

• **Macromodel - level of abstraction 2** → the internal components of an amplifier can be modeled as two components:

- an input impedance Z_{in} .
- an output voltage generator $A_{vo} \cdot V_{in}$ plus an output impedance Z_{out} .



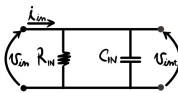
Let's focus on the first one of these components.

• **Input impedance of an amplifier** → the input impedance of an amplifier is determined by its input stage, which can be very complex, and it is defined as: $Z_{in} = \frac{V_{in}}{I_{in}}$.

• The input stage can be modeled in two ways:

• **DC input coupling**

This type of input stage is called "DC coupling" because a constant (DC) input voltage generates a constant (DC) input current, and the input voltage is amplified. Let's see how.



• if V_{in} is constant:

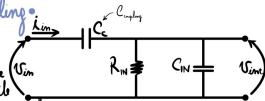
- R_{in} and C_{in} are not in parallel and they are subject to the same constant V_{in} .
- V_{in} is constant, so $f \rightarrow 0$, this means that the capacitor works like an open circuit.
- in that case, the input stage is equivalent to just Z_{in} and the input voltage is passed to the next stages to be amplified.

• if $V_{in} = A_v \cdot \sin(2\pi f t)$:

- the more $f \rightarrow \infty$, the more C_{in} behaves like a short circuit.
- because of this, the voltage that divides with f approaching ∞ .

• **AC input coupling**

This type of input stage is called AC coupling because it "stops" the direct inputs while let pass the alternated ones. Let's see how.



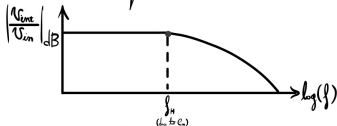
• if V_{in} is constant:

- V_{in} is applied to $C_{in} \parallel (R_{in} \parallel C_{in})$, and so a constant tension is applied to C_{in} .
- Being C_{in} a capacitor, it behaves as an open circuit when the tension applied to it is constant.
- If C_{in} behaves like an open circuit, no current flows in the resistor, and $V_{in} = 0$.

• if $V_{in} = A_v \cdot \sin(2\pi f t)$:

- in this case if $\omega f \ll \infty$ the tension V_{in} will be normally passed to the other stages of the amplifier.
- if $f \rightarrow \infty$, the capacitor C_{in} behaves like an open circuit $\Rightarrow V_{in} = 0$.

The result of the two cases explained is a behavior like this:



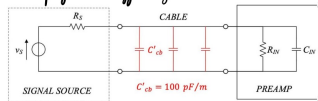
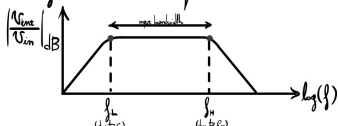
• Actually, f_L is also influenced by the cable that connects the signal source (instrument) to the amplifier. The effect of the cable is to add a capacitance in parallel to C_{in} .

$$\left(\frac{1}{R_{in}} + j\omega C_{in} + j\omega C_c \right)^{-1} = \left(\frac{1 + j\omega C_{in} R_{in} + j\omega C_c R_{in}}{R_{in}} \right)^{-1} = \frac{R_{in}}{1 + j\omega C_{in} R_{in} + j\omega C_c R_{in}}$$

$$V_{in} = V_{in} \cdot \left(\frac{C'_{in} \parallel C_{in} \parallel R_{in}}{(C'_{in} \parallel C_{in} \parallel R_{in}) + R_s} \right) \rightarrow \frac{V_{in}}{V_{in}} = G = \frac{R_{in}}{1 + j\omega C_{in} R_{in} + j\omega C_c R_{in}} \cdot \frac{1 + j\omega C_c R_{in}}{1 + j\omega C_{in} R_{in} + j\omega C_c R_{in}} = \frac{R_{in}}{R_{in} + R_s + j\omega C_{in} R_{in} R_s + j\omega C_c R_{in} R_s}$$

$$\omega = 2\pi f = -\frac{R_{in} + R_s}{j R_{in} R_s (C_{in} + C_c)} \cdot \frac{j}{j} \rightarrow f_H = \frac{1}{2\pi R_{in} R_s (C_{in} + C_c)}$$

The result of the two cases explained is a behavior like this:



in order to find f_H we have to find where the denominator is 0 (portion of the pole)

Operational Amplifiers

Opamps are a particular type of amplifiers, whose name originates from the fact that they were initially used to perform mathematical operations in analog computers.

● **Opamp transfer functions** → in the worst case (no feedback) the transfer function is:

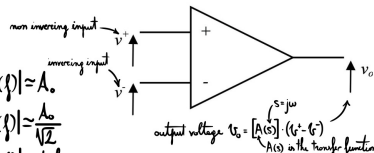
Laplace domain

$$A(s) = \frac{A_0}{1 + sT_0}$$

Frequency domain

$$A(f) = \frac{A_0}{1 + j2\pi f T_0} = \frac{A_0}{1 + j f/f_0}$$

$$\begin{cases} \text{if } f \gg f_0: |A(f)| \approx A_0 \\ \text{if } f = f_0: |A(f)| \approx \frac{A_0}{\sqrt{2}} \\ \text{if } f \ll f_0: |A(f)| \approx \frac{A_0}{f} \end{cases}$$



Typical parameters ranges:
GBWP ∈ [100, 1000]
A_0 ∈ [10^3, 10^7]
f_0 ∈ [10 Hz, 1 MHz]

At the frequency f_0 , the gain of the amplifier is reduced by a factor $\sqrt{2}$, which in dB corresponds to -3 dB.

● **Gain-Bandwidth-product** → A_0 and f_0 are two constants features of each specific opamp, so we can define a third constant $GBWP = A_0 \cdot f_0$.

Given that the gain A_0 is dimensionless and f_0 is a frequency, also the Gain-Bandwidth-product is a frequency. Normally, the GBWP is the frequency of which the gain modulus $|A(f)| = 1$.

The GBWP is an important parameter because it indicates the operating range of the amp:

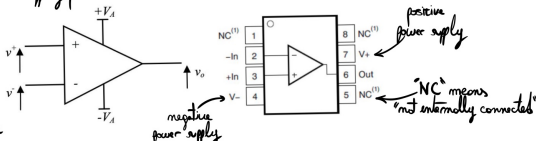
● **Opamp power supply** → Opamps have one or two power supply pins.

The fact that Opamps are subject to $\pm V_{cc}$ limits their working range. If we consider that $f_0 \approx V_{cc}$ is often in the order of mV.

the range of ΔV_{in} would be limited to few mV or even less, making the amplifier practically unusable.

for this reason, as we will see in the next paragraphs, Opamps are not used "as they are", but in "feedback configuration".

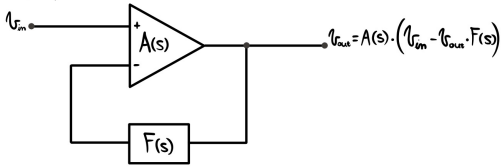
for $f < GBWP$ the gain $A(s) > 1 \Rightarrow$ the amplifier is amplifying.
for $f > GBWP$ the gain $A(s) < 1 \Rightarrow$ the amplifier doesn't amplify.



Feedback OpAmp

Adding a negative feedback to the Opamp allows us to achieve:
- insensitiveness of the gain from the temperature.
- low distortion.

Let's see how this is possible.



The gain of the system is: $V_{out} = A(s) \cdot (V_{in} - V_{out} \cdot F(s)) \rightarrow V_{out} = A(s) \cdot V_{in} - A(s) \cdot V_{out} \cdot F(s) \rightarrow V_{out} (1 + A(s) \cdot F(s)) = A(s) \cdot V_{in}$

$$\frac{V_{out}}{V_{in}} = \frac{A(s)}{1 + A(s) \cdot F(s)} = G_{LF}(s)$$

Nota:
A = Gain dell'Opamp reale (dimensionless)
F = funzione di retroazione (dimensionless)

● **Sensitivity to gain variation** → the gain of an opamp changes over time because of aging or temperature variation. Let's see how this phenomenon impacts on "normal" and on feedback opamps.

No feedback

We know that in this case the transfer function is $V_{out} = A \cdot V_{in}$. When the temperature changes, A changes; for this reason we have the following output variation: $\Delta V_{out} = \Delta A \cdot V_{in}$. We can also calculate the relative variation $\frac{\Delta V_{out}}{V_{out}}$ using the two previous equations:

$$\begin{cases} V_{out} = A \cdot V_{in} \rightarrow V_{in} = \frac{V_{out}}{A} \\ \Delta V_{out} = \Delta A \cdot V_{in} \rightarrow \Delta V_{out} = \Delta A \cdot \frac{V_{out}}{A} \rightarrow \frac{\Delta V_{out}}{V_{out}} = \frac{\Delta A}{A} \end{cases}$$

We can then define the sensitivity as $S = \frac{dV_{out}/V_{out}}{dA/A}$ and in this case we have that $S = 1$.

Feedback opamp

$$G = \frac{A(s)}{1 + A(s) \cdot F(s)}$$

In this case the transfer function is $V_{out} = G \cdot V_{in}$, so the variation due to environmental conditions is $\Delta V_{out} = \Delta G \cdot V_{in}$. From these informations we can derive the sensitivity:

$$\Delta V_{out} = \Delta G \cdot V_{in} \xrightarrow{\text{multiply and divide by } A} \Delta V_{out} = \Delta G \cdot \frac{A}{A} \cdot V_{in} \rightarrow \Delta V_{out} = \frac{\Delta G}{A} \cdot A \cdot V_{in} \rightarrow \frac{\Delta V_{out}}{V_{out}} = \frac{\Delta G}{G} \cdot \frac{A}{A} \cdot \frac{V_{in}}{V_{in}} \rightarrow \frac{\Delta V_{out}}{V_{out}} = \frac{\Delta G}{G}$$

$$\Delta G = \frac{dG}{dA} \cdot \Delta A \rightarrow \Delta V_{out} = \frac{dG}{dA} \cdot \Delta A \cdot V_{in} \rightarrow \frac{\Delta V_{out}}{V_{out}} = \frac{dG}{G} \cdot \frac{\Delta A}{A} \cdot \frac{V_{in}}{V_{in}} \rightarrow \frac{\Delta V_{out}}{V_{out}} = \frac{dG}{G} \cdot \frac{\Delta A}{A}$$

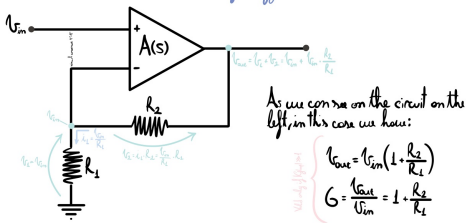
$$S = \frac{dV_{out}/V_{out}}{dA/A} = \frac{dG}{G} \cdot \frac{A}{1 + A \cdot F} \cdot \frac{1}{V_{in}} \rightarrow S = \frac{dA/A}{1 + A \cdot F} \cdot \frac{A}{1 - G_{LF}}$$

This means that we can decrease the sensitivity S increasing the G_{LF} .

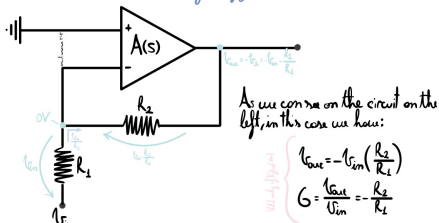
OpAmp-based feedback preamplifiers

Preamplifiers always use negative feedback because of the low distortion rate that this approach allows. There are two main configurations:

Non inverting configuration



Inverting configuration



Temperature dependence of the voltage gain → it is desirable for a preamplifier to have gain independent from temperature. We have seen that the gain in the two configurations is $G_{non-inverting} = 1 + \frac{R_2}{R_1}$ and $G_{inverting} = -\frac{R_2}{R_1}$, so we have to study the behaviour of R_1 and R_2 . In general the temperature dependence of a resistor can be expressed as:

$$R(T) = R(T_0) \cdot (1 + \alpha_r \cdot (T - T_0)) \quad \text{where } T_0 \text{ is the initial temperature, } T \text{ is the current temperature and } \alpha_r \text{ is a Temperature Coefficient.}$$

So, the $\frac{R_2}{R_1}$ temperature dependence is: $\frac{R_2(T)}{R_1(T)} = \frac{R_2(T_0) \cdot (1 + \alpha_r \cdot (T - T_0))}{R_1(T_0) \cdot (1 + \alpha_r \cdot (T - T_0))}$. This means that the gain is independent from temperature variations.

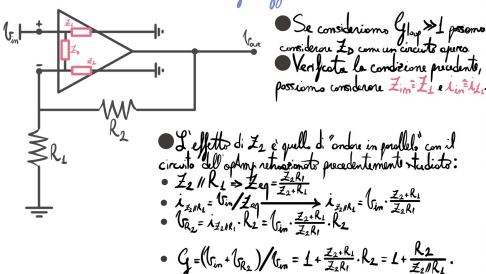
OpAmp input impedances → internally, a OpAmp has three impedances:

- impedance Z_D between the + and - inputs.
- impedance Z_1 between the + input and the ground $\pm$.
- impedance Z_2 between the - input and the ground $\pm$.

We can easily calculate that:

- $i_{Z_1} = (V^+ - V^-) / Z_1$ meaning that $i_{Z_1} = i_{Z_1} + i_{Z_2}$. However, we must note that the negative feedback generates a virtual short circuit between the + and - input pins, meaning that if $G_{loop} \gg 1 \Rightarrow i_{Z_1} = 0$ and it's like if Z_D doesn't exist. Let's see how this property influences the two opAmp configurations:

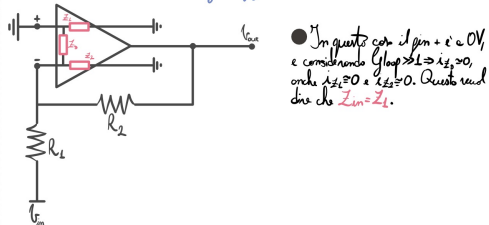
Non inverting configuration



Usually Z_2 is a capacitor, making the gain dependent from the frequency of the input.

Bandwidth of feedback opAmp → ideally ($G_{loop} = \infty$) the gain of a feedback opAmp does not depend from the signal frequency. Actually, G_{loop} cannot be infinite and the gain is frequency dependent. In particular, G_{loop} becomes smaller if frequency increases.

Inverting configuration

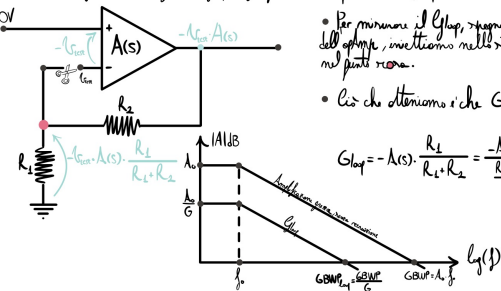


• Per misurare il G_{loop} , rimpiazziamo tutti i generatori impressivi "ideali" con la connessione di pin negativi dell'opAmp, mettiamo nello stesso pin negativo una tensione di test e misuriamo la tensione risultante nel punto rosso.

• ciò che otteniamo è che $G_{loop} = -A(s) \cdot \frac{R_1}{R_1 + R_2}$, che possiamo manipolare nel seguente modo:

$$G_{loop} = -A(s) \cdot \frac{R_1}{R_1 + R_2} = \frac{-A(s)}{\frac{R_1 + R_2}{R_1}} = \frac{-A(s)}{1 + \frac{R_2}{R_1}} = \frac{-A(s)}{1 - G_{1+2, feedback}} = \frac{-A(s)}{1 - G_{1+2, feedback}}$$

As we can see, graphically the G_{loop} is the $A(s)$ threshold crossed by a factor G_m



For what we now observe, we must not f_{max} (in order to cover the range 20Hz - 20kHz for audio application) using the following equation:

$$GBWP_{lap} = \frac{GBWP}{G_{M1}} \rightarrow f_{max} \cdot |G_{f(min)}| = \frac{GBWP}{G_{M1}} \rightarrow f_{max} = \frac{GBWP}{G_{M1} |G_{f(min)}|}$$

From that equation we can notice that, despite the fact that audio amplifiers has to cope with 20Hz < signals < 20kHz, they need to use opamp able to operate in the range of MHz in order to maintain an efficient negative feedback.

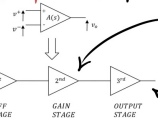
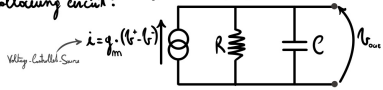
● **Slew Rate and Full Power Bandwidth of opAmps** → opamp are generally designed with two or more amplification stages each one has its own aim:

the first stage is usually a differential amplifier, whose amplify the difference between V_1 and V_2 , where:

- V_1 = non inverting input (ex. guitar signal).
- V_2 = inverting input (ex. ground or feedback).

Note that, amplifying the difference $V_1 - V_2$, we ignore the common component = noise.

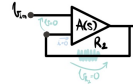
● The transfer function of such a opamp can be modeled using the following circuit:



the second stage add additional voltage gain.

the third stage is a unity voltage gain opamp, able to deliver an adequate output power

Unity Gain Amplifier (Buffer) → the non inverting configuration can be used to realize an opamp with $G_{M1} = 1$, setting $R_2 = \infty$.

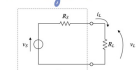


- We can notice that $G_{lap} = -A(s)$.
- $\eta_{lap} = 1$.

In this case we have no voltage amplification, but we do have **power amplification and current amplification**.

To understand that, let's move on example.

Source directly connected



In this case, the signal that reaches the load is: $V_L = V_S \cdot \frac{R_L}{R_L + R_i}$

Source connected through



• The opamp has (usually) $Z_{in} = \infty$, meaning that $Z_{in} = 0$, $V_{in} = 0$ and so the full signal V_{in} reaches the load Z_L .

from the frequency point of view, the t_{rise} time is frequency is limited by the capacitor C, that imposes a maximum voltage variation speed. It can be calculated in the following way:

$$i = C \cdot \frac{dV_L}{dt} \quad \text{is limited because it is not ideal generator} \rightarrow i_{max} = C \cdot \frac{dV_L}{dt}$$

$$\text{this is called slew rate.} \rightarrow \left. \frac{dV_L}{dt} \right|_{min} = \frac{i_{max}}{C} \geq \frac{dV_L}{dt}$$

The slew-rate represents the maximum t_{rise} velocity that the signal can have.

Then means that, when we are dealing with the amplification of a signal, we must be sure that every component of our original signal's variation is smaller than the slew rate. To do so, we have to:

1. Take in consideration the component with highest amplitude → $f = A_{max} \cdot \sin(2\pi f t)$.
2. Calculate its derivative over time → $f = A_{max} \cdot 2\pi f \cdot \cos(2\pi f t)$ $\xrightarrow{\text{maximum of cos}}$ $f'_{max} = A_{max} \cdot 2\pi f$
3. Check that $f'_{max} \leq \text{slew-rate}$.

● From the previous equation, we can also derive the maximum frequency at which the opamp can cope at maximum voltage:

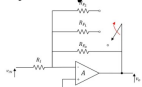
$$\overset{\text{maximum voltage}}{A_{max}} \cdot 2\pi f_{max} = \overset{\text{maximum frequency}}{\text{slew-rate}} \rightarrow f_{max} = \frac{\text{slew-rate}}{2\pi A_{max}} = \text{FPBW}$$

Full Power Bandwidth

Gain control

There are different ways to dynamically change the value of the gain of amplifiers. Let's see them.

● **Switch connected at the output of a differential opAmp** → a way to obtain a "discrete regulation" of the gain is to build a circuit like the following:



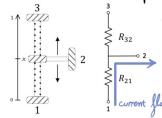
● We can notice that it is an inverting configuration. Ignoring the switch, the gain would be $G_v = -\frac{R_2}{R_1}$. In this case we also must consider the switch and the parallel resistances:

$G = \frac{R_2/R_1}{1}$ with $i = 1, 2, \dots$. The first resistance R_2 needs to be always connected in order to guarantee the feedback.

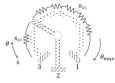
● In this type of circuits, the switch is connected to the output of the opAmp for two main reasons:

- the switch introduces **parasitic capacitance**, and
 - input of the opAmp → characterized by high impedance (ideally ∞). The parasitic capacitance would be in parallel with R_{in} and we know that a capacitor charges in a time that is proportional to a time constant $\tau = R \cdot C$, so if $R \rightarrow \infty$ also $\tau \rightarrow \infty$. This would introduce a very long delay in following the input signal, reducing the opAmp bandwidth.
 - output of the opAmp → it is characterized by an impedance ≈ 0 , so also $\tau = R \cdot C \approx 0$. For this reason, the switch is connected to the output of the opAmp.
- the switch is a physical component that can introduce **electromagnetic disturbances**. These electromagnetic disturbances are modelable as a current generator whose current flows into the input/output resistances, producing a $V_{disturb}$ component:
 - connection to input → Z_{in} is very high, so $V_{disturb} = I_{disturb} \cdot Z_{in}$ would be very big.
 - connection to output → Z_{out} is ≈ 0 , so $V_{disturb} = I_{disturb} \cdot Z_{out}$ would be near to 0.

● **Potentiometers** → potentiometers are variable resistors used to control continuously the gain.



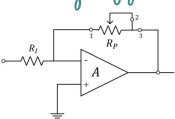
● the pin 2 is connected to a moving contact that allows to modify the values of R_{21} (and R_{22}). The value of R_{21} can go from 0 to R_p , where $R_p = R_{21} + R_{22}$. The resistor can also be curved.



Current flows in that direction, while R_{22} is usually when circuit.

● How to connect a potentiometer to an opAmp:

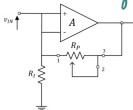
● Inverting configuration



● In practice we replaced the "classic" R_2 with R_{21} , so the gain will be:

$$G_v = -\frac{R_{21}}{R_1} \text{ with } 0 < G_v < -\frac{R_p}{R_1}$$

● Non inverting configuration



● Similarly to the inverting case, we replace R_2 with R_{21} . The gain is:

$$G_v = 1 + \frac{R_{21}}{R_1} \text{ with } 1 < G_v < 1 + \frac{R_p}{R_1}$$

● As we can notice, in both the configurations, the potentiometer pin is connected to the output of the opAmp. This is because the potentiometer is subject to parasitic capacitances, so, as we did for the switch, it's better to connect it to the output (which is characterized by a very low Z_{out} and so $\tau = R_{out} \cdot C$ will be very low).

● How to set the level of a preamplifier → we want to set the gain to guarantee a correct operation of a system. Usually, a pre-amplifier's output is connected to an ADC or to a power amplifier. Let's see how to set the pre-amplifier's gain in these cases.

● Connection to a Analog to Digital Converter

In order to use the entire dynamic of the ADC, it's necessary to properly set the gain of the pre-amplifier. The algorithm is the following:

① We must know the maximum input voltage range of the ADC.

ex. ADC with 16bit and 0V < V_{in} < 1V

② Evaluate the maximum range of the pre-amp input signal.

$V_{in}^{(preamp)} = 0.1V$

③ Set the pre-amp's gain so that $V_{in}^{(preamp)} = V_{in}^{(ADC)}$. It can also be smart to set a "security margin" to prevent peaks higher than expected. In that case: $V_{in}^{(preamp)} = V_{in}^{(ADC)} - V_{margin}$.

ex. $V_{in}^{(ADC)} = 1V$, $V_{margin} = 0.1V$ ⇒ $G_{preamp} = \frac{0.9V}{0.1V} = 9$

④ We can also calculate the minimum voltage step:

- Maximum voltage in input to the ADC = $V_{in}^{(preamp)}$. Gain (ex. 0.1 < G < 0.9).
- Number of levels of the ADC = 2^{16} (ex. $2^{16} = 65,536$ levels).
- ADC resolution = $V_{in}^{(ADC)} / N$ levels (ex. $1V / 65,536 = 15.26 \mu V$).
- ADC dynamic that is actually used = $V_{in}^{(preamp)} / \text{ADC resolution}$.
- Bits that are actually used = $\log_2$ (levels actually used) (ex. $\log_2(58,98) = 15.87$).

● Connection to a Power Amplifier

In this case we have to configure both the pre and power amps in order to get a well-set system. Let's see what we have to configure and how.

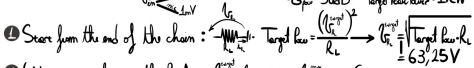
● Power amplifier: its main parameters are:

- $V_{in}^{(power)}$ → input signal peak provided by the preamp output.
- G_{power} → gain of the power amplifier.
- $V_{out}^{(power)}$ → maximum output voltage that the preamp can provide without clipping.

Our aim is to provide to the preamp a $V_{in}^{(preamp)}$ as high as possible so that we can reach the $V_{out}^{(power)}$ using the lowest G_{power} .

● Pre amplifier: in this case the aim is to provide the correct voltage to the preamp.

Let's see an example:



① Start from the end of the chain: $V_{out}^{(power)} = \frac{V_{in}^{(power)}}{R_L} \cdot R_L$. Target $V_{out} = 63.25V$.

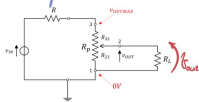
② We can now focus on the PowerAmp: $V_{out}^{(power)} = \frac{V_{in}^{(power)}}{R_L} \cdot R_L$. $V_{in}^{(power)} = \frac{V_{out}^{(power)}}{G_{power}} = \frac{63.25V}{(10^{16})} = 1.262V$.

③ Now we know which should be the preamp output: $V_{in}^{(preamp)} = V_{in}^{(power)} = \frac{V_{out}^{(power)}}{G_{power}} = \frac{63.25V}{(10^{16})} = 1.262V$.

Volume control

In this section we want to control the level of a signal at the input of the preamp (instead of modify the gain). We do so using a component called **fader**:

Simplest version

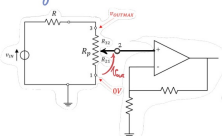


● In this way, the voltage that enters our system is V_{in} on R_L , that can be set modifying the value of R_{21} . In fact:

$$V_{out} = V_{in} \cdot \frac{R_{21}/R_L}{R + R_{32} + R_{21}/R_L} \quad \text{and} \quad V_{out}^{max} = V_{in} \cdot \frac{R_L/R_L}{R + R_L/R_L}$$

The problem is that with this approach we lose a portion of amplitude of V_{in} due to the parallel $R_{21} // R_L$.

Perfect version



● Now R_{21} is placed in parallel with the input impedance of an op-amp, that is ideally infinite. This means that $R_{21} // R_{in,op} \approx R_{21}$.

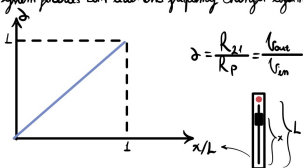
$$V_{out} = V_{in} \cdot \frac{R_{21}}{R + R_{32} + R_{21}} \quad \text{and} \quad V_{out}^{max} = V_{in} \cdot \frac{R_p}{R + R_p}$$

Usually, $R_p \gg R$, so we can approximate the equations as: $V_{out} = V_{in} \cdot \frac{R_{21}}{R_{32} + R_{21}}$ and $V_{out}^{max} = V_{in}$.

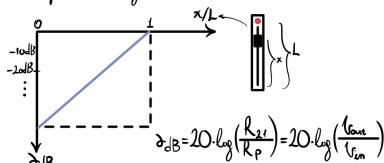
Conventionally, the attenuation is quantified by a factor α (see types of potentiometers).

● **Types of potentiometers** → there are two main types of potentiometers used for volume or gain control:

• **Linear potentiometers**: it is on an attenuator that has a linear attenuation response in a linear scale. It is not very used in audio application because human hearing system processes both level and frequency changes logarithmically.



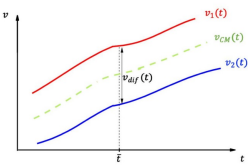
• **Logarithmic potentiometers**: it is on an attenuator that has a linear attenuation response in a logarithmic scale.



	Linear geometry	Circular geometry	
			$\alpha_{dB} \stackrel{\text{def}}{=} 20 \log \left(\frac{V_{out}}{V_{in}} \right) = 20 \log \left(\frac{R_{21}}{R_p} \right) = m \cdot \left(\frac{x}{L} - 1 \right)$
Linear function	$R_{21}(x) = R_p \left(\frac{x}{L} \right)$	$R_{21}(\theta) = R_p \left(\frac{\theta}{\theta_{max}} \right)$	
Logarithmic function	attenuation function $\alpha_{dB} = m \cdot \left(\frac{x}{L} - 1 \right)$ $R_{21} = R_p \cdot 10^{\frac{m}{20} \left(\frac{x}{L} - 1 \right)}$	attenuation function $\alpha_{dB} = m \cdot \left(\frac{\theta}{\theta_{max}} \right)$ $R_{21} = R_p \cdot 10^{\frac{m}{20} \left(\frac{\theta}{\theta_{max}} \right)}$	

Differential Amplifiers

- **Differential signals** → before introducing the differential amplifiers, we need to understand few concepts. Let's consider two signals $v_1(t)$ and $v_2(t)$.



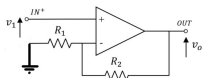
- There are two resulting signals that we can obtain from the interaction between v_1 and v_2 :
- **Differential signal** $\rightarrow v_{diff}(t) = v_1(t) - v_2(t)$. • **Common mode signal** $\rightarrow v_{cm} = \frac{v_1(t) + v_2(t)}{2}$.
- Starting from v_{diff} and v_{cm} it is possible to reconstruct the original signals:
 - $v_1(t) = v_{cm} + v_{diff}/2$.
 - $v_2(t) = v_{cm} - v_{diff}/2$.
- If v_1 and v_2 are much to give:
 - $v_{diff} = 0$, the respective v_{cm} is called **pure common mode signal**. In that case $v_1(t) = v_2(t) = v_{cm}(t)$.
 - $v_{cm} = 0$, the respective v_{diff} is called **pure differential signal**. In that case $v_1(t) = -v_2(t) = v_{diff}/2$.

● Differential feedback precompiler $\rightarrow$ let's start analyzing an approach that is not actually used (HASOCHISUPE). The comp that we've seen in the past chapters can be used as a differential compiler but, for the reasons we will study below, it is not the ideal solution in that sense.

- **Classic configuration** → Let's see how a "classic opamp" behaves if we feed it with a differential input:

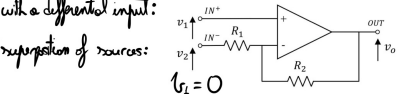
● We want to calculate the Gpm of this configuration, so let's use the principle of superposition of sources:

$$v_2 = 0$$



- It is a simple not-inverting configura-
tion, so the gain is:

$$G^+ = 1 + \frac{R_2}{R_1}$$



- It is a simple inverting configuration, so the gain is:

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

This means that the total output of the system would be $V_{out} = G^+ \cdot G^- + G^- \cdot G^+$. The problem here is that G^+ is very different from G^- but why do we want to have $G^+ = G^-$? To understand that we have to introduce the concept of **Common Mode Rejection Ratio**: The CMRR is the ability of an amplifier configuration to reject the part of the signal that the two input voltages have in common (Common Mode signal). This common component is considered by noise and disturbances that can be "ignored" if the amplifier has a high CMRR.

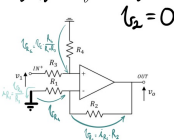
The Common Mode Rejection Rate is defined as: $CMRR = \frac{A_{diff}}{A_{cm}}$.
 A_{diff} Gain for differential signals.
 A_{cm} Gain for common mode signals.

So, if we consider an opamp fed with V_1 and V_2 , we would have that:

• $b_{out} = b_2 \cdot G^+ \cdot b_2^- = (b_{cu} \cdot b_{li}^{(up)}) \cdot G_1^+ \cdot (b_{cu} \cdot b_{li}^{(up)}) \cdot G_2^-$ if $G_1 \rightarrow -G_2 \rightarrow (b_{cu} \cdot b_{li}^{(up)}) \cdot G_1^+ \cdot (b_{cu} \cdot b_{li}^{(up)}) \cdot -G_2 = (b_{li}^{(up)} - b_{li}^{(up)}) \cdot G_1^+$ Only the "unique part" of the signal is amplified.
if $G_1 \rightarrow G_2 \rightarrow b_{cu} \cdot (G_1^+ \cdot G_2) \cdot b_{li}^{(up)} \cdot G_2^-$ the common mode (noise and disturbance) is amplified.

That's why using a "simple" operational amplifier is not an intelligent way to build a differential amplifier.

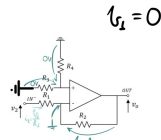
● Improved configuration → on easy solution could be the one of applying two simple renterform of the + input, in order to balance the true gains choosing the right values of R_1 and R_2 . Let's apply the superposition of effects again to find which could be the ideal value for these resistances.



$$\bullet G_{out} = G_{R1} + G_{R2} = G_{R1} \cdot \frac{R_4}{R_1 \cdot R_3} \left(1 + \frac{R_2}{R_1} \right)$$

$$G_{R1} \cdot \frac{R_4}{R_1 \cdot R_3} \quad G_{R1} \cdot \frac{R_4}{R_1 \cdot R_3} \cdot \frac{R_2}{R_1}$$

$$\bullet G^+ = \frac{R_4}{R_1 \cdot R_3} \left(1 + \frac{R_2}{R_1} \right)$$



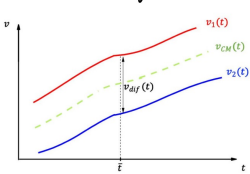
- $I_{G_{eq}} = -I_{G_2} = -I_2 \cdot \frac{R_2}{R_1}$
- $G = -\frac{R_2}{R_1}$

• We want to impose that $G^+ = -G^-$, so: $\frac{R_3}{R_1} \left(1 + \frac{R_2}{R_1} \right) = - \left(\frac{R_2}{R_1} \right) \rightarrow$ we need $R_4 = R_2$ and $R_3 = R_1$.

● Ok, so we finally found a relation. Didn't we? Actually no. "I need g_1 resonant holes values that do not perfectly coincide with their nominal values, so it is impossible to get two resonator of exactly the same values and we cannot impose $R_{\text{eq}} = R_2$ and $R_3 = R_1$ ". It's necessary to find a more stable configuration.

Differential Amplifiers

● **Differential signals** → before introducing the differential amplifiers, we need to understand few concepts.
 Let's consider two signals $v_1(t)$ and $v_2(t)$.



● There are two resulting signals that we can obtain from the interaction between v_1 and v_2 :

● **Differential signal** → $v_{diff}(t) = v_1(t) - v_2(t)$.

● **Common mode signal** → $v_{cm} = \frac{v_1(t) + v_2(t)}{2}$.

● Starting from v_{diff} and v_{cm} it's possible to reconstruct the original signals:

$$\begin{aligned} v_1(t) &= v_{cm} + v_{diff}/2 \\ v_2(t) &= v_{cm} - v_{diff}/2 \end{aligned}$$

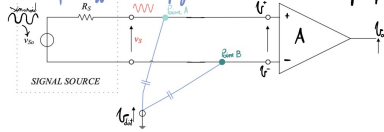
● If v_1 and v_2 are such to give:

● $v_{diff} = 0$, the respective v_{cm} is called **pure common mode signal**. In that case $v_1(t) = v_2(t) = v_{cm}(t)$.

● $v_{cm} = 0$, the respective v_{diff} is called **pure differential signal**. In that case $v_1(t) = -v_2(t) = v_{diff}/2$.

● **How to find a differential (or) amplifier** → Now we want to see which is the best "input configuration" for a CM-rejecting diff amplifier.
 Let's see it step by step.

● **Simple differential amplifier** → we start with the simplest possible configuration. We consider a simple differential amplifier and its behavior in presence of a disturbance.

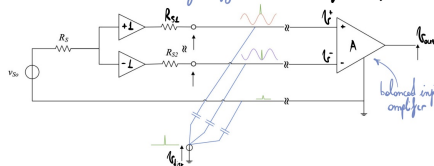


● The injected disturbance has a different effect on each one of the input pins of the amplifier:

- **Point B** → the disturbance is injected on the ground line, and it is very attenuated. The negative pin of the amplifier is connected to that cable, so $v_{cm} = 0$.
- **Point A** → the disturbance is injected on the signal wire and passes almost undisturbed to the positive pin, so $v_{cm} = v_{diff}$.

● This behavior gives as result $v_{out} = A(v_+ - v_-) = A(v_{diff} - 0) \rightarrow v_{cm}$ is amplified undisturbed → very bad differential amplifier.

● **Common-mode cancelling configuration** → a first easy solution could be the following:



● In this configuration we put two unitary gain amplifiers with $G = \pm 1$ in order to split the input signal in a pure differential signal.
 In order to understand the result of this approach, let's write the final formula in an explicit but messy:

$$v_{out} = A(v_+ - v_-) = G \cdot v_+ - G \cdot v_- = \left(\frac{G}{2} \cdot G\right) \cdot v_{diff} + (G \cdot G) \cdot v_{cm}$$

In our specific case, it becomes $v_{out} = G_{diff} \cdot v_{diff} + G_{cm} \cdot v_{cm} = G_{diff}(2G) + G_{cm}(v_{cm})$.

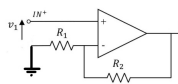
● This means that with this approach, if we use a high CMRR balanced input amplifier, we can cancel common mode disturbances.

● **Internal architecture of a balanced input amplifier** → we will build step-by-step the "perfect balanced-input differential amplifier".

● **Simple op. Amplifier** → Let's see how a "classic opamp" behaves if we feed it with a balanced input:

● We want to calculate the gain of this configuration, so let's use the principle of superposition of sources:

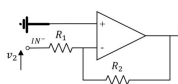
$$v_2 = 0$$



● It is a simple non-inverting configuration, so the gain is:

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

$$v_1 = 0$$



● It is a simple inverting configuration, so the gain is:

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

This means that the total output of the system would be $v_{out} = v_1 \cdot G + v_2 \cdot G$. The problem here is that G is very different from G but why do we want to have $G = G$? To understand that we have to introduce the concept of **Common Mode Rejection Ratio**: the CMRR is the ability of an amplifier configuration to reject the part of the signal that the two input voltages have in common (Common Mode signal). This common component is constituted by noise and disturbances, that can be "ignored" if the amplifier has a high CMRR.

The Common Mode Rejection Ratio is defined as: $CMRR = \frac{A_{diff}}{A_{cm}}$.
 So, if we consider an amplifier fed with v_1 and v_2 , we would have that:

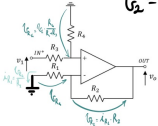
$$v_{out} = v_1 \cdot G + v_2 \cdot G = (v_{cm} + v_{diff}/2) \cdot G + (v_{cm} - v_{diff}/2) \cdot G = v_{cm} \cdot (G + G) + v_{diff} \cdot (G - G)$$

Only the "unique" part of the signal is amplified.
 The common mode (noise and disturbance) is cancelled.

That's why using a "simple" operational amplifier is not an intelligent way to build a differential amplifier.

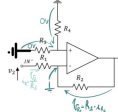
- **Improved configuration 1** → on easy solution could be the one of applying two simple resistors at the + input, in order to balance the two gains choosing the right values of R_3 and R_4 . Let's apply the superposition of effects again to find which could be the ideal value for these resistors.

$$V_2 = 0$$

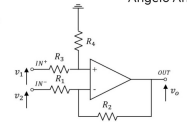


$$\begin{aligned} \bullet V_{out} &= V_2 + V_1 = V_1 = V_1 \cdot \frac{R_3}{R_1 + R_3} \left(1 + \frac{R_2}{R_4} \right) \\ \bullet G^+ &= \frac{R_3}{R_1 + R_3} \left(1 + \frac{R_2}{R_4} \right) \end{aligned}$$

$$V_1 = 0$$



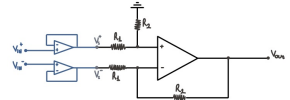
$$\begin{aligned} \bullet V_{out} &= -V_2 = -V_2 \cdot \frac{R_2}{R_4} \\ \bullet G^- &= -\frac{R_2}{R_4} \end{aligned}$$



- We want to improve that $G^- = -G^+$, so: $\frac{R_3}{R_1 + R_3} \left(1 + \frac{R_2}{R_4} \right) = - \left(-\frac{R_2}{R_4} \right)$ → we need $R_4 = R_2$ and $R_3 = R_1$.

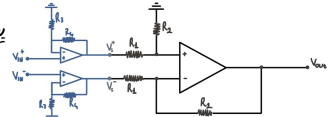
- Or, we finally found a solution. Didn't we? Actually no. In real life, resistors have values that do not perfectly coincide with their nominal value, so it is impossible to get two resistors of exactly the same value and we cannot impose $R_4 = R_2$ and $R_3 = R_1$. It's necessary to find a more stable configuration.

- **Improved configuration 2** → in order to minimize the impact that resistance values imprecisions have on the system, we add two buffers (opamp having $G=1$). In this way, the resistance on the two "balances" is so high that small resistor variation have almost no-effect on the circuit.



The the problem in this configuration is that we are adding a component → we are adding noise without amplify the signal. This leads to a degradation of the SNR. We need to amplify the signal.

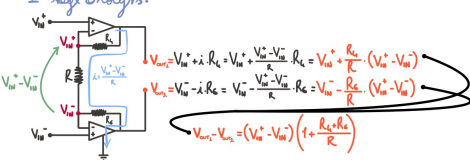
- **Improved configuration 3** → in order to solve the problem of SNR of the previous configuration, we introduce a voltage amplification (gain > 1) at the first stage of the circuit. Again, we solved a problem, but we introduced a new one: R_4 and R_5 and R_3 and R_1 for the same reasons that we saw in the "improved configuration 1". This makes $G_{in} \neq G_{out}$, degrading CMRR.



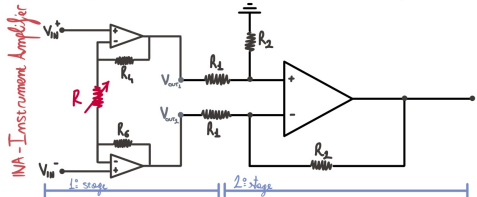
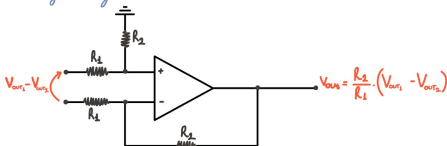
We need to solve this last problem.

- **Improved configuration 4: advanced differential (psu) amplifier** → we use merged together R_3 and R_1 of the previous configuration creating R . Let's analyze the behaviour of this configuration:

- **1° stage analysis:**

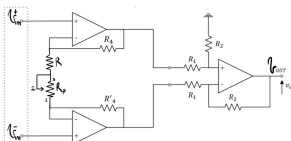


- **2° stage analysis:**



- This configuration has one problem: R is a potentiometer that goes from 0 to R . But if it goes to 0 the gain goes to infinity, making the output diverge. We need to find a solution to this.

- **Improved configuration 5:** to avoid the single resistors to be not at zero, we put it in series with another resistor, that will be the minimum R value.

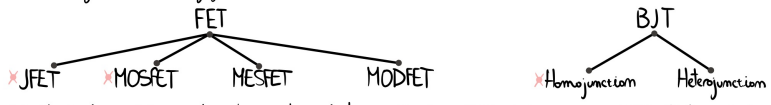


- In that case, the gain is: $G_{TOT} = \frac{R_2}{R_1} \left(1 + \frac{R_0 + R_6}{R \cdot R_{11}} \right)$

Transistors

Transistors can be categorized in two big families:

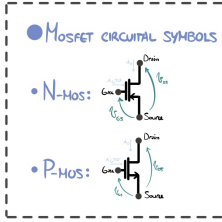
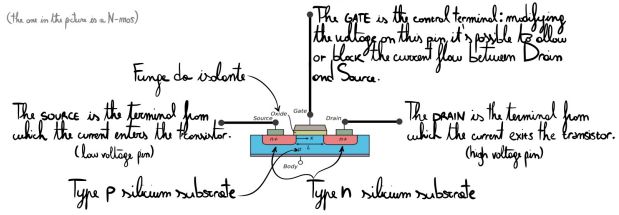
✗ Used for audio



We will study the topologies that are most used in audio applications: MOSFET, JFET and homojunction BJT. Before starting our analysis we have to introduce the concept of **transconductance**: the transconductance g_m is a quantity that measures the efficiency of a transistor in converting a voltage variation applied to its input (gate or base, as we will see) into a current variation of its output (drain or collector). It is defined as $g_m = \frac{\partial I_D}{\partial V_{GS}}$ and its measure unit is Siemens [S].

• **MOSFET** → Metal Oxide Semiconductor field Effect Transistors are **voltage controlled** transistors characterized by a **high input impedance** that makes absorbed current ≈ 0 and by a very **high commutation speeds**. They are constituted as follows:

(The one in the picture is a N-mos)



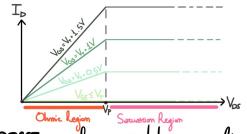
• **Working principle** → applying a voltage to the gate terminal we control the channel between SOURCE and DRAIN. Let's see the possible cases:

• **Cut-off**: there is no conductive channel and the transistor is 'turned off' $\left\{ \begin{array}{l} \text{for N-mos, this happens when } V_{GS} < V_{th} \\ \text{for P-mos, this happens when } V_{GS} < |V_{th}| \end{array} \right\}$ in this condition $I_D = I_S = I_D = 0$.

• **Varistor - ohmic region**: there is a conductive channel along all the source-drain distance and the transistor behaves as a varistor, meaning like a resistor that change its R value according to a control voltage that in this case is the voltage between the gate and the source. This condition happens when $\left\{ \begin{array}{l} \text{if N-mos, } V_{GS} < V_{th} \text{ and } V_{DS} < V_{GS} - V_{th} \\ \text{if P-mos, } V_{GS} < |V_{th}| \text{ and } |V_{SD}| < |V_{GS}| - |V_{th}| \end{array} \right\}$. It's possible to approximate $I_D = I_S = \mu_n C_{ox} \frac{W}{L} \left[(V_{GS} - V_{th}) \left(V_{DS} - \frac{V_{DS}^2}{2} \right) \right]$ after called I_{D0} .

• **Saturation**: when, in ohmic region, we increase 'too much' the V_{GS} (for P-mos) a characteristic phenomenon occurs: at the end of the conductive channel, the channel itself become 'narrower' and from that moment, if we further increase the control voltage, this action will not correspond to an increasing of the I_D voltage: the transistor's behavior can be described to the one of a current generator. This condition occurs when $\left\{ \begin{array}{l} \text{for N-Mos, if } V_{GS} > V_{th} \text{ and } V_{DS} > V_{GS} - V_{th} \\ \text{for P-Mos, if } V_{GS} > |V_{th}| \text{ and } |V_{SD}| > |V_{GS}| - |V_{th}| \end{array} \right\}$. It's possible to approximate $I_D = I_S = \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{th})^2$.

This 'sloping' made in also called **transconductor stat**, and the transconductor value can be calculated as $g_m = 2 \sqrt{I_D \cdot I_0}$.



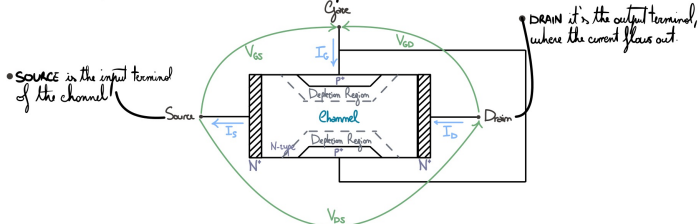
• **Small signals model for MOSFETs** → MOSFETs are characterized by a non linear input-output relation, but we don't have to worry with non-linear equations. There is a relation to that:

• If we consider a MOSFET working around its operating point (one of the red points in the draw) and considering only **little signal variations** (little AC component), it is possible to approximate the behaviour of the transistor as it was linear. The equivalent circuit is the following:

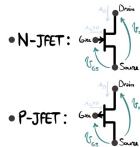
• **JFET** → Junction Field Effect Transistors are characterized by a ^{higher than MOSFETs} very high input impedance and a very low noise level, making it a very diff. type of transistor in audio applications. On the other hand, they are not very suited to drive "heavy" loads because of their output current limitations and they have a lower transconductance g_m , limiting the gain in some applications.

• It is a "normally on" device, meaning that if $V_{GS} = 0$ the conductive channel is open.

• **GATE** is the control pin: changing the tension on this terminal we modify the current that flows between drain and source.



JFET CIRCUITAL SYMBOLS



• **Working principle** → it is possible to manage the current that flows through the channel, making it (the channel) wider or thinner by changing the "negative polarization" of the gate: the higher the polarization is, the narrower the channel becomes (the depletion region increases). There are three possible cases:

- **Cut-off**: in this condition there is no conductive channel between source and drain. It occurs:
 - for N-JFETs, if $V_{GS} < V_{TH}$
 - for P-JFETs, if $V_{GS} > |V_{TH}|$
 In this condition $i_D \approx i_S \approx 0$.
- **Variacor-Ohmic region**: in this condition the transistor works as a voltage controlled resistor. It occurs:
 - for N-JFETs, if $V_{GS} > V_{TH}$ & $V_{DS} < V_{GS} - V_{TH}$
 - for P-JFETs, if $V_{GS} > |V_{TH}|$ & $V_{SD} < V_{GS} - |V_{TH}|$

where V_{GS} is the gate-source voltage, and V_{TH} is the threshold voltage.

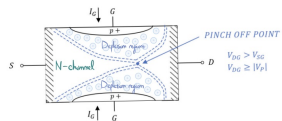
In this condition it is possible to approximate the current equation linearly:

$$\begin{aligned} \text{for N-JFET, } i_D \approx i_S &\approx \frac{2I_{DSS}}{V_{TH}} \left(1 - \frac{V_{GS}}{V_{TH}}\right) \cdot V_{DS} \\ \text{for P-JFET, } i_D \approx i_S &\approx \frac{2I_{DSS}}{V_{TH}} \left(1 - \frac{V_{GS}}{|V_{TH}|}\right) \cdot V_{SD} \end{aligned}$$

• **Saturation**: if we increase "enough" the voltage between drain and source, the pinch-off phenomenon occurs: the conductive channel becomes narrower near to the "output" of the circuit, making the output current independent from V_{DS} further increments.

This condition occurs:

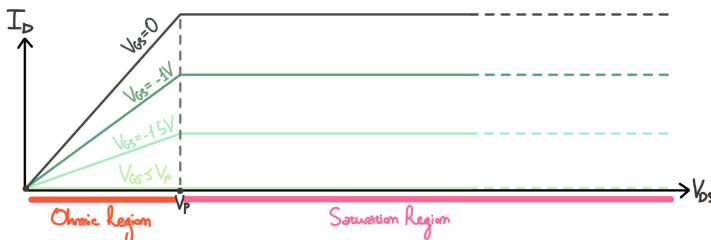
- for N-JFET, if $V_{GS} > V_{TH}$ & $V_{DS} > V_{GS} - V_{TH}$. In this case, $i_D \approx i_S \approx I_{DSS} \left(1 - \frac{V_{GS}}{V_{TH}}\right)^2$.
- for P-JFET, if $V_{GS} > |V_{TH}|$ & $V_{SD} > V_{GS} - |V_{TH}|$. In this case, $i_D \approx i_S \approx I_{DSS} \left(1 - \frac{V_{GS}}{|V_{TH}|}\right)^2$.



As we saw for MOSFETs, the saturation region is also called transconductive region, and the transconduction value in that case

is equal to:

- for N-Mos: $g_m = \frac{2I_{DSS}}{V_{GS} - V_{TH}}$
- for P-Mos: $g_m = \frac{2I_{DSS}}{V_{GS} - |V_{TH}|}$

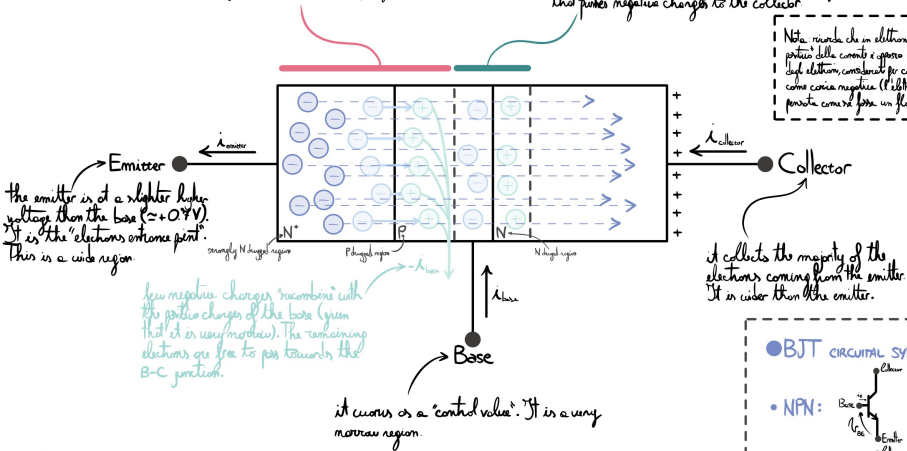


● **BJT** → Bipolar Junction Transistors are current controlled transistors characterized by low input impedance and high conductance, making it suitable to be applied for "high" amplifications.

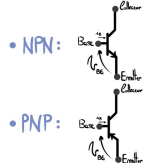
Emitter-Base junction: when $V_{BE} \geq V_{BE_{on}} \approx 0.7V$ the potential barrier between the emitter and the base disappears, allowing a lot of electrons to pass from E to B.

Base-Collector junction: if $V_{BC} = V_{BE} + V_{CE}$ the electrons that previously passed the E-B junction are pushed through the collector. The fact that $V_{BC} > V_{BE}$ generates an electric field that pushes negative charges to the collector.

Nota: ricorda che in elettronica il "corrente positiva" della corrente è spesso alla direzione degli elettroni, considerati per convenzione come carica negativa. (l'elettricità era pensata come se fosse un fluido).



● BJT CIRCUITAL SYMBOLS



● **Working principle** → there are three working regions for this type of transistor:

• **Cut-off:** the conductive channel is closed, $I_B \approx I_C \approx 0$. This occurs for NPN-BJT's if $V_{BE} < V_{BE_{on}}$.

• **Forward active-linear amplifier:** in this conditions, the transistor operates like a linear amplifier. This occurs for PNP-BJT's if $V_{BC} < V_{BE_{on}}$.

In this condition is possible to approximate the I_C equation as:

- for NPN-BJT's, $I_C = \beta \cdot I_B = I_S \cdot e^{\frac{V_{BE}}{V_T}}$
- for PNP-BJT's, $I_C = \beta \cdot I_B = I_S \cdot e^{\frac{V_{BE}}{V_T}}$

where I_S is the saturation current, V_T is the thermal voltage, and β is the current gain.

Thus in the forward active region for BJTs, and the transconductance value is $g_m = \frac{I_C}{V_T}$.

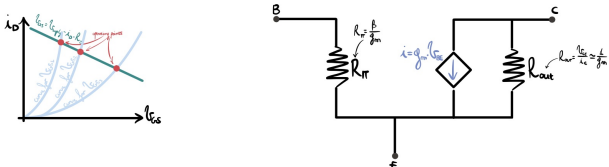
• **Saturation:** in this region the transistor is "turned on at maximum power" and an increase of I_B doesn't lead I_C to increase. This working region is often used to build switches, because the behaviour of the transistor is similar to the one of a relay-circuit.

This condition occurs for NPN-BJT's, if $V_{BE} > V_{BE_{on}}$ and $V_{BC} < V_{BE_{on}}$.

for PNP-BJT's, if $V_{BE} > V_{BE_{on}}$ and $V_{BC} < V_{BE_{on}}$.

Note: usually $V_{BE_{on}} = 0.7V$ and $V_{CE_{sat}} = 0.2V$

● **Small signal model for BJTs** → in the same way we did for MOSFET, we can build a "small signal model" for BJTs, meaning a model that approximates the transistor characteristic curve or a straight line around operating points where AC component is very low.



● **Dissipated power** → the power dissipated by this type of transistor is: $P_{dissipated} \approx I_{C_{sat}} \cdot V_{CE_{sat}}$

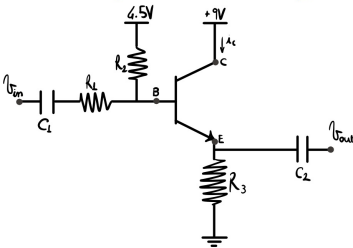
Useful BJT implementations

Common emitter amplifier → it is a BJT configuration used to amplify a voltage input signal. The name of this configuration comes from the fact that the emitter and the output circuit have the emitter pin in common (connected to ground).

- Let's move on example to understand how it works:
 $V_{CC}=12V$, $R_C=5k\Omega$, $I_C=1mA$, $V_{CE} \approx 25mV$.
 and $\Rightarrow V_{CE}=V_C$, $V_C=5V$ and $V_E=V_{CE}=7V$.
- It's more possible to calculate the transconductance:
 $g_m = \frac{I_C}{V_T} = 0.04S$
- In this type of configuration the gain is $G = -g_m R_C = -225$.
- If for example, we feed the input pin (Base) with a 100mV sinusoidal amplitude $\Delta V_{in} = 10mV$ we obtain in output another sinusoidal amplitude $\Delta V_{out} = 2.25V$, achieving an amplification of the signal.

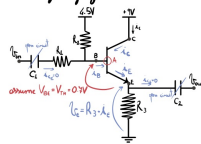
This configuration is subject to distortions and gain modulation due to temperature changes. We use it instead of feedback opamp because it's cheaper.

Further analysis on emitter follower configuration (from tube screamer)



In the "tube screamer" circuit we saw this emitter follower. We can assert that it's an emitter follower because the output is placed on the emitter pin. The across "emitter" from the emitter, it's a NON-BJT.

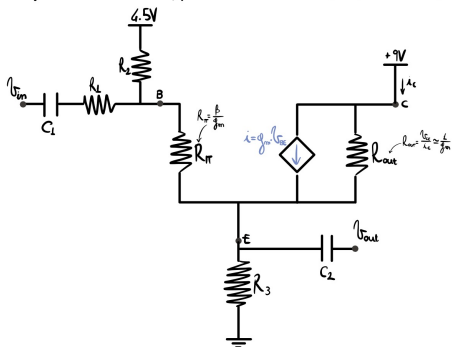
Let's first perform a DC analysis, to individuate the operating point of the circuit:



- We know that base-collector current relation: $I_C = \beta \cdot I_B$.
 KCL in A: $I_B + I_C + I_E = 0 \Rightarrow I_E = I_B + I_C$.
 So, putting all together, we get that $I_E = I_B (1 + \beta)$.
- We can also see that $V_{BE} = V_B - V_E \Rightarrow V_E = V_B - V_{BE}$.
 $I_E = \frac{V_E}{R_3} \Rightarrow I_E = \frac{V_B - V_{BE}}{R_3}$.
 So, putting all together, we get that $I_E = \frac{V_B - V_{BE}}{R_3}$.

- With the informations gained until now, we can write the equation $\frac{V_B - V_{BE}}{R_3} = I_E = I_B (1 + \beta)$. In order to get V_B , we must know the value of I_B . Let's find it: $I_B = (4.5V - V_B) / R_2$.
- Let's solve the final equation: $\frac{V_B - V_{BE}}{R_3} = \frac{(4.5V - V_B)}{R_2} (1 + \beta) \Rightarrow V_B R_2 - V_{BE} R_2 = 4.5V R_3 (1 + \beta) - V_B R_3 (1 + \beta) \Rightarrow V_B (R_2 + R_3 (1 + \beta)) = 4.5V R_3 (1 + \beta) + V_{BE} R_2 \Rightarrow V_B = \frac{4.5V R_3 (1 + \beta) + V_{BE} R_2}{R_2 + R_3 (1 + \beta)}$
- From V_B we can easily get $V_E = V_B - 0.7V$.
- Before we calculated that $I_E = I_B (1 + \beta)$, while we know that $I_C = I_B \cdot \beta$. Actually, usually β is very big while I_B is very low, so $I_C \approx I_E$. Then, we can calculate the transconductance $g_m = \frac{I_C}{V_T}$ and the input impedance of the BJT $r_{\pi} = \frac{\beta}{g_m} = \frac{\beta V_T}{I_C} \approx \frac{1}{g_m}$.

If we want to be very precise and consider the effect of R_{in} and R_{out} of the BJT, we can use the small signals model:



Using this representation we could demonstrate that $Z_{in} = R_{in} + (\beta + 1) R_E$ but I'm too tired to study the calculations, I'll trust the prof.

$$Z_{out} = \frac{R_E + R_{out}}{\beta + 1}$$

Common collector amplifier (emitter follower) → this is a power amplifier: its voltage gain is ≈ 1 (<1 if it is not ideal). We use it to amplify the signal's power (current) in order to make it adapt to drive a load output impedance.

Let's move on example to understand how it works:
 $V_{CC}=12V$, $R_C=5k\Omega$, $I_C=1mA$, $V_{CE} \approx 25mV$.

Let's imagine we want to obtain an output tension in DC of $V_C = 6V$. We know that (in order to put the BJT in amplifying region) $V_{CE} \approx 0.7V$, so the base tension must be $V_{BE} = 0.7V$ on the emitter: $V_{CE} = V_C - V_E = 6V - 0.7V = 5.3V$.

As we saw, $V_{CE} = 12V$ and we want $V_C = 6V$, then means that $V_{CE} = V_C - V_E = 6V$. Knowing that we can choose the correct resistor: $R_E = V_E / I_E = 6k\Omega$.

We can also calculate the intrinsic resistance of the emitter $r_e = V_{Tm} / I_E \approx 25\Omega$.

With all these informations we can calculate the voltage gain: $G = \frac{R_E}{r_e + R_E} = 0.995 \approx 1$.
 How we can see we're not amplifying the voltage: it's a buffer.

Riassunto sui transistor per comprenderli meglio → revisione la cosa già spiegata nelle pagine precedenti per assicurarsi di averle capite bene.

MOSFET



• **Costruzione e funzionamento:** su una lastra di silicio P sono impiantate due regioni a drogaggio N, separate da una zona nel quale è montata una regione di ossido, chiamata al gate.

• **Cutoff:** quando non è applicata alcuna tensione al gate ($V_{GS} < V_{TH}$) il canale è chiuso perché gli elettroni che vorrebbero attraversarlo trovano la barriera P come ostacolo.

• **Linear region:** quando viene applicata una tensione al gate ($V_{GS} > V_{TH}$) gli elettroni acquisiscono abbastanza energia per saltare da una regione P all'altra, aprendo il canale. In tali situazioni, la quantità di corrente che passa è proporzionale alla tensione applicata al gate:

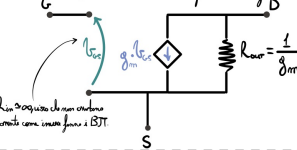
$$I_D \approx \frac{\mu_n C_{ox} W}{L} \cdot (V_{GS} - V_{TH}) \cdot (V_{GS} - \frac{V_{DS}}{2})$$

• **Saturation region:** se aumenti anche la tensione tra il gate e il source ($V_{GS} > V_{TH}$ e $V_{DS} > V_{GS} - V_{TH}$) il canale si apre completamente ed aumentano i carichi di corrente. V_{DS} non possono aumentare nel pozzo di corrente del canale. In tali condizioni, il transistor può essere considerato come un generatore di corrente I_D :

$$I_D \approx \frac{1}{2} \cdot (V_{GS} - V_{TH})^2$$

• **Small signals model:** MOSFET are non-linear components, but can be considered as linear around their operating point (I_D, V_{GS}).

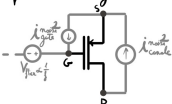
In that case, the MOSFET can be represented as follow:



Ricorda: questa è una non lineare, come invece fanno i BJT

• **Elettrici noise in MOSFET** → in un MOSFET la principali sorgenti di rumore sono:

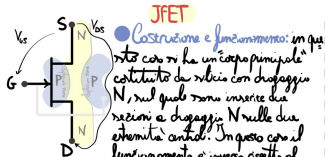
- **Rumore termico:** è dovuto all'agitazione termica dei portatori di carica all'interno del canale conduttivo. Si genera con una corrente di drain: $I_{noise} = 4kTg_{D0}$.
- **Shot noise:** molto più alta, ma presente: $I_{noise} = 2qI_D$.
- **Fluor** → a basse frequenze offre un rumore di tipo fluor. Si modella come una sorgente di tensione posta in serie al gate.



• **Comparazione tra rumore MOSFET e JFET** → come appena visto, questi due tipi di transistor sono molto simili. Ci sono però alcune considerazioni da fare:

- **Rumore termico:** per i MOSFET è $4kTg_{D0}$, mentre per JFET è $4kTg_{D0}$ (il livello di fluor è molto più basso).
- **Rumore di fluor:** i MOSFET presentano fluor non trascurabile nelle basse frequenze, mentre nei JFET il rumore di fluor è più basso.
- **Interferenza:** i MOSFET sono più suscettibili al rumore ambientale.

JFET



• **Costruzione e funzionamento:** in questo caso si ha un corpo p-n-p-n-p, costituito da silicio con drogaggio N, nel quale sono inserite due regioni a drogaggio P nelle due estremità centrali. In questo caso il funzionamento è inverso rispetto al MOSFET: il canale è "normalmente" aperto e può essere chiuso agendo sul gate.

• **Chiusa region:** in condizioni "statiche" ($V_G < V_{GS} < V_{GS}^{sat}$) il canale elettrico resta aperto. L'ampiezza del canale dipende dalla tensione V_{GS} quindi il transistor si comporta come un resistor.

$$I_D = \frac{2 \cdot I_{D0}^{sat}}{V_G} \cdot \left[(V_{GS} - V_G) \cdot (V_{GS} - \frac{V_G}{2}) \right]$$

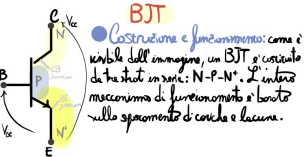
$$R_{DS} = \left[\frac{2 \cdot I_{D0}^{sat}}{V_G} \cdot (V_{GS} - V_G) \right]^{-1}$$

• **Saturation region:** quando, in condizioni dinamiche, la tensione tra il gate e il source aumenta ($V_{GS} > V_{GS} - V_G - V_{GS}^{sat}$), si ha un "collo di bottiglia" intorno al drain (pinch off) che impedisce alle cariche di aumentare ulteriormente il suo flusso all'interno di V_{GS} . Si verifica quindi, in condizioni di saturazione e il comportamento del transistor è simile ad un generatore di corrente I_D .

$$I_D = I_{D0}^{sat} \cdot (1 - \frac{V_{GS}}{V_G})^2$$

• **Cutoff:** se applichiamo "abbastanza" tensione al gate ($V_G < V_G$) le due estremità si separano fino a catturare tutti gli elettroni, chiudendo il canale elettrico → $I_D = 0$.

BJT



• **Costruzione e funzionamento:** come è visibile dall'immagine, un BJT si caratterizza da tre strati in serie: N-P-N. L'intero meccanismo di funzionamento è basato sullo spostamento di cariche e lacune.

• **Cutoff:** se la tensione tra il collector e la base è bassa ($V_{CE} < 0.7V$) gli elettroni non possono attraversare la giunzione BE e a causa delle differenze di potenziale non entrano nel pozzo BE, non ci sono neanche nella giunzione CB, per cui $I_C = I_E = 0$.

• **Forward Active:** se la tensione V_{BE} è abbastanza elevata ($V_{BE} > 0.7V$) alcuni elettroni della zona N della emitter passano alla zona P, combinandosi con le lacune presenti in questa zona. Si generano così due correnti:

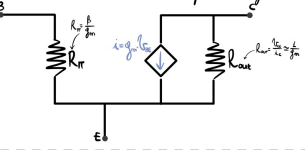
• I_E : gli elettroni si combinano con le lacune vicino alla base, creando una piccola corrente I_E .

• I_C : la maggior parte degli elettroni passano dalla zona P alla zona N del collector a causa della polarizzazione inversa, generando una corrente I_C .

• **Saturation:** se oltre alla tensione V_{BE} aumentiamo anche V_{CE} oltre una certa soglia, il canale si apre completamente e la corrente non aumenta né esponenzialmente, ma rimane fissa a I_{sat} .

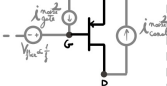
• **Small signals model:** BJTs are non-linear components, but can be considered as linear around their operating point (I_D, V_{GS}).

In that case, the BJT can be represented as follow:



• **Elettrici noise in BJT** → in un BJT la principali fonti di rumore sono:

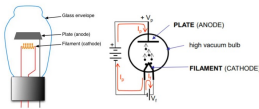
- **Rumore di shot della corrente di base:** come abbiamo visto, il flusso di corrente di un BJT è generato dalle cariche che attraversano da una zona N all'altra, attraversando la zona a drogaggio P. Tale flusso di cariche non è regolare ed è soggetto a variazioni aleatorie. Tali variazioni possono essere modellate come una corrente $I_{noise} = 2qI_B$.
- **Fluor:** è presente anche una componente I_{noise} di tipo fluor, che si modella come una corrente $I_{noise} = 2qI_B$.



Valves

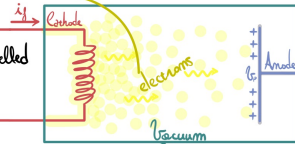
The valves are the first electronic devices to be built, and some modern electronic devices still use it. Different types of valves have been developed, each one solving the problems of the "previous versions".

● **Thermionic diode valve** → This component is constituted by:



Electrons → the electrons that are present in the filament (cathode) are very "energetic" because of the high filament temperature, and some of them "exit" the filament and are emitted in the vacuum. This phenomenon is called thermionic emission. The electrons that are free to move in the vacuum are attracted from the anode, thanks to the electromagnetic field generated by its positive polarization.

● **Cathode** → it is constituted by a filament travelled by a current i_f that brings its temperature at 1774 °C.



● **Anode** → it is a metallic plate partially polarized (usually at +200/300V).

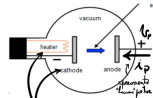
● **Vacuum** → in the valve a strong vacuum is created. This is useful to remove air molecules in order to let the electrons be free to move.

● **Thermionic diode: improved version** → the previous valve configuration has been improved separating the filament from the cathode. The physical laws that describe the behaviour of the valve diode are the following:

- **Off mode:** if $i_f \ll 0$, the electromagnetic field is not able to attract the electrons emitted by the cathode, so there is "no conductive channel" and $i_p \approx 0$.
- **Limited-emission regime:** if $i_f > 0$ but not too much, the current flow between cathode and anode is described by the formula:

$$i_p = i_s e^{\frac{eV}{kT}} \quad \text{where } i_s = S \cdot A \cdot T^2 \cdot e^{\frac{e\phi}{kT}}$$

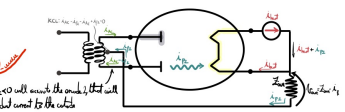
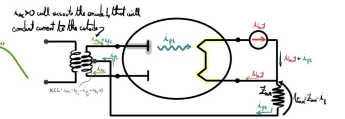
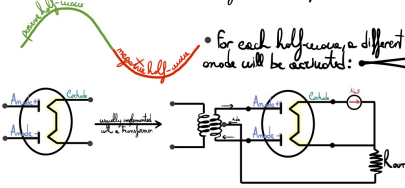
i_p : plate current
 i_s : saturation current
 S : Richardson constant
 A : area of the cathode
 T : cathode temperature
 e : electron charge
 V : anode-cathode voltage
 ϕ : work function



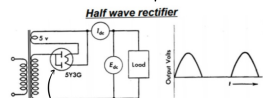
• in this case, the filament is used as a "heater" while the cathode is a plate separate from the filament and heated by this one.

- **Space Charge limited regime:** if $i_f \gg 0$, the amount of emitted electrons generates a "cloud" between anode and cathode. This cloud generates an electric field characterized by opposite direction to the one produced by the anode, limiting the current flow. In this case, the current flow is described by the equation: $i_p = K V^{1.5}$ (where K is a constant).

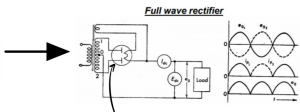
● **Double anode rectifying tube** → it is a variation of the thermionic diode, characterized by two anodes that share the same cathode. To understand how it works, let's imagine on AC input (sinusoidal):



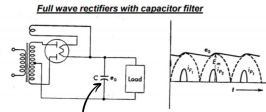
● Let's see how this component is used:



in this (simple) configuration the two anodes are connected by a short-circuit; the component works like a simple thermionic diode.

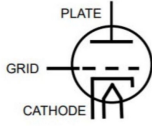


here the two anodes are connected to two different terminals of the transformer in order to rectify the full wave



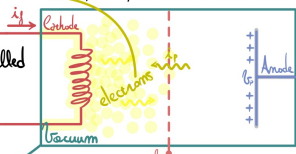
here we also add a capacitor in order to keep the output current "constant".

• **Triode** → the triode is an advancement over the diode, adding a third electrode, adding a third electrode called GRID, positioned between the cathode and the anode. This allows us to use the triode not only as a resistor but also as an amplifier.



$$i_p = i_g = K [V_g \log V_p]^M$$

• **Electrons** → the electrons that are present in the filament (cathode) are very "energetic" because of the high filament temperature, and some of them "exit" the filament and are emitted in the vacuum. This phenomenon is called thermionic emission. The electrons that are free to move in the vacuum, are attracted from the anode, thanks to the electromagnetic field generated by its positive polarization.



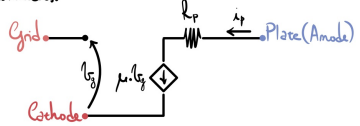
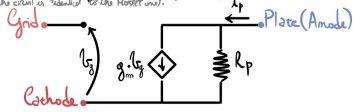
• **Cathode** → it is constituted by a filament travelled by a current i_f that brings its temperature at 774°C.

• **Anode** → it is a metallic plate partially polarized (usually at +200/400V).

• **Vacuum** → in the tube a strong vacuum is created. This is useful to remove air molecules in order to let the electrons be free to move.

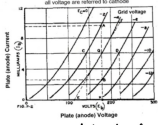
• **Control Grid** → this electrode acts as a "control pin": if it is negatively polarized it doesn't allow electrons to pass. Little variations of the grid tension lead to big variations in the current flow between anode and cathode. This gives the possibility to use the triode as a voltage controlled current amplifier.

• **Small signal model** → as we did with transistors, we can approximate the non-linear transfer function of the triode as a linear one, with the consequence that the tube is working around its operating point (i_p, V_g, V_p). There are two equivalent circuit models.

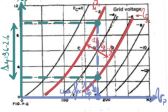


where the parameters are the following:

• **transconductance** → it describes how the current flow between anode and cathode changes when the voltage of the grid pin is modified: $g_m = \frac{\partial i_p}{\partial V_g} \Big|_{V_p = \text{constante}}$.

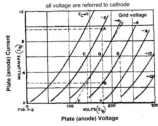


↳ V_g, V_p underscaled

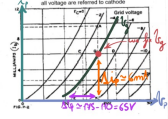


$$g_m = \frac{\partial i_p}{\partial V_g} \Big|_{V_p = \text{constante}} = \frac{9.6 - 2.6}{-2 - (-6)} \left[\frac{\text{mA}}{\text{V}} \right] = 4.75 \text{ mA/V}$$

• **plate resistance** → it describes how the plate voltage changes in function of the current flow: $R_p = \frac{\partial V_p}{\partial i_p} \Big|_{V_g = \text{constante}}$.

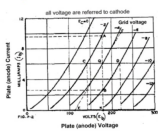


↳ V_g, V_p underscaled

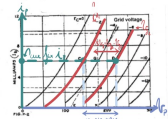


$$R_p = \frac{\partial V_p}{\partial i_p} \Big|_{V_g = \text{constante}} = \frac{65\text{V}}{6\text{mA}} = 10.8 \text{ k}\Omega$$

• **amplification factor** → represent the maximum voltage gain we can achieve with a device: $\mu = g_m \cdot R_p = - \frac{\partial V_p}{\partial V_g} \Big|_{i_p = \text{constante}}$.



↳ V_g, V_p underscaled



$$\mu = - \frac{\partial V_p}{\partial V_g} \Big|_{i_p = \text{constante}} = - \frac{218 - 162}{-4} \left[\frac{\text{V}}{\text{V}} \right] = 14$$

● **Tetrode** → the tetrode adds a fourth electrode, called **SCREEN GRID**, placed between the control grid and the anode.

Electrons → the electrons that are present in the filament (cathode) are very "energetic" because of the high filament temperature, and some of them "exit" the filament and are emitted in the vacuum. This phenomenon is called **thermionic emission**. The electrons that are free to move in the vacuum are attracted from the anode, thanks to the electromagnetic field generated by its positive polarization.

Cathode → it is constituted by a filament travelled by a current i_f that brings its temperature at 774°C.

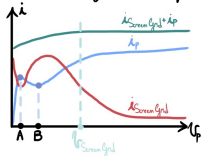


Screen Grid → it works as an electromagnetic shield, isolating the anode from the control grid. This is useful because it reduces the parasitic capacitance between anode and control grid.

Anode → it is a metallic plate partially polarized (usually at +200V/400V).

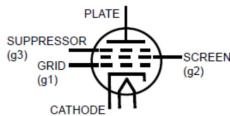
Vacuum → in the tube a strong vacuum is created. This is useful to remove air molecules in order to let the electrons be free to move.

● Let's analyze the component's characteristic curve:



- $V_{g1} = 0$: all the electrons emitted by the wire (cathode) are collected by the Screen Grid $\Rightarrow i_p = 0$.
- $0 < V_{g1} < V_{g2}$: increasing V_{g1} , the electrons emitted by the cathode start to be collected by the anode $\Rightarrow i_p$ increases and $i_{screen\ grid}$ decreases of the same quantity.
- $A < V_{g1} < V_{g2}$: here we have that:
 - the electrons coming from the wire hit the plate with enough strength to cause a secondary electron emission from the plate.
 - in this region ($V_{g1} < V_{g2}$), so the electrons emitted by the plate are collected by the screen grid.
 - $V_{g1} > V_{g2}$: at this point very few electrons are captured by the grid. This is the operative region of the tetrode.

● **The pentode** → here we try to eliminate the electron secondary emission. We do so through the introduction of another electrode, called **SUPPRESSOR GRID**, placed between the screen grid and the anode (plate).



Electrons → the electrons that are present in the filament (cathode) are very "energetic" because of the high filament temperature, and some of them "exit" the filament and are emitted in the vacuum. This phenomenon is called **thermionic emission**. The electrons that are free to move in the vacuum are attracted from the anode, thanks to the electromagnetic field generated by its positive polarization.

Cathode → it is constituted by a filament travelled by a current i_f that brings its temperature at 774°C.



Screen Grid → it works as an electromagnetic shield, isolating the anode from the control grid. This is useful because it reduces the parasitic capacitance between anode and control grid.

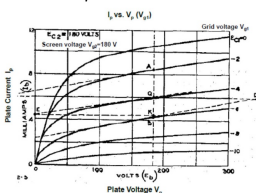
Anode → it is a metallic plate partially polarized (usually at +200V/400V).

Vacuum → in the tube a strong vacuum is created. This is useful to remove air molecules in order to let the electrons be free to move.

Control Grid → this electrode acts as a "control pin": if it is negatively polarized it doesn't allow electrons to pass. Little variations of the grid tension lead to big variations in the current flow between anode and cathode. This gives the possibility to use the tetrode as a voltage controlled current amplifier.

Suppressor Grid → it generates an electric field that always points to the plate (anode).

● Let's see the pentode characteristic curve:



Electronic noise

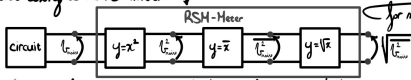
• **What is electronic noise?** It's a signal to transmit a signal that is a constant voltage. If we zoom-in on the real signal, we will see something similar to this: The signal is subject to random fluctuations, that are not avoidable and are caused by how the charges move in the circuits.

Summarizing a real signal is composed as: $V_{real} = V_{ideal} + V_{noise}$, with $E[V_{noise}] = 0$.
mean, deterministic and given "favorable" variations of the signal, we can fit them

• **How to measure noise** → It is find the best way to measure noise.

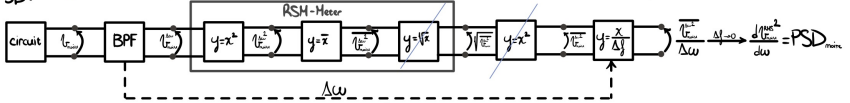
• **Mean** → as we saw, noise has $E[V_{noise}] = \lim_{T \rightarrow \infty} \frac{1}{T} \int_0^T V_{noise}(t) dt = 0$, so this measure doesn't provide any useful information about noise intensity.

• **Root Mean Square and Power** → a better way to measure noise intensity could be calculating its power: $P_{noise} = \frac{V_{rms}^2}{R}$.
 We can measure the noise power using a RMS-meter.



for noise it is usually used the RMS voltage: $V_{rms} = \sqrt{\frac{1}{T} \int_0^T V_{noise}^2(t) dt}$

• **Noise Power Spectral Density** → if we put a BPF in between the circuit and the RMS-meter, at the output of the RMS-meter we obtain a quantity called PSD:



Possiamo quindi ricavare che $PSD_{noise} = \frac{dP_{noise}}{df}$ e di conseguenza $V_{rms}^2 = \int_0^\infty PSD_{noise}(f) df$.

• Now that we know what is the PSD we can define two types of noise: white noise → it has the same intensity at every frequency.
 Electronic devices are characterized by both of them:

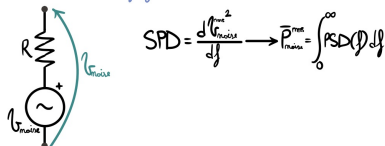
Corner frequency = f_c where $f_{noise}(f_c) = \text{white noise}(f_c)$. The lower is f_c , the better it is for audio applications.

• **Persival Theorem** → this theorem connects the time and the frequency domains, affirming that $P_{noise} = \frac{V_{rms}^2}{R} = \frac{1}{R} \cdot \int_0^\infty PSD_{noise}(f) df$.

• **Noise in resistors** → in resistors, electrons, usually move (even without an applied tension) due to thermal agitation. This random movement generates V_{noise} .
 We can consider this as a white noise having $PSD_{noise} = k_B T R$.

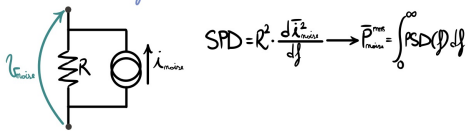
• There are two equivalent models to represent noise in resistors:
 Boltzmann constant = $1.38 \times 10^{-23} \text{ J/K}$

• **Model 1 (voltage generator)**



$$SPD = \frac{dV_{noise}^2}{df} \rightarrow \bar{P}_{noise} = \int_0^\infty PSD(f) df$$

• **Model 2 (current generator)**



$$SPD = R^2 \cdot \frac{dI_{noise}^2}{df} \rightarrow \bar{P}_{noise} = \int_0^\infty PSD(f) df$$

• **Noise in diodes** → a diode shows characteristics of an electronic device called shot noise. Tale rumore è indipendente dal verso di percorrenza della corrente, ed è generato dal fatto che l'attraversamento della barriera da parte degli elettroni è un processo casuale, per cui ogni passaggio crea una piccola variazione di corrente.

La formula della shot noise è $\frac{dI_{noise}^2}{df} = 2qI$, ed q è un valore costante.
 q = elementary charge = $1.6 \cdot 10^{-19} \text{ C}$

• **Noise in transistors** → there are different types of noises for transistors, depending on their type:

• **BJTs** → in this type of transistors $I_{noise} \approx I_{base}$ with I_{base} the power through the base-emitter junction in order to reach the emitter. This passage generates a shot noise, more likely to what happens with diodes: $\frac{dI_{noise}^2}{df} = 2qI_c$.
diode drop: 0.6-0.7 V

This type of noise does not change with temperature.

• Also the base input current is subject to a noise: $\frac{dI_{noise}^2}{df} = 2qI_{base}$.

• In addition to the previously described white noise, there is also a $1/f$ frequency component, so: $\frac{dI_{noise}^2}{df} = 2qI_c + \frac{A_{1/f, BJT}}{f}$ and $\frac{dI_{noise}^2}{df} = 2qI_{base} + \frac{A_{1/f, BJT}}{f}$.

• **FETs (mosFET and JFET)**

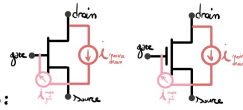
In this case we have a thermal noise similar to the one of simple resistors.

similar to the one in ohmic region:

$$\frac{dI_{noise}^2}{df} = \frac{4kT}{R_{eq}} \approx \frac{4}{g_m}$$

• Also the gate input current is subject to a noise: $\frac{dI_{noise}^2}{df} = 2qI_{gate}$.

• In addition to the previously described white noise, there is also a $1/f$ frequency component, so: $\frac{dI_{noise}^2}{df} = \frac{4kT}{R_{eq}} + \frac{A_{1/f, FET}}{f}$ and $\frac{dI_{noise}^2}{df} = 2qI_{gate} + \frac{A_{1/f, FET}}{f}$.
very little compared to BJT



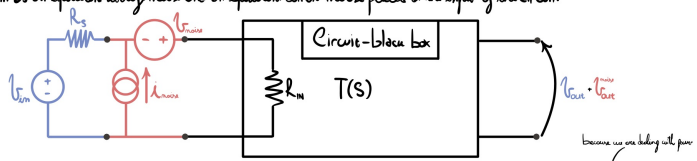
• Often, to compare the noise sources at the input of the circuit, we convert the circuit noise in a tension equivalent:

$$i_{n, \text{noise}} = g_m \cdot v_{n, \text{noise}} \Rightarrow v_{n, \text{noise}} = \frac{i_{n, \text{noise}}}{g_m} = \frac{\frac{2qI_c}{g_m} \cdot \frac{1}{f}}{g_m} = \frac{2qI_c}{g_m^2} = \frac{4kT}{g_m}$$

• Often, to compare the noise sources at the input of the circuit, we convert the circuit noise in a tension equivalent:

$$\frac{d v_{n, \text{noise}}}{df} = \frac{4kT}{g_m}$$

• **Noise in complex linear circuits** → if we think to a linear circuit as a black box containing many components that generate noise, we can lump together the noise and represent them as an equivalent voltage noise and an equivalent current noise placed at the input of the circuit.



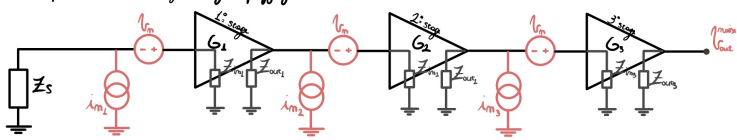
The signal is "processed" through the formula $v_{out} = v_{in} \cdot T(s)$, while the noise follows the equation $\frac{d v_{n, \text{noise}}}{df} = \frac{d v_{n, \text{noise}}}{df} \cdot |T(s)|^2$.

We should also take in account the loss resistances R_s and R_n . Let's apply the superposition effect:

- Considering $v_{in} \rightarrow$ $v_{in} \rightarrow \frac{v_{in}}{R_s} \rightarrow \frac{v_{in}}{R_s} \cdot \frac{R_n}{R_s + R_n} \rightarrow v_{n, \text{noise}} = v_{in} \cdot \frac{R_n}{R_s + R_n} \cdot T(s) \rightarrow \frac{d v_{n, \text{noise}}}{df} = \frac{d v_{n, \text{noise}}}{df} \cdot \left| \frac{R_n T(s)}{R_s + R_n} \right|^2$
- Considering $i_{n, \text{noise}} \rightarrow R_s \rightarrow i_{n, \text{noise}} \rightarrow i_{n, \text{noise}} \cdot R_n \parallel R_s \rightarrow v_{n, \text{noise}} = i_{n, \text{noise}} \cdot R_n \parallel R_s \cdot T(s) \rightarrow \frac{d v_{n, \text{noise}}}{df} = \frac{d i_{n, \text{noise}}}{df} \cdot \left| R_s \parallel R_n \cdot T(s) \right|^2$

So the combined equation is: $\frac{d v_{n, \text{noise}}}{df} = \frac{d v_{n, \text{noise}}}{df} \cdot \left| \frac{R_n T(s)}{R_s + R_n} \right|^2 + \frac{d i_{n, \text{noise}}}{df} \cdot \left| R_s \parallel R_n \cdot T(s) \right|^2$

• **Noise in multi-stage amplifiers** → in order to understand which is the best configuration that helps to minimize the impact of noise to the output, let's make an example. Let's consider the following amplifying chain:



Let's calculate the noise contributions for each stage:

$$v_{n, \text{out}} = \left(i_{n1} \cdot Z_{in} \parallel Z_s \cdot \left(i_{n1} \cdot \frac{Z_s}{Z_{in} \parallel Z_s} \right) \cdot G_1 G_2 G_3 + \left(i_{n2} \cdot Z_{in1} \parallel Z_s + i_{n2} \cdot \frac{Z_s}{Z_{in1} \parallel Z_s} \right) \cdot G_2 G_3 + \left(i_{n3} \cdot Z_{in2} \parallel Z_{out1} + i_{n3} \cdot \frac{Z_s}{Z_{in2} \parallel Z_{out1}} \right) \cdot G_3 \right)$$

in terms of PSD...

$$\frac{d v_{n, \text{out}}^2}{df} = \left(\left(\frac{d i_{n1}}{df} \cdot \left| Z_{in} \parallel Z_s \right|^2 + \left(i_{n1} \cdot \left| \frac{Z_s}{Z_{in} \parallel Z_s} \right|^2 \right) \right) \cdot |G_1 G_2 G_3|^2 + \left(\frac{d i_{n2}}{df} \cdot \left| Z_{in1} \parallel Z_s \right|^2 + \left(\frac{d i_{n2}}{df} \cdot \left| \frac{Z_s}{Z_{in1} \parallel Z_{out1}} \right|^2 \right) \right) \cdot |G_2 G_3|^2 + \left(\frac{d i_{n3}}{df} \cdot \left| Z_{in2} \parallel Z_{out1} \right|^2 + \left(\frac{d i_{n3}}{df} \cdot \left| \frac{Z_s}{Z_{in2} \parallel Z_{out1}} \right|^2 \right) \right) \cdot |G_3|^2 \right)$$

$$= PSD_1 \cdot |G_1 G_2 G_3|^2 + PSD_2 \cdot |G_2 G_3|^2 + PSD_3 \cdot |G_3|^2$$

So, as we can see, the noise at the input of the first stage is the one that impact more on the system output because it is amplified at each stage of the circuit. This means that an intelligent approach to build a chain of amplifiers is to:

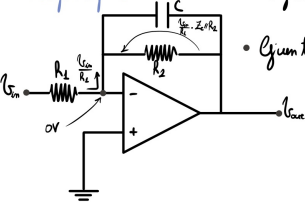
- minimize the 1st stage noise (e.g. using feedback opamp)
- maximize the 1st stage gain G_1 in order to have on higher SNR.

There are two main categories of filters:
 • **Passive filters** → made by combination of passive components such as resistors, capacitors and inductors
 • **Active filters** → contain also active components such as operational amplifiers
 Let's analyze in detail various types of active filters.

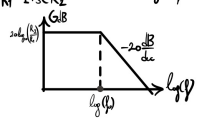
Active LPF

• **1st order LPF** → let's see how we can build a 1st order LPF.

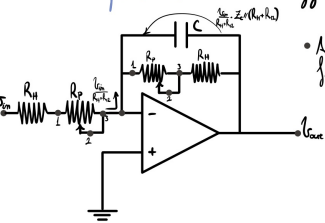
Simple implementation



• As for this thing, let's calculate the transfer function of the circuit: $v_{out} = v_{in} \cdot \left(\frac{Z_2 + R_2}{R_1} \right) = v_{in} \cdot \left(\frac{R_2 + \frac{1}{sC}}{R_1} \right) = \frac{R_2 + R_1}{1 + sCR_1} = TF(s)$
 • Given that $TF(s) = \frac{R_2}{R_1} \cdot \frac{1}{1 + sCR_1}$ we have only a pole in $f_p = \frac{1}{2\pi CR_1}$, so the frequency response will be:



• **Parametric implementation** → we can modify the position of the pole by applying the following changes to the circuit:

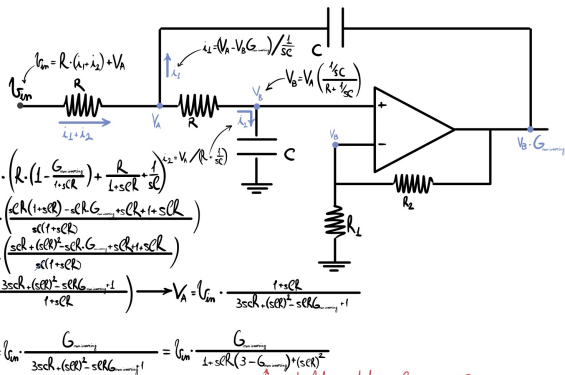


• As we can notice, in this case the TF(s) = $\frac{1}{R_1 + R_2} \cdot \frac{R_2 + R_1}{sC(R_1 + R_2) + 1} = \frac{1}{sC(R_1 + R_2) + 1}$, with a $f_p = \frac{1}{2\pi sC(R_1 + R_2)}$

2nd order LPF: Sallen Key LPF - Biquadratic LPF

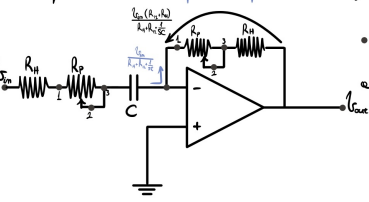
• Let's see the equations that we obtained on the circuit to get its TF.

$$\begin{aligned}
 & i_1 = v_{in} / R_1 = v_A / R_1 + v_A / R_2 \\
 & v_{in} = R_1 \left(\frac{v_A}{R_1} + \frac{v_A}{R_2} \right) = v_A \left(\frac{R_1 + R_2}{R_1 R_2} \right) \cdot sC + v_A \left(\frac{1}{1 + sCR_2} \right) \\
 & v_{in} = v_A \left(\frac{R_1 + R_2}{R_1 R_2} \right) \cdot sC + v_A \left(\frac{1}{1 + sCR_2} \right) \\
 & v_{out} = v_A \cdot \frac{1}{1 + sCR_2} = v_A \cdot \frac{1}{1 + sCR_2} \\
 & v_{out} = v_{in} \cdot \frac{1}{1 + sCR_2} = v_{in} \cdot \frac{1}{1 + sCR_2}
 \end{aligned}$$

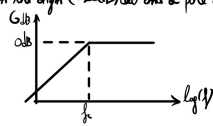


Active HPF

• **1st order HPF** → we can build on HPF using a very similar configuration to the LPF, but this time placing the capacitor in series with R2. Also in this case is possible to build a simple and a parametric version of the HPF, so I will briefly describe the parametric one.



• In this case, the TF(s) = $\frac{R_2 + R_1}{R_1 R_2} \cdot \frac{1}{sC} = \frac{1}{sC} \cdot \frac{R_2 + R_1}{R_1 R_2} = \frac{1}{sC} \cdot \frac{R_2 + R_1}{R_1 R_2}$, so we have a zero in the origin (→ 20dB/dec) and a pole at $f_p = \frac{1}{2\pi C(R_1 + R_2)}$. Obviously, also in this case $f_p = \frac{1}{2\pi C(R_1 + R_2)}$.



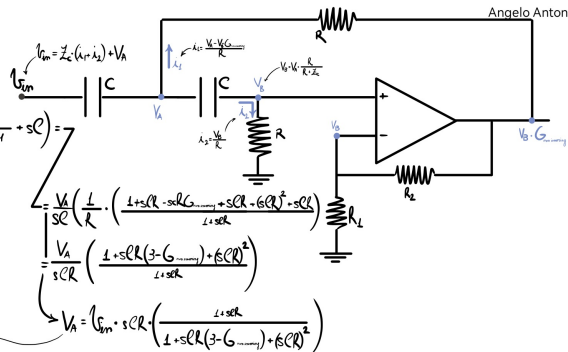
Stability condition: G_m inverting < 3

2° order HPF: Sallen Key HPF - Biquadratic HPF

This configuration is identical to the correspondent LP one with the C and R swapped. Let's calculate the TF(s):

$$V_{in} = Z_C \cdot (i_1 + i_2) + V_1 = \frac{V_1}{sC} \left(\frac{1}{R} \cdot \left(\frac{1+sLR-sCR}{1+sCR} \right) + \frac{sC}{sCR+1} \right) = \frac{V_1}{sC} \left(\frac{1+sLR-sCR}{1+sCR} + \frac{sCR}{1+sCR} \right)$$

$$V_1 = V_{in} \cdot sCR \cdot \left(\frac{1+sCR}{1+sCR(3-G_{min})+s^2CR^2} \right)$$



We can finally calculate: $V_{out} = V_1 \cdot G_{min} = G_{min} \cdot V_1 \cdot \frac{R}{R+Z_C} = G_{min} \cdot \frac{sCR}{1+sCR} \cdot V_{in} \cdot sCR \cdot \left(\frac{1+sCR}{1+sCR(3-G_{min})+s^2CR^2} \right) = V_{in} \cdot \frac{G_{min} \cdot (sCR)^2}{1+sCR(3-G_{min})+s^2CR^2}$

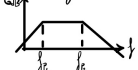
Active BPF



We can realize in this circuit: $i = \frac{V_{in}}{R_1 + Z_{C1}} \rightarrow i \cdot R_2 \parallel Z_{C2} = \frac{V_{in}}{R_1 + Z_{C1}} \cdot R_2 \parallel Z_{C2}$
 where $R_1 + Z_{C1} = R_1 + \frac{1}{sC_1} = \frac{sCR_1 + 1}{sC_1}$ and $R_2 \parallel Z_{C2} = \left(\frac{1}{R_2} + sC_2 \right)^{-1} = \frac{R_2}{1+sCR_2}$
 Merging the results together we get: $i \cdot R_2 \parallel Z_{C2} = V_{in} \cdot \frac{sC_1 R_2}{(1+sCR_1)(1+sCR_2)}$
 $V_{out} = V_{in} \cdot \frac{sC_1 R_2}{(1+sCR_1)(1+sCR_2)}$

From the equation we notice that there are one zero in the origin and two poles at $f_{p1} = \frac{1}{2\pi C_1 R_1}$ and $f_{p2} = \frac{1}{2\pi C_2 R_2}$.

Setting the right values of C_1 and R_1 we obtain a frequency response like this:



We can also notice that this is a combination of the 1° order HP and LP filters.

2° order BPF: Sallen Key BPF - Biquadratic BPF

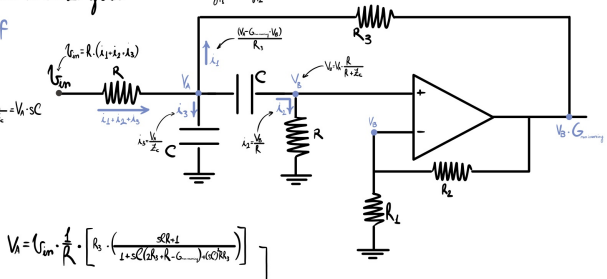
Let's calculate the TF:

$$i_1 = \frac{V_{in} - V_1}{R_1} = \frac{V_1}{R_1} \left(1 - \frac{sCR_1}{1+sCR_1} \right) \rightarrow i_2 = \frac{V_1}{R_2} \cdot \frac{1}{1+sCR_2} = V_1 \cdot \frac{sC}{1+sCR_2}$$

$$V_{in} = R_1 \cdot (i_1 + i_2) = R_1 V_1 \left[\frac{1}{R_1} \left(1 - \frac{sCR_1}{1+sCR_1} \right) + \frac{sC}{1+sCR_2} \right]$$

$$= R_1 V_1 \left[\frac{1}{R_1} \left(\frac{1+sCR_1}{1+sCR_1} - \frac{sCR_1}{1+sCR_1} \right) + \frac{sC}{1+sCR_2} \right]$$

$$= R_1 V_1 \left[\frac{1}{R_1} \left(\frac{1+sC(R_1 R_2 - R_1 G_{min})}{1+sC(R_1 R_2 - R_1 G_{min})} \right) \right]$$



$$V_1 = V_{in} \cdot \frac{1}{R_1} \cdot \left[R_2 \cdot \left(\frac{sCR_1 - 1}{1+sC(R_1 R_2 - R_1 G_{min})} \right) \right]$$

$$V_{out} = V_1 \cdot G_{min} = V_{in} \cdot \frac{sCR_1}{1+sCR_1} \cdot G_{min} \cdot \frac{1}{R_1} \cdot \left[R_2 \cdot \left(\frac{sCR_1 - 1}{1+sC(R_1 R_2 - R_1 G_{min})} \right) \right] = V_{in} \cdot \frac{R_2}{R_1} \cdot \left(\frac{sCR_1 G_{min}}{1+sC(R_1 R_2 - R_1 G_{min})} \right)$$

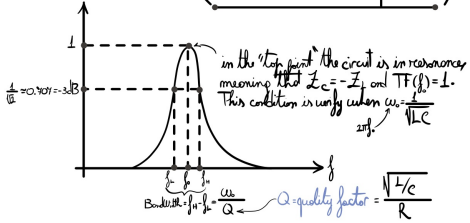
Passive resonant filter → it is a BPF with a very narrow band. Let's see how we can implement it:

The transfer function is just a voltage divider

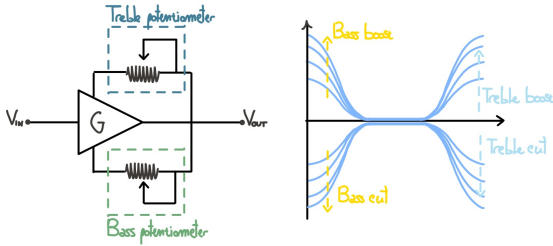
$$TF(s) = \frac{R}{R + Z_C + Z_L} = \frac{R}{R + \frac{1}{sC} + sL}$$

We can see that if $f \rightarrow \infty$ we have $\frac{R}{R + \frac{1}{sC} + sL} \rightarrow 0$

and if $f \rightarrow 0$ then $\frac{R}{R + \frac{1}{sC} + sL} \rightarrow 0$, so the frequency response will be something like:



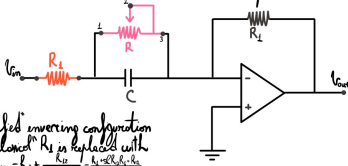
• **Tone control** → the aim of a tone-control filter is to allow to set the gain of the bass/treble frequencies all in one filter. The high level working mechanism is sketched below:



We will build it step-by-step.

• **Tone control - Bass control** → we want to be able to increase or decrease the low frequencies. Let's first see how to implement these two functions separately, so that it will then be easier to build a "complete" circuit that includes both the functions.

• Bass cut

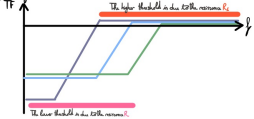


• It is a "modified" inverting configuration in which the "closed" R_1 is replaced with $R_{eq} = R_1 + Z_C \parallel R_{12} = R_1 + \frac{R_{12}}{sCR_1 + 1} = \frac{R_1(sCR_1 + 1) + R_{12}}{sCR_1 + 1}$.

$$\text{Thus means that } TF(s) = -R_L / (R_1 + Z_C \parallel R_{12}) = -\frac{R_L(sCR_1 + 1)}{R_1(sCR_1 + 1) + R_{12}} = -\frac{R_L(sCR_1 + 1)}{R_1 + R_{12} + sCR_1 R_{12}}$$

(one conf. $\frac{R_L}{R_1}$)

• The frequency response is:



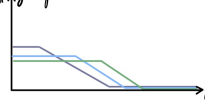
• Bass boost

• Also in this case we have a "modified" inverting configuration with the "closed" R_L replaced with $R_{eq} = R_L \parallel Z_C + R_1$.

• Let's calculate the TF(s):

$$TF(s) = -\frac{R_L \parallel Z_C + R_1}{R_1 + R_{12}} = -\frac{\frac{R_L}{sC(R_L + R_1)} + R_1}{R_1 + R_{12}} = -\frac{R_L + sC(R_L + R_1)R_1}{R_1(1 + sCR_1)} = -\frac{R_L + sC(R_L R_1 + R_1^2)}{R_1(1 + sCR_1)}$$

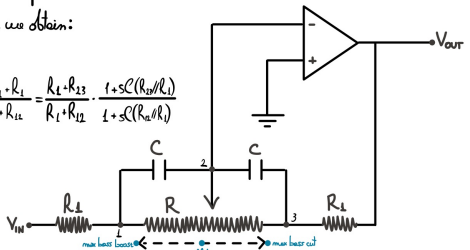
• The frequency response is:



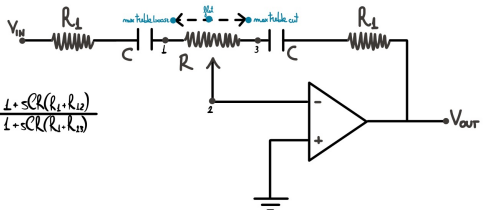
• **Complete circuit of bass control:** merging the previous two circuits we obtain:

$$\text{• } TF(s) = -\frac{Z_C \parallel R_{12} + R_1}{R_{12} + Z_C \parallel R_1} = -\frac{\frac{R_{12}}{sC(R_{12} + R_1)} + R_1}{R_{12} + \frac{R_1}{sC(R_{12} + R_1)}} = -\frac{R_{12} + sC(R_{12} R_1 + R_1^2)}{R_{12} + R_1 + sC R_{12} R_1} = -\frac{R_{12} + sC(R_{12} R_1 + R_1^2)}{R_1 + R_{12} + sC R_{12} R_1}$$

(one conf. $\frac{R_L}{R_1}$)



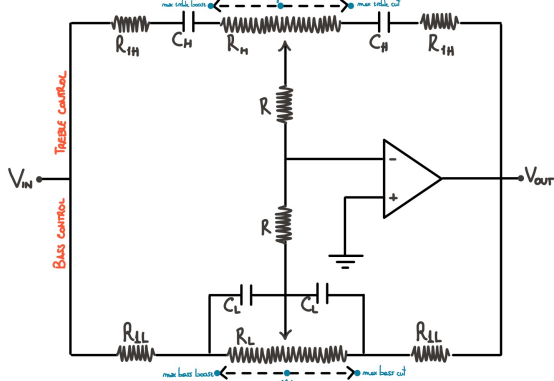
• **Tone control - Treble control** → we want to be able to increase or decrease the high frequencies. The process to obtain that is very similar to the bass control one, we just analyze the complete treble control circuit:



$$\text{• } TF(s) = -\frac{R_1 + Z_C \parallel R_{12}}{R_{12} + Z_C \parallel R_1} = -\frac{R_1 + \frac{R_{12}}{sC(R_{12} + R_1)}}{R_{12} + \frac{R_1}{sC(R_{12} + R_1)}} = -\frac{R_1 + sC(R_{12} R_1 + R_1^2)}{R_{12} + R_1 + sC R_{12} R_1}$$

(one conf. $\frac{R_L}{R_1}$)

- **Full tone control circuit - Baxandall tone control** → we can finally merge the bass and treble control circuit to obtain the following:

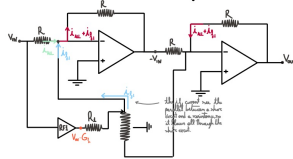
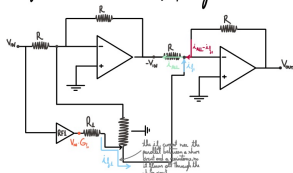
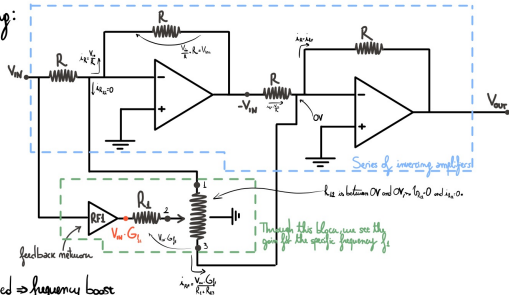


- **Graphic equalizers** → in a graphic equalizer we use different gain control to set the level of different specific frequencies. Let's see how to implement it.
- The basic circuit for the equalization of a single frequency is the following:

$$V_{out} = -R(i_R + i_{R2}) = R \left(\frac{V_{in}}{R} - \frac{V_{in} \cdot G_1}{R_1 + R_{23}} \right) = V_{in} \left(1 - \frac{R G_1}{R_1 + R_{23}} \right)$$

if R_{23} is minimized → frequency cut

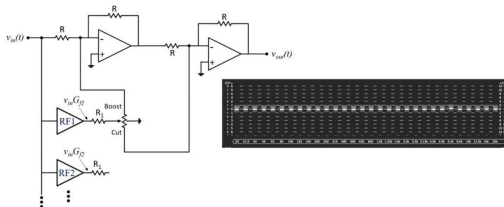
if R_{23} is maximized → frequency boost



$$V_{out} = R \cdot (i_{R1} + i_{R2}) = R \left(\frac{V_{in}}{R} - \frac{V_{in} \cdot G_1}{R_1} \right) = V_{in} (1 - \beta G_1)$$

$$V_{out} = R \cdot (i_{R1} + i_{R2}) = R \left(\frac{V_{in}}{R} - \frac{V_{in} \cdot G_1}{R_1} \right) = V_{in} (1 - \beta G_1)$$

- To build a complete equalizer that allows to manage different frequencies, different RF modules are placed in parallel, as showed below:



- **Bi-quadratic filters** → it is a particular type of filter having the transfer function in the form $TF(s) = \frac{a_2 s^2 + a_1 s + a_0}{s^2 + b_1 s + b_2}$. For example:

$$TF_{LP}(s) = \frac{K}{s^2 + (\frac{1}{Q})(\frac{s}{\omega_0}) + 1} \quad TF_{BP}(s) = \frac{KS}{s^2 + (\frac{1}{Q})(\frac{s}{\omega_0}) + 1} \quad TF_{HP}(s) = \frac{KS^2}{s^2 + (\frac{1}{Q})(\frac{s}{\omega_0}) + 1} \quad TF_{LP}(s) = \frac{K(s - \omega_0)}{s^2 + (\frac{1}{Q})(\frac{s}{\omega_0}) + 1}$$

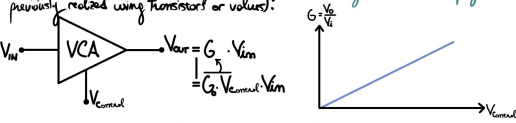
As we can see all the filters have the same denominator, while the numerator presents 0, 1, 2 zeros, as required by the filter's frequency response.

Automatic Gain Control

AGCs are a family of components that allow to automatically change their gain in order to reach a certain aim. In this section we will see which are the basic constitutive elements of an AGC and which are the most used AGCs.

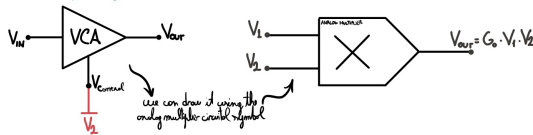
● **How to build an RMS meter** → we know that, given a tension V , the correspondent I_{RMS} is given by the formula $I_{RMS} = \sqrt{\frac{1}{T} \int_0^T i^2 dt}$. This means that, in order to obtain I_{RMS} we need the following three circuits:

● **Squarer circuit**: let's start from a common voltage controlled amplifier (that we previously realized using transistors or valves):

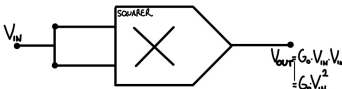


The gain $G = G_0 \cdot V_{control}$ changes linearly according to $V_{control}$.

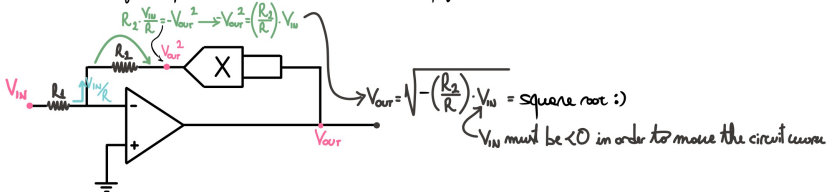
● If then we connect the $V_{control}$ pin with "another voltage" of the circuit, we are building an analog voltage multiplier.



● At this point it's easy to get a squarer: we just need to connect the only multiplier inputs:



● **Square root circuit**: using the square it's possible to build a circuit that performs the square root:

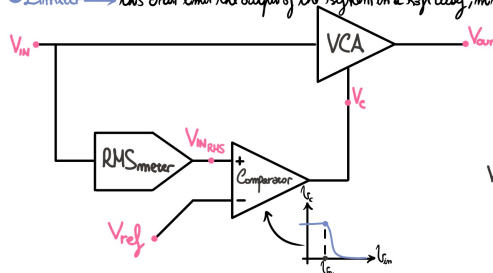


● **RMS meter**: finally we can combine the previous elements to build an RMS meter:

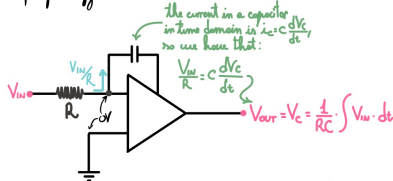


● **AGC different circuits** → using the RMS meter it's possible to build various very useful devices:

● **Limiter** → this circuit limit the output of the system in a right way, minimizing the distortion:



● **Integrator**: it is implemented using the following feedback opamp configuration:



The output of the system is $-V_{out} = Z_c \cdot i_c = Z_c \cdot \frac{V_{in}}{R}$.

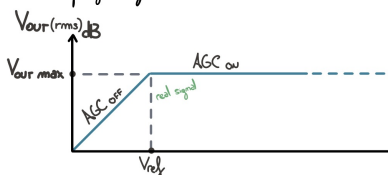
Instead of working in the S domain, let's work in the time domain. Here, the equation that describe the behaviour of a capacitor is:

$$i_c = C \cdot \frac{dV_c}{dt} \text{ where (in this case) } i_c = \frac{V_{in}}{R}$$

Let's extrapolate $V_{out} = V_c$ from the previous equation:

$$\frac{V_{in}}{R} = C \cdot \frac{dV_c}{dt} \rightarrow \frac{V_{in}}{RC} = \frac{dV_c}{dt} \rightarrow V_c = \frac{1}{RC} \int V_{in} dt$$

So, the output of the system is $V_{out} = -\frac{1}{RC} \int V_{in} dt$, making the circuit an analog voltage integrator.

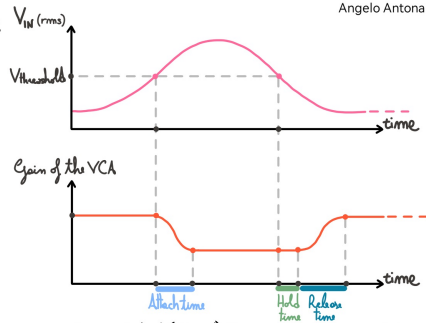


We can also analyse how the gain of the VCA (and so V_c) evolves in function of V_{in} :

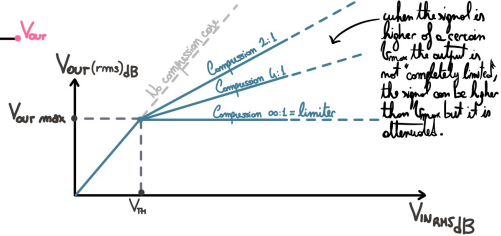
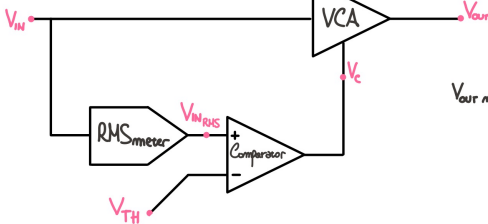
● When V_{in} overcomes the V_{TH} , the circuit takes a while (*Attack*) to adjust the gain.

● When V_{in} returns lower than V_{TH} , the circuit takes a while to start increasing the gain (*Hold*) and then it needs other time to reach the correct gain level (*Release*).

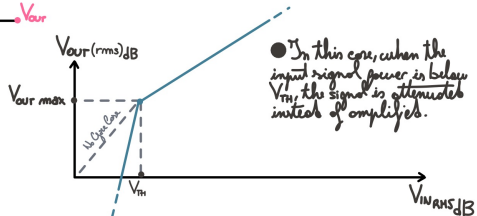
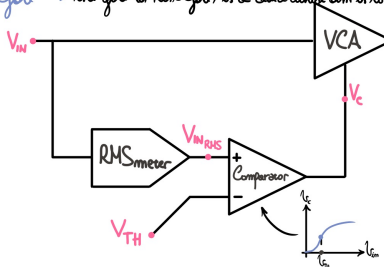
These "delays" in the activation and deactivation of the gain control makes this circuit a limiter and not a clipper.



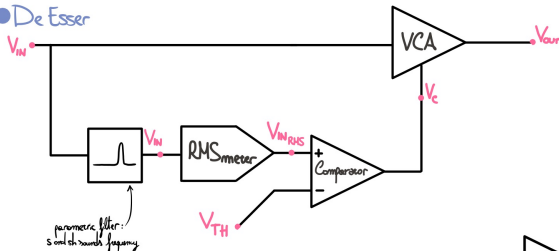
● **Compressor** → it is a "more delicate limiter": it doesn't limit the signal below a certain V_{TH} , instead it "attenuates" the signal with a certain factor $\alpha:1$. Notice that a limiter can be seen as a compressor with ratio of $\infty:1$.



● **Gate** (or **Non Gate**) is a device whose aim is to "remove" the noise from the silent sections of the signal. Let's see how it is implemented:

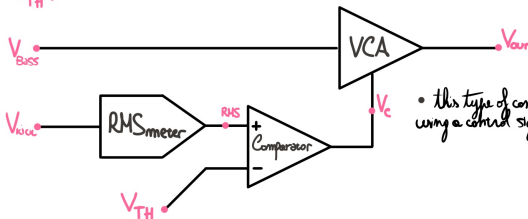


● **De Esser**



• its aim is to attenuate sibilances, often in the 5kHz-8kHz frequency range.

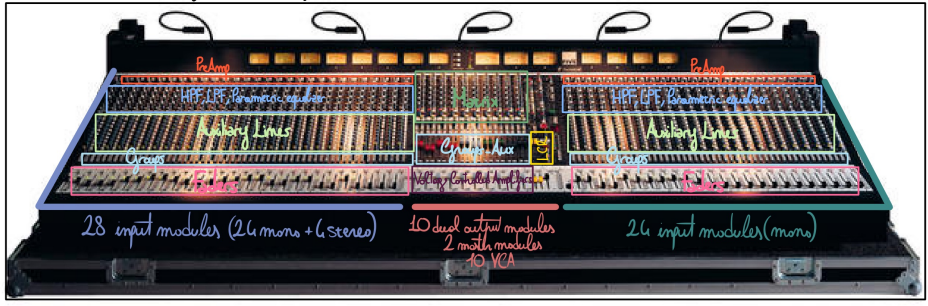
● **Side-chain compressor**



• this type of compressor acts on a target signal using a control signal, called "side chain".

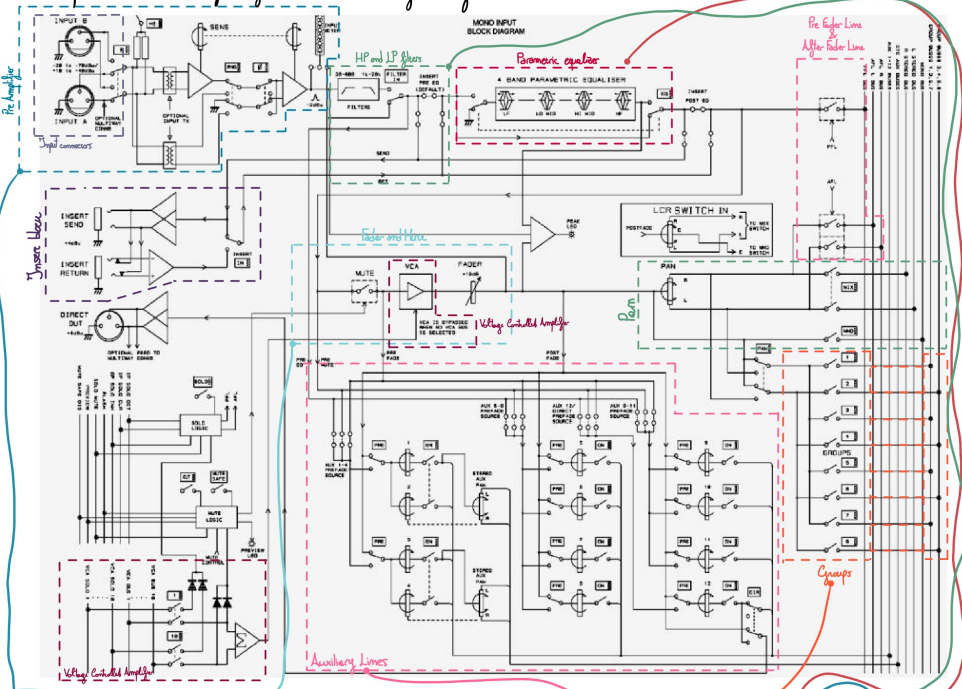
Mixing consoles

In this section we will analyze the Soundcraft Series V. This console is divided in several sections:

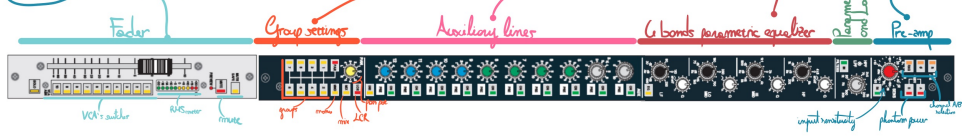


We will analyze in detail each one of these sections.

● **Input section** → as we can see in the picture of the console, the input section is composed by many repeated mono/stereo channels. The complete circuit model of a single (mono) channel is the following:



Each section in the circuit represent a part of the input line on the real console:



Let's analyse the input channel studying each one of the elements that compose it.

● Pre-amplification section

Phantom power: this switch allows to provide a +68V (nominal) to devices that need it. The current passes through a couple of resistors towards the A and B inputs.

This potentiometer regulates the sensitivity of the pre-amp gain.

Range: it changes the excursion of the input:

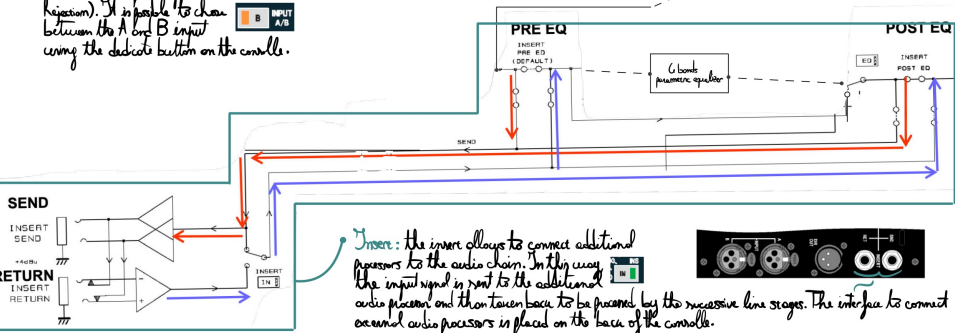
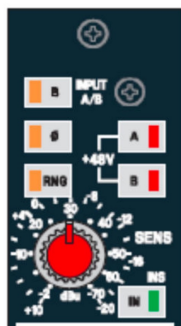
- mic $\rightarrow$ microphones has a very "weak" signal so they need a higher pre-amplification.
- line $\rightarrow$ other instruments provide "stronger" signals to the system so it's necessary to lower the pre-amplification (-20dB).

Phase inversion: it allows to shift the signal's phase by 180°.

the inputs A and B allow to take in input a differential signal, allowing to perform noise cancelling exploiting all the differential techniques we saw in the first chapter (Common Mode Rejection). It is possible to choose between the A and B input using the dedicated button on the console.

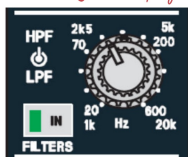
the "optional" input TX: are a couple of transformers used to adapt the input impedance according to the input device.

Parametric High and Low pass



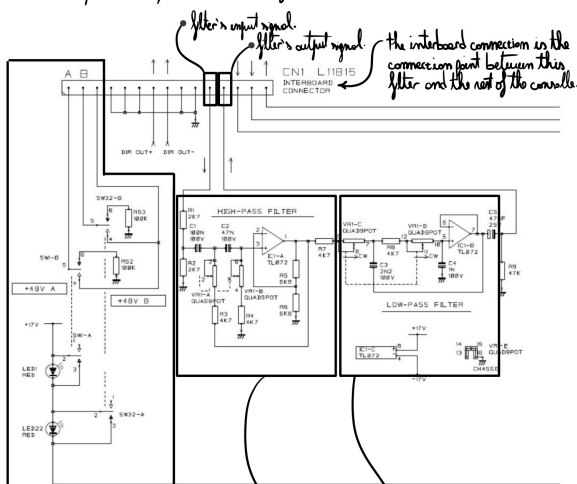
Insert: the insert allows to connect additional processors to the audio chain. In this way the input signal is sent to the additional audio processor and then taken back to be processed by the successive line stages. The interface to connect external audio processors is placed on the back of the console.

● Parametric High and Low pass filters



Let's analyze the high level schematic of this section of the console:

Let's analyze the "low level" schematic of the filter:



This switch turns on/off the filter:

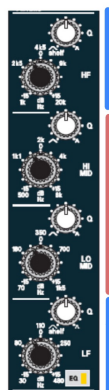
- if it is connected to the "upper pin" the signal that flows in the next stages is the output of the filter.
- if it is connected to the "lower pin" the signal that flows in the next stages is the input of the filter.

this is actually the phantom power section that we already seen in the pre-amp section.

this is the HPF module.

this is the LPF module.

4 band Parametric equaliser



HF and LF have the same controls:

- this knob allows to set the **quality factor** of the filter. If we set the knob 'completely on the left' the filter has a 'shallow slope' otherwise it has a 'bell shape'.

this knob contains **two** controls:

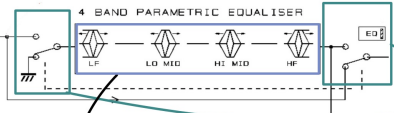
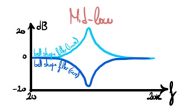
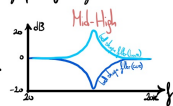
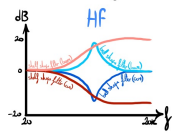
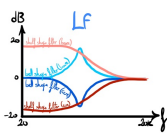
- the external one allows to select the **central frequency** of the bell shape filter.
- the internal one is used to set the **gain boost or cut** to the selected frequency range.

Hi-mid and Lo-mid filters have the same controls:

- similarly to hi and lo filter, this is the **Q-factor** select knob.
- In this case, the filter is always in 'bell shape mode'.

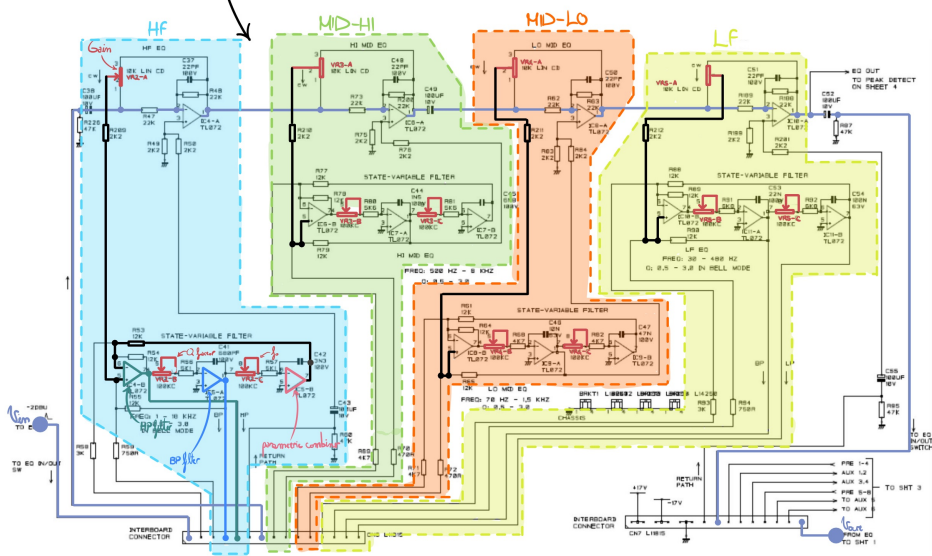
this knob contains **two** controls:

- the external one allows to select the **central frequency** of the bell shape filter.
- the internal one is used to set the **gain boost or cut** to the selected frequency range.

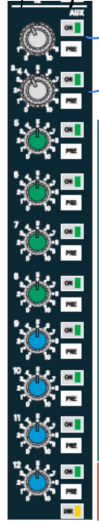


This switch allows to activate/de-activate the parametric filter module.

Let's see the low level circuit schematic.



● **Auxiliary lines** → they are additional channel exits that allow to duplicate the signal and send it to a secondary destination (ex stage monitor) that is separated from the main mix.



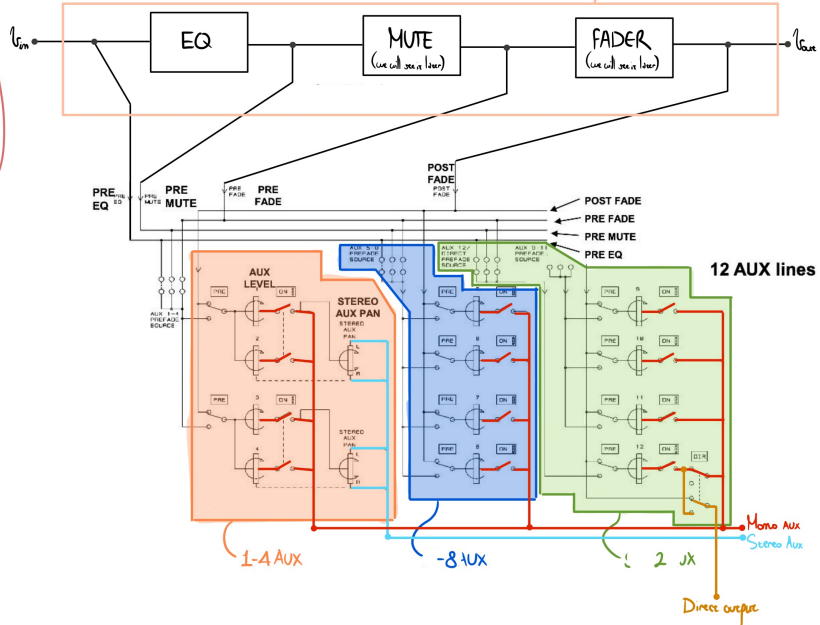
The first 6 AUX are stereo. Here we can not:

- ◀ double potentiometer → external source not the pre
- ◀ the On/off button allows to activate or deactivate the aux signal
- ◀ the Pre button allows to chose if the signal has to be the pre or the post fader signal.

The 5-11 AUX are mono exits. (In this case we find the pre and the on/off button)

The 12 aux is similar to the previous ones but in this case this is also a **direct button** which allows to send the signal to a direct exit, meaning a dedicated output of the back of the console.

● Main audio chain.



Mute and Fader

② This is a **absence module**:

- R130 and C36 form a LPF, avoiding "burrings" of the signal due to the commutation of the switch ($\tau > 1\text{ms}$) when we PRESS MUTE.
- DS deals with the possibility that, for some reason, the TR8 turns on its own, preventing limiting the transistor to OFF.
- R127 allows the C36 capacitor discharge when we RELEASE MUTE.

Also in this case $\tau > 1\text{ms}$.

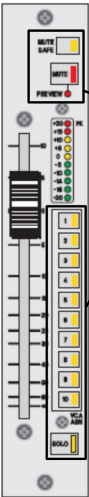
• TR8: when the SW5-A and SW5-B switches are commutated to "ground", the $V_{BE} > V_{BE1}$ and the transistor enters SATURATION MODE. In this way it increases the tension on the base of the TR10 transistor.

③ When TR8 "turns on", the tension on the base of TR10 is lowered and $V_{BE} < V_{BE1}$ turning off the TR10 transistor.

④ When TR10 turns on, the current generated from this module is directed to ground and doesn't feed the gate of TR7.

① When we press the "mute" button, we are mechanically these transistors (SW5-A and SW5-B).

Effect section

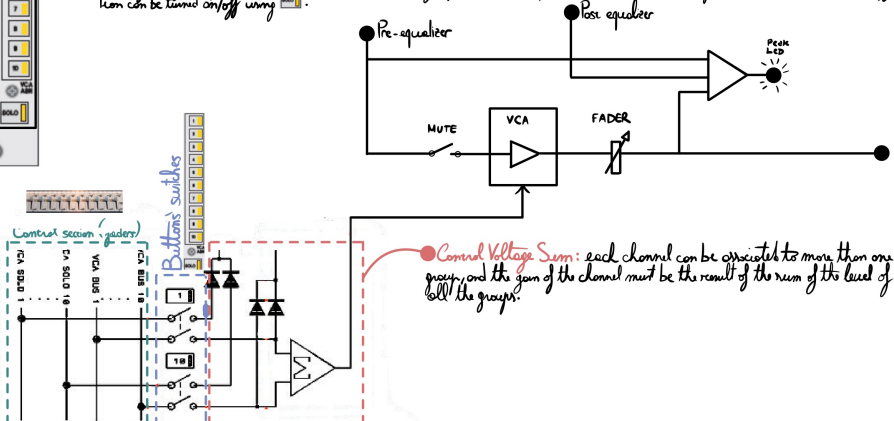


⑤ When the TR7 is "turned on" the input signal passes to the buffer TL07250 and goes to the output. When TR10 is in saturation $\Rightarrow$ TR7 becomes non conductive and the signal goes to "ground".

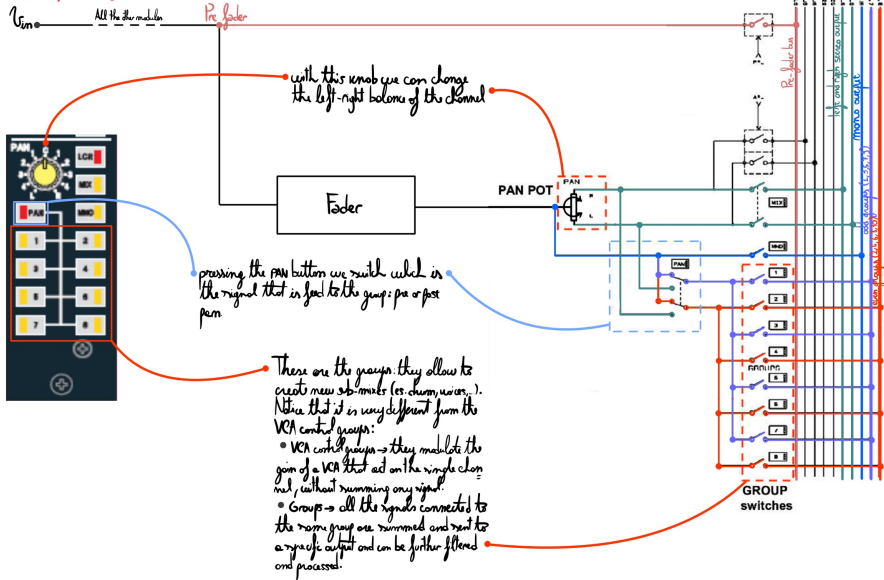
⑥ This is the buffer section. The transistor TR12 commutation is managed by the same voltage of TR10 so when TR10 is on the "mute mode" = saturation, also TR12 is turned on and being to ground possible signal leakage (even that TR10 is in non conductive mode).

● This button  is used to mute or commute the channel. The schematic above describes the internal circuit of the mute implementation. Below we can see how it is implemented together with the other elements of this channel section.

● Each one of these buttons allows to associate the channel to one of the 10 VCA groups. These channels are useful: for each one of these groups there is a fader in the "main section" of the console . If we change the position of one of these faders, we modify the gain of all the channel associated to that group. Below it is possible to see how this is implemented. Note that this configuration can be turned on/off using .

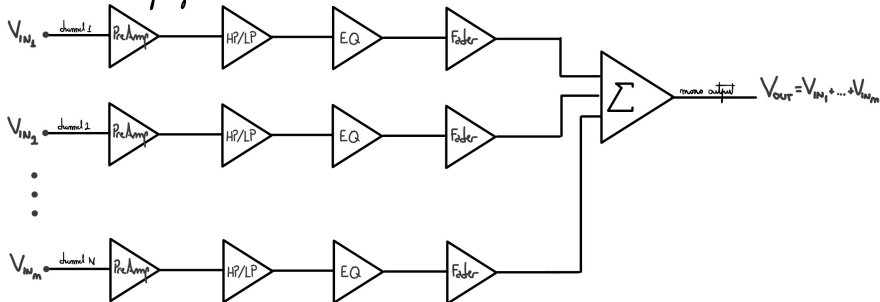


Group settings



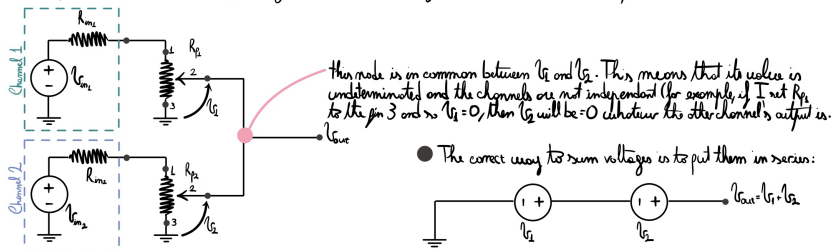
How to correctly sum signals $\rightarrow$ the schematic above shows the low-level implementation of the panning and signal summation functionality of the console. In this sub-section we will see how these modules are built from a low-level point of view.

Let's consider the mono output of the console:

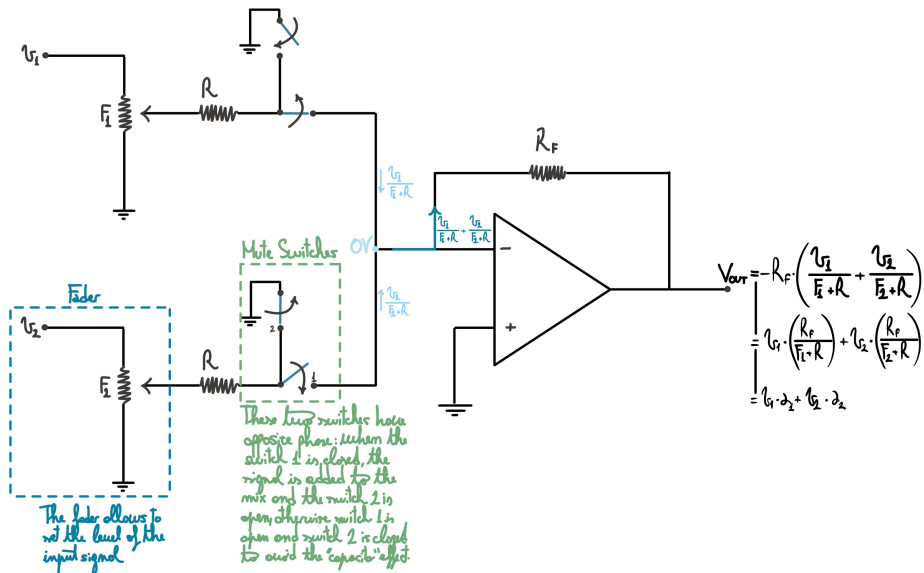


The previous picture shows what we would like to do. But how can we implement it?

A stupid way would be to just join together two or more voltage sources, but this leads to many problems. Let's see it.



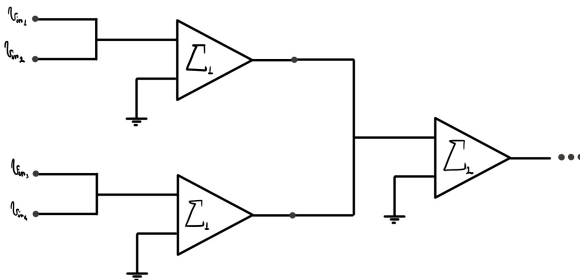
- Actually, in mixing stages we usually sum currents instead of voltages:



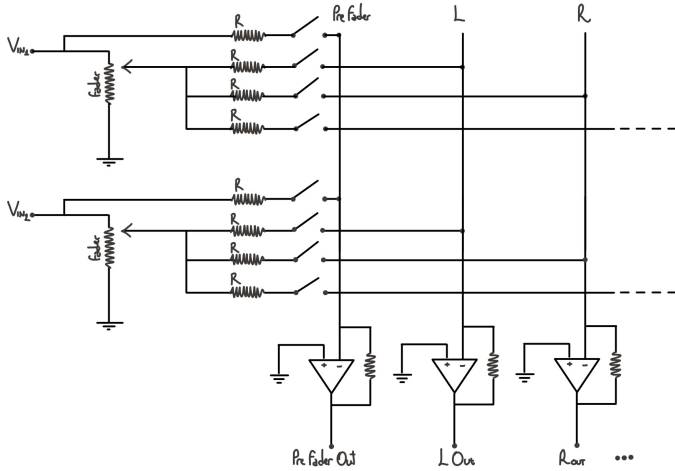
Ideally, we could sum infinite currents using a single operational amplifier, but in reality we have to deal with several limitations:

- Maximum voltage range: V_{out} must be $< V_{DD}$, otherwise the amplifier will saturate.
- Maximum output current: as we saw in the dedicated chapter, an opamp is able to source a limited amount of current.
- Adding more input resistors we are adding more noise and we are decreasing the BW of the opamp, given that $C_{flop} = A(s) \cdot \frac{R_{in}}{R_{out} \cdot R_F}$ and $R_{in} = R_1 // R_2 // \dots // R_n$.

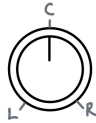
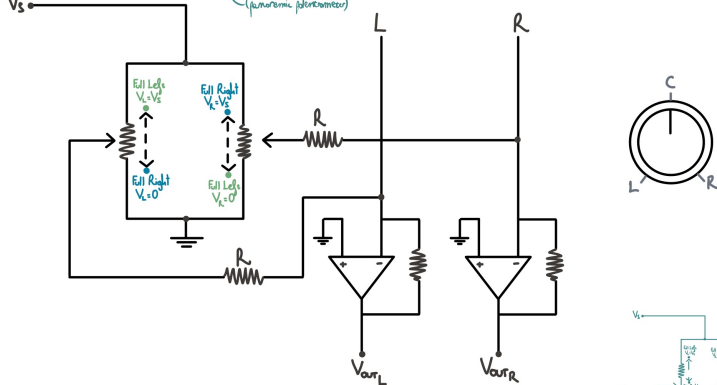
For these reasons a "multi-stage mixing" approach is used:



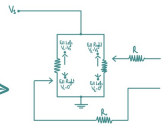
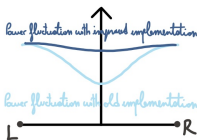
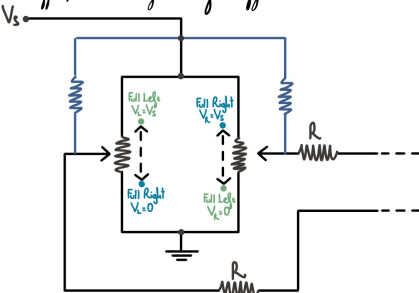
● **Mixing matrix** → the mixed signal are sent to various output buses:



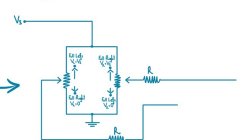
● **Pinning** → until now we've seen the "pin pit" on a high level point of view. Let's see how it is actually implemented:



● The previous configuration has one problem: power. Let's consider two cases:
 - full right case
 - center case
 As you can see there is a huge difference in total output power. To mitigate this effect, we use the following configuration:

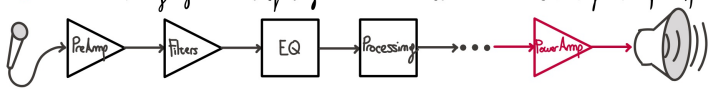


$$P_{L,R} = \frac{V_L^2}{R_L} \Rightarrow P_{out} = \frac{V_L^2}{R_L} \cdot R_L$$



$$V_L = \frac{V_L}{2} \Rightarrow P_{out} = 2 \cdot \frac{V_L^2}{4} \cdot R_L = \frac{V_L^2}{2} \cdot R_L$$

Power Amplifiers → this is the "last stage" of the "electronic part" of the audio chain: its aim is to drive the output loudspeaker, providing enough electrical power.



Power Amplifiers are categorized in different groups according to their efficiency, meaning "how much time" every transistor remain "active" during the waveform. We will talk only about BJT power amplifier.

To correctly describe these classes we must introduce some new concepts:

- **Angle of conduction (θ)** → it is the portion of the sinusoidal period in which a transistor conducts current. It is measured in (0-360) degrees.
- **Rest point (Q-point)** → it is a point on the plane in which the coordinates are: I_{CQ} = collector current in DC: $I_{CQ} = \frac{V_{CC} - V_{CEQ}}{R_C} = 0.1V$
- This point is important because the variable part of the signal oscillates around it, and it's necessary to carefully choose Q-point to keep it into the linear range of the amplifier.
- **Swing** → maximum excursion of the signal that does not cause distortion.
- **Efficiency** → it is the ratio $\eta = \frac{P_{out}}{P_{in}}$.



Now we have the necessary vocabulary and we can analyze the four Amplifier Classes.

• **Class A** → in a class-A power amplifier, the transistor is "always turned on", even when there's no signal ⇒ angle of conduction = 360°.

For this reason, this kind of amplifiers have very high linearity (sound quality) but very low efficiency.

- As we saw, the current and the tension that characterize the c and e pins of the transistor are conserved by:

$$i_{Ctot} = I_{CQ} + i_c(t) \quad \begin{matrix} \text{DC current} \\ \text{variable component} \end{matrix} \quad \begin{matrix} V_{CEtot} = V_{CEQ} + v_{ce}(t) \\ \text{DC voltage} \\ \text{variable component} \end{matrix}$$

- Let's analyze the rest point = Q-point = DC component:

① We know that $i_C = \beta \cdot i_B \Rightarrow i_B = \frac{i_C}{\beta}$

In this type of implementations $\beta \in [10, 300]$ so i_B is very low and can be neglected.

② Given that $Z_C = \infty$, the current i_C flows completely into R_C , so it's very easy to calculate $V_{CE} = V_{CC} - i_C \cdot R_C$.

linear case analyzing the DC case

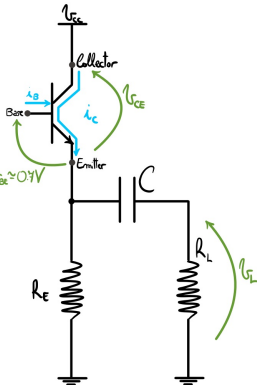
③ For the VCE we can write that $V_{CE} = V_{CEQ} + v_{ce}$, meaning that $V_{CEQ} = V_{CC} - I_{CQ} \cdot R_C = V_{CC} - i_C \cdot R_C$

④ $\omega = 0$, so the capacitor impedance $Z_C = \frac{1}{j\omega C} = \infty \Rightarrow$ the capacitor behaves as an open-circuit.

⑤ from the equation of the point ③ it's easy to get that:

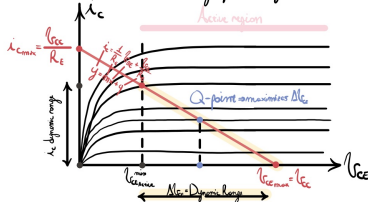
$$V_{CEQ} = V_{CC} - I_{CQ} \cdot R_C \Rightarrow I_{CQ} = \frac{V_{CC} - V_{CEQ}}{R_C}$$

$$I_{CQ} = I_{CQ}$$



- Given that, as we saw in point ③ $V_{CEQ} = V_{CC} - I_{CQ} \cdot R_C$, we often choose $V_{CEQ} = \frac{V_{CC}}{2}$ in order to keep $V_{CEQ} = \frac{V_{CC}}{2}$.
- In this way, the swing = $\pm V_{CEQ}/2$ and $I_{CQ} = V_{CEQ}/R_C = \frac{V_{CC}}{2R_C}$

- Let's see this situation graphically:

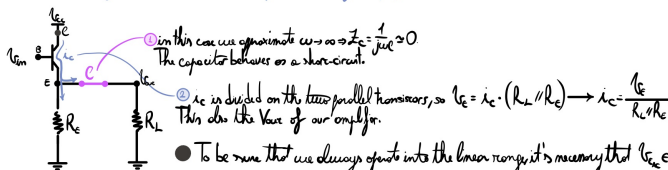


• from point ⑤ we saw that $i_C = \frac{V_{CC} - V_{CE}}{R_C}$, and we can re-write that equation as $i_C = \frac{1}{R_C} (V_{CC} - V_{CE})$

line equation: $y = mx + q$

- This straight line is called "retta di carico".

- Let's analyze the AC behavior of this kind of amplifiers:



① in this case we approximate $\omega \rightarrow \infty \Rightarrow Z_C = \frac{1}{j\omega C} \approx 0$
The capacitor behaves as a short-circuit.

② i_C is divided on the two parallel transistors, so $V_{CE} = i_C \cdot (R_C \parallel R_L) \rightarrow i_C = \frac{V_{CE}}{R_C \parallel R_L}$
This is the Vce of our amplifier.

• To be sure that we always operate into the linear range, it's necessary that $V_{CE} \in \left[-\frac{V_{CEQ}}{2}, \frac{V_{CEQ}}{2} \right]$ and that $i_{CQ} \in \left[-\frac{I_{CQ}}{2}, \frac{I_{CQ}}{2} \right]$.

- For this type of amplifiers, harmonic distortion is usually very low: THD < 0.1% (Very high linearity).

• **Power and efficiency of Class A amplifiers** → let's study which percentage of the amplifier's supply voltage is actually converted to output power.

• We can note that there are two main components:

• **Dissipated power** → it is the power used by the amplifier during its operations: $P_{diss}(t) = P_{transistor}(t) - P_L(t)$.
- DC component:

- $P_{transistor}^{DC} = (i_C \cdot i_B) \cdot V_{CC}$, but $i_C = \beta i_B$ and β is very high, so $i_B \approx 0 \Rightarrow P_{transistor}^{DC} \approx i_C \cdot V_{CC} = i_C \cdot \frac{V_{CC}}{2}$
- $P_{R_E}^{DC} = i_C \cdot V_{R_E} = i_C \cdot \frac{V_{CC}}{2} = \frac{V_{CC}^2}{4 R_E}$
- $P_{diss}^{DC} = P_{transistor}^{DC} + P_{R_E}(t) = \frac{V_{CC}^2}{2 R_E}$

- AC component: in this case, $Z_C \approx 0 \Rightarrow$ the two resistors are in parallel.

- $P_{transistor}^{AC} = i_C \cdot V_{R_E \parallel R_L} = i_C \cdot \frac{V_{CC}}{2} = \frac{V_{CC}^2}{4 R_E \parallel R_L}$
- $P_{R_E}^{AC} = P_{transistor}^{AC}$
- $P_{diss}^{AC} = \frac{V_{CC}^2}{2 R_E \parallel R_L} \rightarrow$ usually $R_E \gg R_L \rightarrow P_{diss}^{AC} = \frac{V_{CC}^2}{2 R_L}$

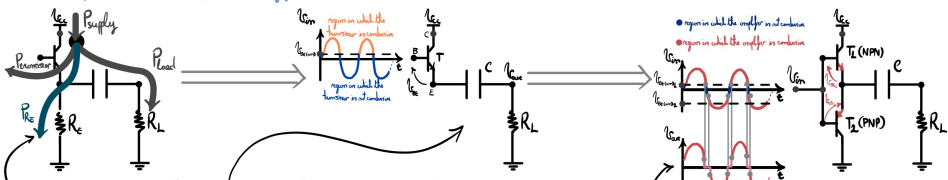
• **Output power** → it is the portion of power that is actually converted in output amplified signal: $P_{out} = P_L$.
It has only an AC component:

- let's consider the case in which the output signal is $i_C(t) = \hat{I}_C \sin(\omega t)$.
- The instantaneous power provided to the load is $P_{out}(t) = i_C(t) \cdot v_L(t) = \frac{V_L^2(t)}{R_L} = \frac{V_L^2 \cdot \sin^2(\omega t)}{R_L}$.
- Knowing that the average value of $\sin^2(\omega t) = \frac{1}{2}$, we can write that $P_{out} = \frac{V_L^2}{2 R_L}$.

• At this point, we calculate the efficiency or $\eta = \frac{P_{out}}{P_{diss}} = \frac{\frac{V_L^2}{2 R_L}}{\frac{V_{CC}^2}{2 R_L}} = \frac{V_L^2}{V_{CC}^2} \cdot \frac{R_E}{R_L} \rightarrow$ to maximize η , we set $V_L = \pm \frac{V_{CC}}{2}$ and $R_E = R_L \Rightarrow \eta = \frac{R_E}{R_L} \cdot \frac{V_{CC}^2}{V_{CC}^2} = 0.25 = 25\%$

• **Class B** → it is an "evolution" of the class A amplifier, created with the aim of increasing the efficiency.

• Steps to get a Class-B amplifier starting from a Class-A amplifier:



① The main "efficiency problem" of this configuration is the power dissipated over R_E , so a first approach could be to simply remove this resistor and search for a solution.

② Let's analyze the behaviour of the new component when we feed it with a sinusoidal signal:
• if $v_{in}(t) < v_{be(on)}$ the transistor is in cut-off region and $i_{B1} \approx 0$
• if $v_{in}(t) > v_{be(on)}$ the transistor is in forward active region and $i_{B1} \approx 0$
To reduce this problem (get rid of the negative part of the sinusoid) we can add another transistor of the opposite polarization type, where one of these was R_E .

③ As we can notice, we got back a big portion of negative waveform, but there is still a region where $v_{be(on)} < v_{in} < v_{be(on)}$ in which both the transistors are in non-conductive region and the output is 0. This discontinuity causes the phenomenon of **crossover distortion**.

• **Main characteristics of Class-B amplifiers:**

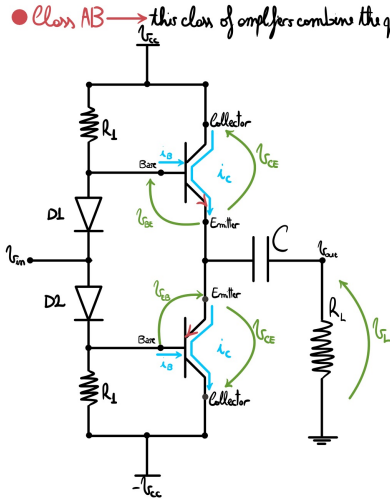
- **angle of conduction** → each transistor is conductive only for $\frac{1}{2}$ of the waveform $\Rightarrow \theta = 180^\circ$.
- **efficiency** → let's calculate the efficiency of this type of amplifier:

- **Load power** → if $i_{in}(t) = \hat{I}_C \sin(\omega t)$ we have $i_{C1}(t) \approx i_{C2}(t) \Rightarrow P_{out}(t) = \frac{V_L^2(t)}{R_L} = \frac{V_L^2 \cdot \sin^2(\omega t)}{R_L} \Rightarrow P_{out} = \frac{V_L^2}{R_L} \cdot \frac{1}{2} = \frac{V_L^2}{2 R_L}$

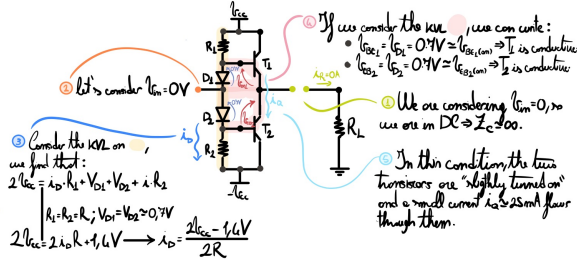
- **Source power** → during the positive half-cycle, T_1 engages a current equal to $i_{C1}(t) = \frac{V_L \sin(\omega t)}{R_L}$
negative half-cycle, T_2 engages a current equal to $-i_{C2}(t) = \frac{V_L \sin(\omega t)}{R_L}$
the average power in $P_{S1} = V_{CC} \cdot i_{C1}(t)$
 $= \frac{2}{\pi} V_{CC} \frac{V_L}{R_L}$

• **quality of the signal** → the signal presents a lower quality compared to the output of a Class-A amplifier, due to the crossover distortion.

● **Class AB** → this class of amplifiers combines the qualities of A-class amplifiers (linearity) and AB-class ones (efficiency). Let's see how.



● The main idea is to take the class B amplifier and to modify it in order to have one of the transistors "slightly turned on" during the $v_{CE1} < v_{CE2}$ phase, minimizing the crossover distortion. To do so, we introduce $D1$ and $D2$ and $R1$:



● In this way, in the "control region" of the waveform, at least one of the transistors is activated.

● **AB-class amplifier's properties:**

● **Angle of conduction** → each transistor conduct for a bit more of the half waveform $\Rightarrow 80^\circ$ to 200° .

● **Efficiency:**

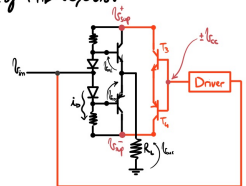
- **Load power** → we have $v_{out} = v_{in} = V_p \sin(\omega t) \rightarrow \bar{P}_{load} \approx \frac{V_p^2}{2R_L}$.
- **Source power** → $i_C(\omega) \approx \frac{V_p \sin(\omega t)}{R_L} \Rightarrow \bar{P}_S = \frac{2}{\pi} \cdot \frac{V_{CC} V_p}{R_L} \xrightarrow{\text{load in the } V_p} \bar{P}_S = \frac{2}{\pi} \cdot \frac{V_{CC}^2}{R_L}$.
- **Efficiency** → $\eta = \frac{\bar{P}_{load}}{\bar{P}_S} = \frac{P_{load}}{P_S} \approx 78.5\%$.

(for further details see TDE - 2 July 2013)

● **Sound quality:** in AB-class amplifiers there is still a small crossover distortion, but it is often in the range of $THD < 0.05\%$.

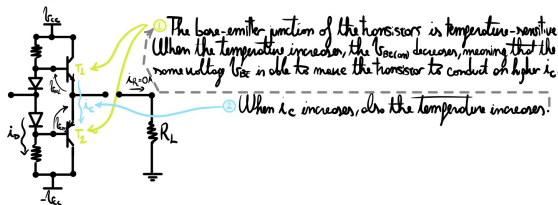
● **Class AB amplifier with driver** → in order to reduce the overlapping between the two transistors $T1$ and $T2$, we can introduce a driver and two transistors (draw on the right). The driver is a digital device that:

- generates $+v_{CE}$ if $v_{in} > 0$ → when $+v_{CE} \rightarrow T3$ is turned on and $v_{CE} = +v_{CE}$ while $v_{CE} = 0$.
- generates $-v_{CE}$ if $v_{in} < 0$ → when $-v_{CE} \rightarrow T4$ is turned on and $v_{CE} = 0V$ while $v_{CE} = -v_{CE}$.



● **Problems - solutions of Class AB amplifiers:**

● **Thermal compensation** → consider to be in the situation in which the amplifier has been turned on for a long time and the transistors' temperature is around $70^\circ C$.

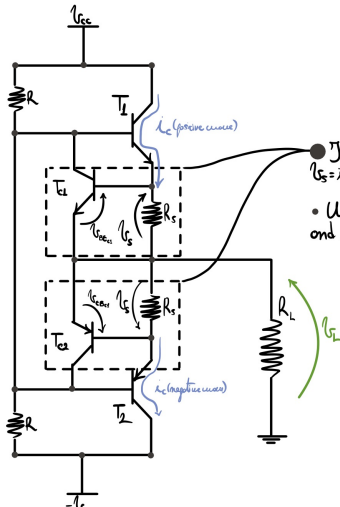
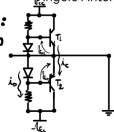


→ this cycle leads i_C to diverge, desaturating $T1$ and $T2$. To avoid this effect, there are 2 main possible solutions:

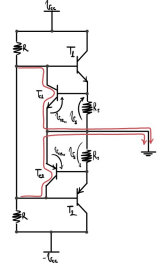
● **Solution 1: thermal compensation diodes:** this approach consists in limiting the diodes $D1$ and $D2$ to the same positive dissipation system that is connected to $T1$ and $T2$. In this way, when the temperature of $T1$ increases also $D1$ will become hotter, and v_{D1} will be lower, avoiding i_C to diverge.

● **Solution 2: trimmer:** we insert a potentiometer in the bias network of the amplifier. When we turn on the amp, we wait for it to reach the "same temperature", then we set the potentiometer in order to have the correct i_{B2} .

● **Short-circuit protection** → Consider the case in which, for any reason, the emitter of the Transistors is connected to ground. In this case, we are practically short-circuiting the power supply to ground and the transistor is subject to a very high current that flows through it, leading to the possibility to burn the transistor. To avoid this possibility, the following configuration is used:

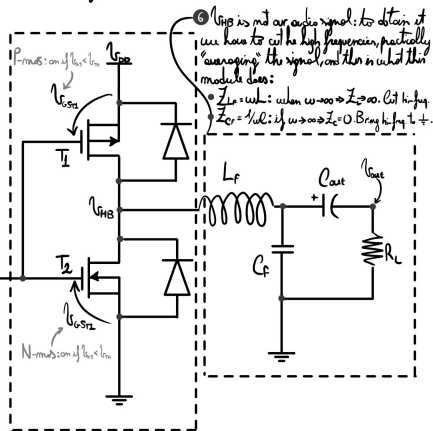
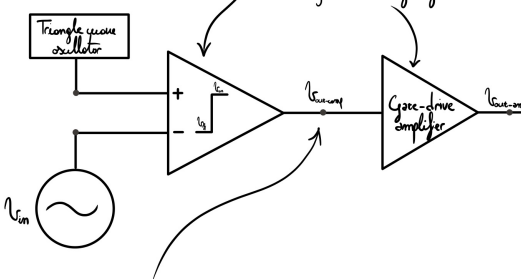


- If the emitter of the transistors T_1 and T_2 is connected to ground, i.e. ground, and also the voltage $V_E = V_C$ is ground. We can notice that $V_E = V_C \Rightarrow$ also T_1 (or T_2) turns on.
- When T_1 (or T_2) turns on the base of T_2 (or T_1) is connected to $V_B = 0 \Rightarrow T_1$ (or T_2) is turned off and I_C goes to zero.

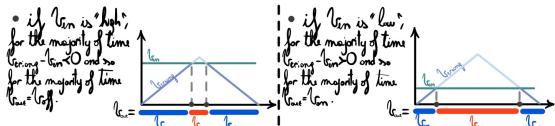


● **Class D** → the aim of Class-D amplifiers is to maximize the efficiency. To do so, we have to dissipate the lesser amount of heat possible. The 'switching setting' in which transistors consume the less amount of heat when they are used as simple switches, to get '1' and '0' signals. But how can we use just 'on' and 'off' signals to reconstruct the amplified signal at the output of the amplifier? We do so using the **Pulse Width Modulation technique**.

- The comparator is an amplifier used in its saturation range. In this way, the only two output possible are V_{on} and V_{off} . The working principle is the following:
 - if $V_{in} - V_{ref} > 0 \Rightarrow V_{out} = V_{on}$
 - if $V_{in} - V_{ref} < 0 \Rightarrow V_{out} = V_{off}$
- This amplifier moves the signal strong enough to drive the gates of the transistors



● Let's consider one triangle wave cycle and let's see what happens to V_{comp} :



● This module is called 'half bridge' and it works in the following way:

- if $V_{comp} = V_{on} \Rightarrow T_1$ is turned on and T_2 is turned off $\Rightarrow V_{out} = 0V$
- if $V_{comp} = V_{off} \Rightarrow T_1$ is turned off and T_2 is turned on $\Rightarrow V_{out} = V_{DD}$
- In this way, there is not one moment in which both the transistors are turned on simultaneously, so there is not a short-circuit connection between V_{DD} and $0V$, and both the transistors work only in saturation region when they are turned on. This means the dissipated power is very little.

• The output of the comparator is a square wave $V_{comp} = \dots$

● Advantages and disadvantages of Class-D amplifiers:

- **Efficiency** → it is very high, usually $\eta \approx 85-90\%$ as a result of the very low heat. This leads also to a decrease of the size of the dissipating system and less problems in managing high temperatures (vs. AB amplifiers).
- **High power** → it is easier to scale this type of amplifiers to hundreds of Watt maintaining high efficiency.
- **Sound quality** → it is lower compared to other types of amplifiers due to the 'averaging' of the input signal $\Rightarrow$ THD $\approx 0.5\%$, but we can lower that value using good (complex and expensive) LC circuits to cut the High-Frequency components.