CALIFORNIA INSTITUTE OF TECHNOLOGY
PHYSICS MATHEMATICS AND ASTRONOMY DIVISION

Sophomore Physics Laboratory (PH005/105)

Analog Electronics
TheOperationalAmplifier

Chapter 4

The Operational Amplifier

4.1 Introduction

Operational amplifiers are one of the most extensively used analog integrated circuits especially because of their ability to approximate reasonably well the ideal behavior. For this reason, real operational amplifiers can be quite often modeled as ideal or quasi-ideal. Moreover, the versatility this device, which hides a large internal complexity [1], makes the operational amplifier suitable for many different applications.

In the first section, the mathematical formulation makes the ideal operational amplifier concept quite awkward, especially after a first reading. Using the ideal device properties in conjunction with the so called feedback network, which links the output of the amplifier inputs, will clarify the definition of the ideal operation amplifier.

The subsequent section is dedicated to the explanation of some basic operational amplifier circuits. Finally, section 4.4 introduces a more realistic model of operational amplifier together with some of the particular behavior of this electronic device.

[1] A modern operational amplifier made of a cascade of stages, each one designed mainly to match the ideal characteristics, can have around 50 components both active and passive. See the Analog Devices Web site, for example.

4.2 The Ideal Operational Amplifier

The ideal operational amplifier (Op-Amp) is a linear amplifier with two differential inputs v_+, v_- and one output v_o (see figure 4.1) and with the following characteristics:

- $v_0 = A_v(v_+ - v_-), \qquad A_v > 0$, (linearity)
- input resistance $R_i \to \infty$,
- output resistance $R_o \to 0$,
- voltage gain $A_v \to \infty$,
- frequency response constant for any frequency.

Aside the welcome property of linearity and infinite frequency response, the need of all the other characteristics can be justified as follows. Infinite input resistance R_i means essentially that the Op-Amp inputs do not produce perturbations to any circuit to which they are connected to. Zero output resistance R_o perfectly isolates the Op-Amp from any perturbation. Infinite input impedance and zero output impedance implies also no dissipation of energy. The condition of infinite voltage gain A_v is necessary if we want a device able to deliver any gain, once a network which connects the output to the input is added to the Op-Amp. In general, this kind of network is called *feedback network*.

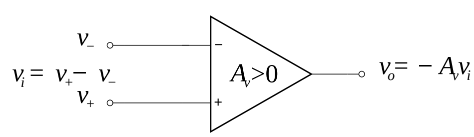

Figure 4.1: Op-Amp symbol, and some definitions of variables.

4.2.1 Ideal Op-Amp Fundamental Equation (Golden Rules)

The consequence of the following conditions

- $A_v \to \infty$,
- $v_o < \infty$ if $v_i = v_+ - v_- < \infty$,

is the following very useful and important formula

$$v_+ - v_- = 0, \qquad \text{at all times.} \tag{4.1}$$

Equation 4.1 will be called the *first Op-Amp golden rule.*
The consequence of the following condition

- $R_i \to \infty$,

is

$$i_+ - i_- = 0, \qquad \text{at all times,} \tag{4.2}$$

Equation 4.2 will be called the *second Op-Amp golden rule*.

These two rules are fundamental for the solution of any circuit involving Op-Amps. We will see in the next sections the importance of this equations once a feedback network is connected to the Op-Amp.

4.2.2 Op-Amp Input Output "Logic"

It is worthwhile here to notice the behavior of Op-Amp output as function of the two inputs. From the definition of Op-Amp we have that a signal sent to the negative input V_- is amplified and changed in sign. A signal sent to the positive input V_+ is just amplified. Two signals sent each to one input are indeed subtracted and amplified.

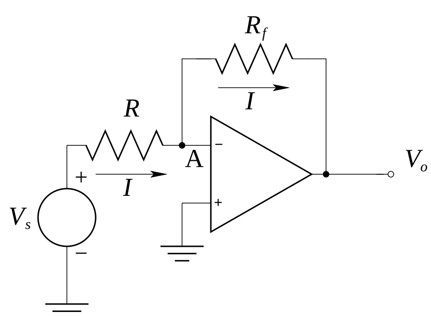

Figure 4.2: Op-Amp with a feedback network.

4.2.3 Op-Amp with a Feedback Network

Let's consider the circuit in figure 4.2, where a feedback resistance R_f is connected to the negative input. The current through the resistors R and

R_f is the same because the ideal Op-Amp input does not drive any current ($R_i = \infty$). Furthermore, since $V_i = V_+ - V_- = 0$ and with the use Ohm's law and the KVL, it follows that

$$I = \frac{V_s}{R} = -\frac{V_o}{R_f}. \tag{4.3}$$

The output voltage V_o and voltage gain A will be

$$V_o = AV_s, \qquad A = -\frac{R_f}{R}.$$

The gain of the Op-Amp depends just on the resistances ratio R_f and R.

4.2.4 The Virtual Ground

Let's re-analyze the circuit in figure 4.2. Considering that the golden rule imposes $V_i = V_+ - V_- = 0$, and the negative input is grounded, the node **A** must always be at zero voltage. This is equivalent to having **A** virtually grounded. The adjective virtual is necessary because even if **A** is at the potential of the ground there is no current flowing through **A** ($R_i = \infty$) as in a real ground**.** In other words, the virtual ground happens to be because the Op-Amp does its best to keep $V_i = 0$.

4.3 Commonly Used Op-Amp Circuits

In the study of the several common Op-Amp configurations, we will use the approximation of an ideal circuit. A more realistic model is often necessary to understand some behaviors of real circuits. For an initial design, and where the the ideal Op-Amp characteristics are well approximated, the ideal model is quite often sufficient.

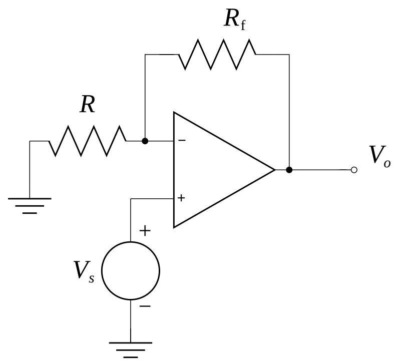

Figure 4.3: Non-inverting configurations of the Op-Amp.

4.3.1 Non-Inverting Amplifier

Let's consider the non-inverting configuration of the Op-Amp in figure 4.3. Because of $V_i = 0$, we will have

$$V_s - V_- = V_s - RI = 0.$$

Considering that the output voltage V_o is

$$V_o = (R_f + R)I,$$

we can use the expression of I to obtain

$$V_s = \frac{R}{R + R_f} V_o.$$

The output voltage V_o and voltage gain A will be

$$V_o = AV_s, \qquad A = 1 + \frac{R_f}{R}.$$

Considering that in this configuration V_s is directly connected to $V_+$and V_- is not a virtual ground, the input impedance of the amplifier is $R_i + R$, where R_i is the real input impedance of the Op-Amp.

4.3.2 Inverting Amplifier

This circuit has been already discussed in section 4.2.3. For completeness, the solution and some comments are here reported

$$V_o = AV_s, \qquad A = -\frac{R_f}{R}.$$

It is worthwhile to notice that because $V_- = 0$, the circuit input impedance is just R. Having values of R typically of few $k\Omega$, the inverting configuration doesn't preserve the high impedance characteristic of an Op-Amp. A connection of the circuit input to a network can potentially create appreciable perturbations.

4.3.3 Differential Input Stage

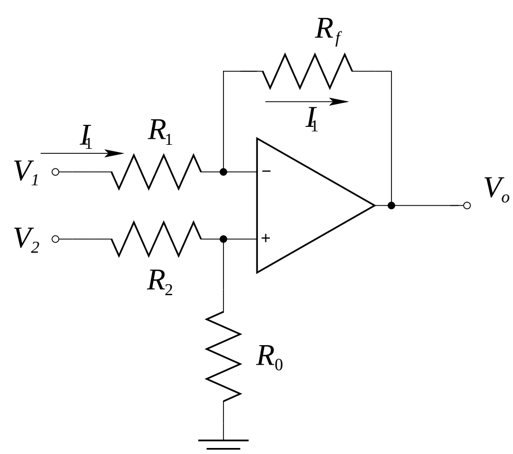

Figure 4.4: Differential input configuration of the Op-Amp.

Let's now solve the differential input circuit of the Op-Amp in figure 4.4.

Writing the voltage drop across $R_1$and R_f, we obtain the linear system

$$\begin{aligned} V_- - V_1 &= R_1 I, \\ V_o - V_- &= R_f I. \end{aligned}$$

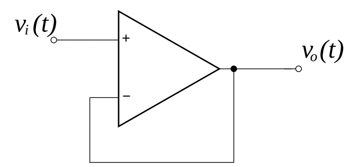

Figure 4.5: Voltage follower or unity gain buffer.

Solving the system with respect to V_o, we get

$$V_o = \left(1 + \frac{R_f}{R_1}\right) V_- - \frac{R_f}{R_1} V_1.$$

Using the voltage divider equation to obtain $V_+$and because $V_+ - V_- = 0$, we have

$$V_- = V_+ = \frac{R_0}{R_2 + R_0} V_2,$$

and finally, we get

$$V_o = \frac{R_1 + R_f}{R_1} \frac{R_0}{R_2 + R_0} V_2 - \frac{R_f}{R_1} V_1.$$

A way to to obtain the same voltage gain for V_2 and V_1 is to impose $R_0 = R_f$ and $R_1 = R_2 = R$. The output voltage becomes

$$V_o = A(V_2 - V_1), \qquad A = \frac{R_f}{R}.$$

This differential configuration is not very convenient because it does not preserve the high input impedance of the Op-Amp. In fact , considering that the Op-Amp input impedance is very high, we have that the resistance seen from V_1 is $R_2 + R_0$. Usually, the sum of those resistors is at least one order of magnitude smaller than the Op-Amp input impedance. If we need to build a variable gain differential amplifier, we will need to change more than one resistor value. Matching the resistances values can become an issue when thermal drifts become important.

More practical and stable configurations called instrumentation amplifiers are available "off the shelf".

4.3.4 Voltage Follower (Unity Gain "Buffer")

The circuit sketched in figure 4.5 is called voltage follower or unity gain buffer. The feedback line with no load gives

$$v_o(t) = v_-.$$

Moreover, because of the golden rule we will have

$$v_- = v_+,$$

which implies

$$v_o = v_i.$$

The output voltage $v_o(t)$ follows the input voltage $v_i(t)$ with unitary gain.

Considering that the high impedance input and the low impedance output values of Op-Amps are close to the state of the art in the electronic design[2], the voltage follower can be used as an isolation stage (buffer) between two circuits.

4.3.5 Integrator Amplifier

Let's consider the circuit in figure 4.6 without the resistance R_f. The voltage drop v_o across the capacitor C_f is

$$v_o(t) = -\frac{1}{C_f}\int_{-\infty}^{t} i(\tau)d\tau \tag{4.4}$$

and the current flowing through the resistance R is

$$i(t) = \frac{v_i(t)}{R}.$$

[2]Devices expressly made to work as input unity gain buffer, and output unity gain buffer are also available. Analog Devices SSM2141 and SSM2142 are complementary buffers devices which can drive long delay lines for example.

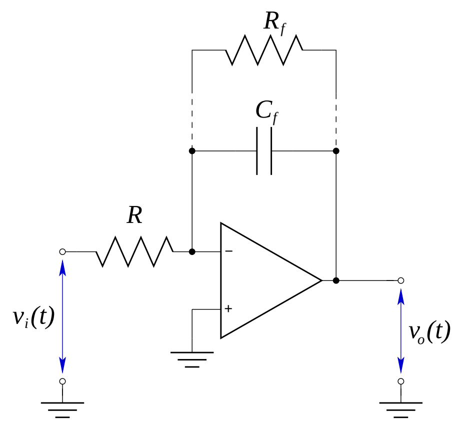

Figure 4.6: Active integration stage using an Op-Amp.

Placing the expression of $i(t)$ obtained from the previous equation into eq.(4.4), we will obtain

$$v_o(t) = -\frac{1}{\tau}\int_{-\infty}^{t} v_i(t')dt', \qquad \tau = RC_f.$$

Real Op-Amps or signals connected to the input have often (always) a DC offset. This offset is indeed integrated and after a given time will saturate the amplifier output. This saturation is essentially a manifestation of the instability of the circuit at low frequency. Moreover, the initial charge of the capacitor is undefined, making the initial output state unpredictable.

A common way to avoid these problems is to introduce the resistance R_f in parallel with the capacitor C_f which reduces the amplifier DC gain. An intuitive way to understand the effect of this feedback resistance is that it does not allow the capacitor to be charged "ad libitum". The choice of the R_f is not so trivial if we want to preserve the characteristic of good integrator. Using the simple phasor analysis it easy to prove that the good integrator condition is $\omega \gg 1/C_f R_f$.

If the DC current must be integrated, we can place a switch in parallel with the capacitor to be opened when the integration is started. In this way we will have the capacitor state completely defined.

4.3.6 Differentiator Amplifier

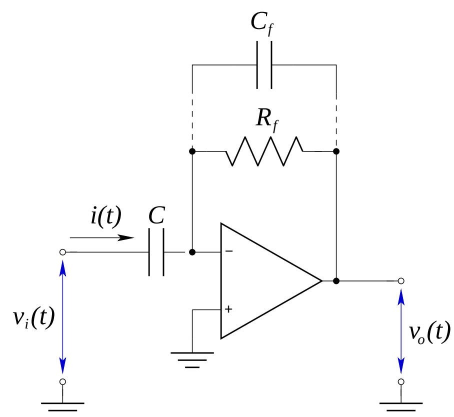

Figure 4.7: Differentiator stage using an Op-Amp.

Let's now consider the circuit in figure 4.7 without the feedback capacitor C_f. Applying a similar analysis to that used in the integrator amplifier we will have

$$\begin{aligned} i(t) &= C\frac{dv_i}{dt}, \\ v_o(t) &= -R_f i. \end{aligned}$$

and indeed

$$v_o(t) = -\tau\frac{dv_i}{dt}, \qquad \tau = R_f C.$$

This configuration without C_f doesn't work well with real Op-Amps, because of stability problems. In fact, the introduction of the capacitor compromises the internal compensation of the Op-Amp. Placing a capacitor C_f in the feedback network restores the compensation making the overall circuit stable. The choice of C_f is not trivial if we want to preserve the circuit differentiator characteristics.

Section 4.4.3 explains in more details the effect of this configuration on the compensation of a real Op-Amp.

4.4 The Real Op-Amp

Lets consider in this section a more realistic model of the Op-Amp by including a finite input impedance R_i, non zero output impedance R_o, finite gain A, bias currents and voltage offsets. Using ideal components, the equivalent circuit of the real Op-Amp is shown in figure 4.8.

4.4.1 Bias Currents and Voltage and Current Offsets

Imbalances inside of the Op-Amp, mainly due to differences in the electronics components, produce undesirable bias currents and a voltage offset at the inputs. Input voltage and current offsets can be modeled by introducing ideal generators as shown in figure 4.8. Current Offset is defined as the difference in the magnitude of the bias currents, i.e.

$$i_{os} = |i_{b+}| - |i_{b-}|$$

A way to characterize the voltage offset is to use the voltage follower configuration (see section 4.3.4) with the input V_i connected to the ground. The voltage offset will be directly the output voltage V_o.

Current input biases can be studied connecting a resistor between the one input and ground and measuring the voltage drop across the resistor.

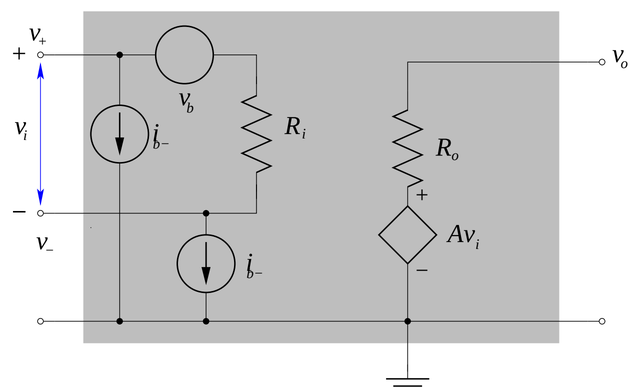

Figure 4.8: Equivalent circuit for an Op-Amp using ideal components. Voltage offsets and current biases are taken into account using ideal voltage and current generators.

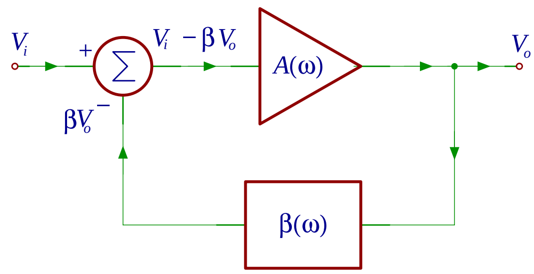

Figure 4.9: Amplifier with negative feedback.

4.4.2 Feedback Amplifiers

Let's consider an amplifier with a negative feedback network as show in figure 7.1. Considering that the summation point output is

$$V_i - \beta(\omega)V_o\,,$$

and the amplifier gain is $A_{OL}(\omega)$, the output voltage will be

$$V_o = A(V_i - \beta V_o)\,.$$

Collecting V_o, we will have

$$V_o = \frac{A}{1+\beta A}\,V_i\,,$$

and the so called *closed loop transfer function*, A_{CL}, will finally be

$$A_{CL}(\omega) = \frac{A}{1+\beta A}\,. \tag{4.5}$$

We can clearly see that if the denominator goes to zero for a given frequency ω^*, we are in trouble, $A_{CL}(\omega^*)$ diverges, and the amplifier saturates. The trick to avoid this situation, is to study the following equation

$$A_{OL}(\omega) = \beta(\omega)A(\omega) = -1, \tag{4.6}$$

where A_{OL} is the *feedback amplifier open loop transfer function*. If the phase where the magnitude of $A_{OL}$is equal to one is different from 180^0plus multiples of 360^0, the denominator never goes to zero and the saturation is avoided. However, this is not enough because we can have just an oscillation with no saturation if the A_{OL} phase is too close to 180^0. The rule of thumb is to have a so called phase margin of about $60^0$from 180^0. Finally, we can formulate the criterion for the stability:

$$\text{where } |A_{OL}| = 1 \qquad \Rightarrow -120 < \arg(A_{OL}) < 120\,.$$

Another important result of the theory of feedback amplifier is the following straightforward result

$$\text{if} \quad \beta A \gg 1 \qquad \Rightarrow \qquad A_{CL}(\omega) \simeq \frac{1}{\beta}\,.$$

Where the open loop transfer function $A\beta$ is greater than one the feedback amplifier response does not depend on the response $A(\omega)$ of the amplifier with no feedback. It is worthwhile to notice that the ideal amplifier ($A \to \infty$) has $A_{CL} = 1/\beta$ for all angular frequencies. A close loop ideal amplifier does not have undesired instabilities, but just the ones that can be introduced by the feedback network.

4.4.2.1 Non-Inverting Configuration

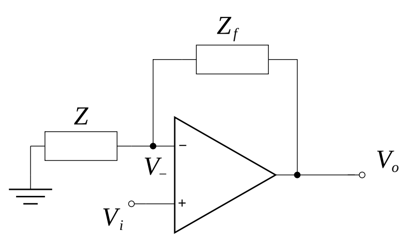

Figure 4.10: Non-inverting configuration Op-Amp with generic impedance.

Considering the Op-Amp Non-Inverting configuration as shown in figure 4.10, and the voltage divider equation we have

$$V_- = \frac{Z}{Z_f + Z} V_o\,, \tag{4.7}$$

and the feedback network transfer function is

$$\beta(\omega) = \frac{V-}{V_0} = \frac{Z}{Z_f + Z}\,.$$

The approximate gain of the feedback amplifier is as expected

$$A_{CL} \simeq \frac{1}{\beta} = 1 + \frac{Z_f}{Z}.$$

4.4.2.2 Inverting Configuration

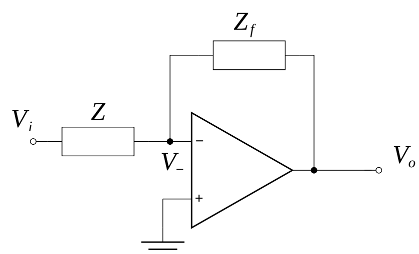

Figure 4.11: Inverting configuration Op-Amp with generic impedance.

In this case the feedback network transfer function β is

$$\beta = \frac{V_i}{V_0}$$

Considering the inverting configuration stage as shown in figure 4.11, because of the virtual ground we have

$$\begin{cases} V_i = ZI \\ -V_o = Z_f I \end{cases} \Rightarrow \quad \beta(\omega) = -\frac{Z}{Z_f},$$

and the gain of the feedback amplifier is simply

$$A_{CL} \simeq \frac{1}{\beta} = -\frac{Z_f}{Z}.$$

4.4.3 Compensated Op-Amp Transfer Function

Practical Op-Amps are often designed to have a frequency response dominated by a single pole, i.e. the transfer function with no feedback is just a simple low-pass filter. In this case, the Op-Amp transfer function with no feedback can be written as

$$A(\omega) = \frac{A_0}{1 + j\frac{\omega}{\omega_0}}, \tag{4.8}$$

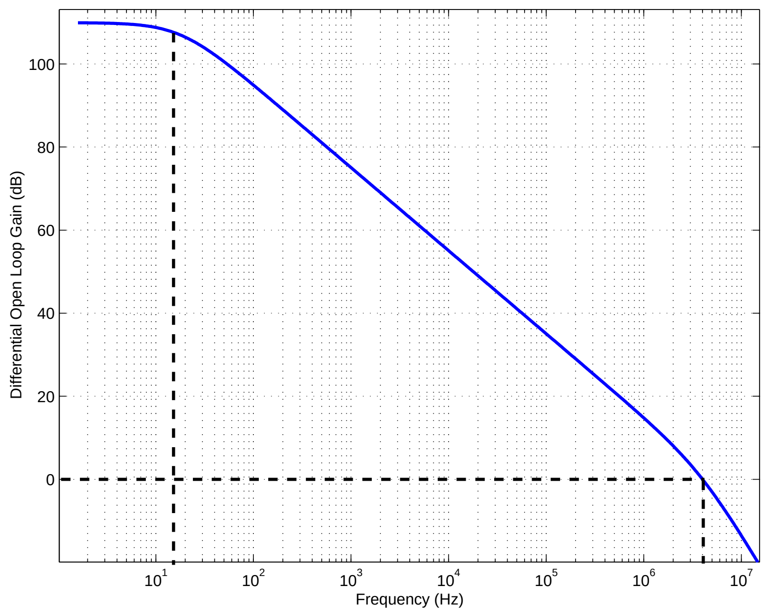

Figure 4.12: Differential gain of the AD711 Op-Amp (cut-off frequency $\nu_0 =$ 18Hz, unity gain frequency $\nu_1 = 4$MHz, and DC gain $a_0 = 110$dB). Dominant single pole behavior is valid up to about 4MHz, where the slope becomes steeper than $1/\omega$.

where A_0 is the DC gain and ω_0 is the angular frequency of the dominant pole (the cut-off angular frequency of the low pass filter). This behavior is obtained by introducing a compensating circuit (quite often a capacitor) in the architecture of the Op-Amp.

This choice comes from the stability requirement that we mentioned in the previous section. In fact, an amplifier with a transfer function with a dominant pole cannot lose more than 90^0 making quite easy the design of a stable feedback network. For example, feedback networks with just resistors, (ideal resistors don't loose phase) will not generate oscillations.

Typical values for pole frequencies are between 5Hz and 100Hz. Figure 4.12 shows the differential transfer function of the Op-Amp AD711 with no feedback network.

Let's study more in details the compensated Op-Amp response with a feedback.

4.4.4 Compensated Op-Amp with Constant Frequency Response Feedback

Considering the frequency response of a compensated feedback amplifier, we will have

$$\begin{aligned} A(\omega) &= \left(\frac{A_0}{1+j\frac{\omega}{\omega_0}}\right) \Bigg/ \left(1+\frac{\beta A_0}{1+j\frac{\omega}{\omega_0}}\right), \\ &= \frac{A_0}{1+\beta A + j\frac{\omega}{\omega_0}}, \\ &= \left(\frac{A_0}{1+A_0\beta}\right) \Bigg/ \left(1+j\frac{\omega}{\omega_0(1+\beta A_0)}\right). \end{aligned}$$

In the particular case that $\beta(\omega)$ is constant and $\beta = \beta_0 \geq 1$, the previous equation becomes

$$A(\omega) = \frac{A_1}{1+j\frac{\omega}{\omega_1}}, \qquad \begin{cases} A_1 = \frac{A_0}{1+A_0\beta_0} \\ \omega_1 = \omega_0(1+\beta_0 A_0) \end{cases}$$

In this case, the feedback Op-Amp response is the same as of the open loop transfer function A_{OL}but with a smaller DC gain (about $1/\beta_0$) and higher cut off angular frequency ω_1 $\omega_0\beta_0A_0$.

4.4.5 Compensated Op-Amp in the Differentiator Configuration

Let's find the frequency response of the "real" Op-Amp differentiator circuit of figure 4.7. For the basic differentiator (no capacitor C_f) the feedback network transfer function is

$$\beta = -\frac{Z}{Z_f} = -\frac{1}{j\omega RC} = j\frac{\omega_1}{\omega}, \qquad \omega_1 = \frac{1}{RC}.$$

Considering eq. (4.5), we have that the gain of the differentiator is

$$A_{CL}(\omega) = \frac{A(\omega)}{1+jA(\omega)\,\omega_1/\omega} = \frac{1}{1/A + j\,\omega_1/\omega},$$

and finally

$$A_{CL}(\omega) = \frac{A_0}{1 + j\left(A_0\frac{\omega_1}{\omega} + \frac{\omega}{\omega_0}\right)} = -A_0\omega_0\frac{j\omega}{\omega^2 - j\omega\omega_0 + A_0\omega_0\omega_1}\,.$$

Resonance occurs when the denominator goes to zero, i.e for

$$\omega^* = j\frac{\omega_0}{2} \pm j\sqrt{\frac{\omega_0^2}{4} + A_0\omega_0\omega_1} \simeq \pm j\sqrt{A_0\omega_0\omega_1}\,, \qquad \omega_0 \ll \omega_1,\ A \geq 1\,.$$

4.4.6 The Common Mode Rejection Ratio (CMRR)

We want characterize the rejection of an Op-Amp output as a differential amplifier, of signals sent to both inputs. For an ideal Op-Amp we expect to obtain $V_o = 0$ for all frequencies, i.e. a perfect rejection. To define a convenient parameter which measures the rejection it is necessary to define the following ones, the *common mode gain*

$$A_C(\omega) = \frac{V_o}{V_+ - V_-}, \qquad V_+ = V_- = V_s\sin(\omega t),$$

and the *differential mode gain*

$$A_D(\omega) = \frac{V_o}{V_+ - V_-}, \qquad V_+ = V_s\sin(\omega t), \quad V_- = 0.$$

The *Common Mode Rejection Ratio (CMRR)* is defined as the modulus of the ratio of the differential gain A_D over the common mode gain A_C, i.e.

$$CMRR(\omega) = \left|\frac{A_D(\omega)}{A_C(\omega)}\right|$$

Ideally, the $CMRR$ should be infinity for all frequencies.

This parameter can be measured using the Op-Amp differential configuration (see figure 4.4) and measuring A_C and A_D as a function of the frequency. To minimize possible large systematic errors, it is necessary to have the same gain for the to inputs V_1 and V_2. This can be achieved by placing a trimmer in the voltage divider mesh of the differential configuration circuit. Adjusting the trimmer we can minimize V_o for a single frequency and study the $CMRR$ for a given bandwidth.

Figure 4.13shows the $CMRR$ as a function of frequency of a typical Op-Amp. A typical value for $CMRR$ is 90dB.

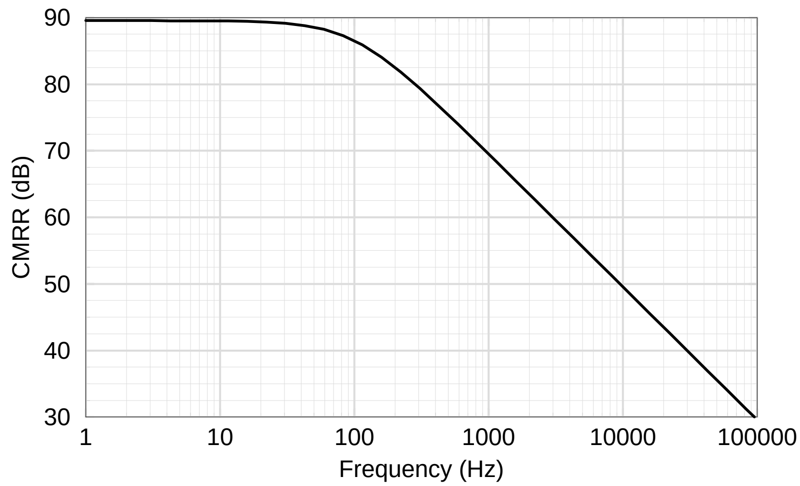

Figure 4.13: *CMRR* as a function of frequency of a typical Op-Amp.

4.4.7 The Gain Bandwidth Product (GBWP)

The gain bandwidth product is a common way to characterize the gain with respect to the available bandwidth of amplification. It is defined as

$$GBWP = A_0\omega_0.$$

The larger the GBWP the better is the Op-Amp, and the closer the Op-Amp is to the ideal operational amplifier.

4.4.8 The Slew Rate (SR)

The *slew rate* or *maximum slew rate SR* of an Op-Amp is defined as the maximum rate of the output voltage v_o per unit time

$$SR = \max\left\{\left|\frac{\Delta v_o(v_i)}{\Delta t}\right|\right\}.$$

This parameter essentially measures the ability of an Op-Amp to follow voltage changes for large voltage inputs.

The slew rate can be easily observed sending a square wave (see figure 4.14) to the Op-Amp input v_i, and looking at the raising and falling slope of the output signal v_o. If the slopes do not change changing the input amplitude, then the Op-Amp is slew rate limited.

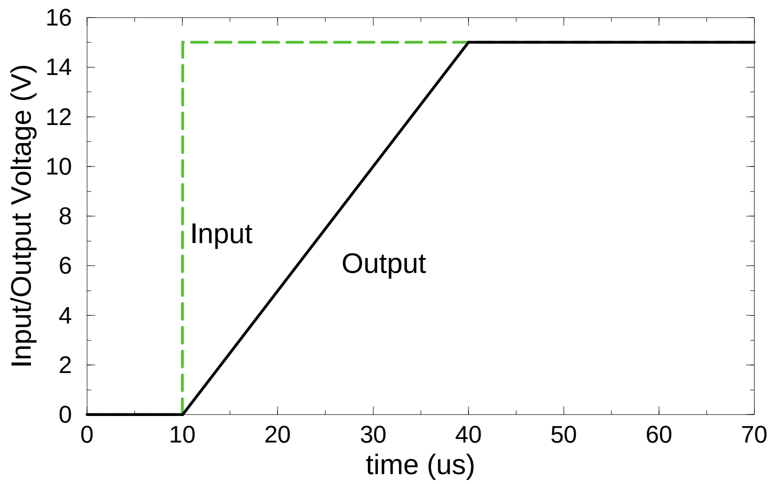

Figure 4.14: Slew rate illustration. The output v_o takes 40μs to reach the desired voltage. The device is said to be slew rate limited.

A similar procedure can be applied using a low frequency sinusoidal signal as input. In this case if we increase the input amplitude too much, the output will become distorted. Considering a sinusoid of frequency $\omega_0$and amplitude V_o

$$SR = \max\left\{\left|\frac{d}{dt}V_o \sin\,\omega_0 t\right|\right\} = \omega_0 V_o \max\left\{\cos\,\omega_0 t\right\} = \omega_0 V_0.$$

For an undistorted signal with amplitude V and maximum frequency ω, we must have

$$SR \gg \omega V\ .$$

Usually, a frequency compensated Op-Amp has a passive integrating stage at the output, and therefore the slew rate is often proportional to the inverse of the capacitance of the compensating network.

Typical good slew rate values are of the order of few V/μs.

4.4.9 Ideal versus Real and Practical Considerations

The following table summarizes the main characteristic of an ideal Op-Amp together with those of typical real Op-Amp. In some cases we can find Op-Amps excelling some of the mentioned characteristics , quite often at the expense of other characteristics.

Property	Ideal Op-Amp	Typical Op-Amp
Open-Loop DC Gain A_v	∞	$> 10^4$
Open-Loop Bandwidth	∞	$\sim$ 10Hz (dominant pole)
Common Mode Rejection Ratio $CMRR$	∞	> 70dB
Input Resistance R_i	∞	$> 10\text{M}\Omega$
Output Resistance R_o	0	$< 500\Omega$
Input Current $\delta I_\pm$	0	$< 0.5\mu$A
Input Offset Voltage $\delta V_\pm$	0	< 10mV
Input Offset Current δI_i	0	< 200pA

What are the conditions that dictate the range of the feedback impedances R_f? Apart from special cases, the feedback current I should be only a small fraction of the maximum output current I_o, i.e. $I = 1\% I_o$. A typical Op-Amp has a maximum output of 10mA at 10V, i.e.

$$R_f = \frac{V}{I} = \frac{10\text{V}}{10 \cdot 1\%\text{mA}} = 100\text{k}\Omega$$

Typical feedback resistors should be in the range of $R_f = 50 - 1\text{M}\Omega$.

Small difference on the differential stage of the Op-Amp produces a DC offset δV at the input, which can produce large DC output if the gain is extremely high. For example, if we have

$$\delta V = 10\text{mV}, \quad G \geq 10^4, \Rightarrow V_o \geq 10V.$$

4.5 Problems Preparatory to the Laboratory

1. Supposing that the open-loop gain of an Op-Amp is a simple low pass filter (first order) with DC gain 144dB and cut-off frequency $f_0 = 10$Hz, sketch in a bode diagram, the magnitude of the frequency response of a non-inverting stage with gain $G = 10$ at 10Hz.

2. Prove that the good integrator condition for the circuit in figure 4.6 is $\omega \gg 1/R_f C_f$. (Hint: calculate the response of the circuit and compare it with the ideal response of the ideal inverting integrator, i.e. $A(\omega) = -1/(j\omega RC)$. Calculate the DC gain of the integrator transfer function.

3. Using an integrator stage with a feedback resistor R_f, and time constant $\tau = RC$, compute the values for τ and R_f needed to integrate a sinusoidal wave with frequency $f > 1$kHz and with 10% of losses in the integration. Choose a value of $R \gg R_s$, where $R_s = 50\Omega$ is the input impedance of the used function generator.

4. Show that the slope of an integrated square wave is the inverse of of the time constant $\tau = RC_f$ of the Integrator shown in figure 4.6. Which characteristics of the square wave are needed to fulfill the requirement to not saturate the integrator output ?

5. Consider a differential stage having the following resistances values: $R_1 = R_2 = 50\,\text{k}\Omega$, $R_f = R_0 = 100\ \text{k}\Omega$. Calculate the following quantities

   (a) the two input impedances $Z_1$and Z_2 ,

   (b) the output impedance Z_o,

   (c) Considering that the Op-Amp max output current and voltage are respectively I_{max} =10mA and $V_{max} = 10$V, calculate the smallest load R_{min} it can drive .

4.6 Laboratory Procedure

Read carefully the entire procedure before starting the experiment, and note on your log book all the unpredicted behavior you experience in the circuits response.

Consult the data-sheet to properly map the μ741 and AD711 Op-Amp pin-out.

Op-Amp output high frequency noise can be reduced by adding 100nF capacitors closest as possible to the $\pm 15V$ power supplies input of the Op-Amp.

Before powering your circuit up, cross-check the power supply connections.

It is always a good practice to turn on the power supplies at the same time to avoid potential damages of the Op-Amps.

Using the μ741 Op-Amp, do the following steps:

- Using a non-inverting configuration with a gain of 100, verify the transfer function of the Op-Amp.
- Using the same previous circuit, estimate the slew rate of the Op-Amp. Redo the same measurement using an AD711 Op-Amp.
- Study the $CMRR$ using a differential configuration. Use a potentiometer to balance the gains at just one frequency and then measure the $CMRR$. Verify that the obtained values are in agreement with the specifications reported in the Op-Amp data-sheet. Mount and tune the null adjustment circuit as specified in the Op-Amp data-sheet.

Build an integration stage using an Op-Amp having a time constant $\tau \sim 100\mu s$. Include a feedback resistor R_f to avoid saturation at the output and do the following steps:

- Measure the impulse response.
- Measure the frequency response.
- Estimate the integrator time constant τ using a square wave.

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