Introduction to Operational Amplifier Design

vers 1.0 October 22, 2007

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1 Preface

The scope of this course is the design of basic voltage feedback operational amplifier circuits. Using the ideal op amp model and solving for the currents and voltages at each terminal we get the transfer function as a Laplace Transform. This course provides a practical way of going from paper design to prototyping working circuits.

This course is intended for professional electrical engineers. The course-taker should be familiar with the Laplace and inverse Laplace Transforms and basic AC network analysis.

After completing the course, there is a quiz consisting of 16 multiple choice questions. On completion, 4 professional development hours will count towards satisfying PE licensure renewal requirements.

Navigating the course is facilitated by hyperlinked table of contents on each page or the tags in the bookmark pane.

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2 Introduction

The voltage feedback differential amplifier ("op amp" as it is called) is used in a wide variety of electronic applications such as: linear amplifier/attenuator, signal conditioner, signal synthesizer, computer, or simulator.

A practical way to approach designing and implementing an op amp circuit is to start with the ideal model and get an expression that relates the output to the input, regardless of the input. This is accomplished by working with the loop equations in the frequency or s domain [1].

In summary, the key to getting the transfer function is that the voltages at the input terminals of a closed-loop op amp circuit mirror each other. Also, no current flows into or out of an input terminal. When a signal is applied to either or both terminals, the output will adjust itself to meet these constraints.

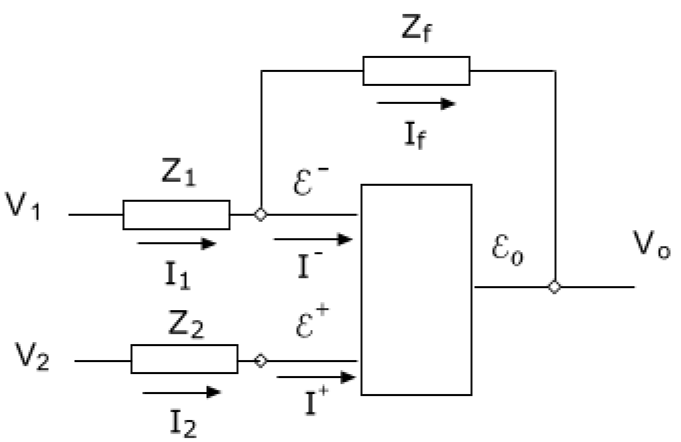

Figure 3.1-1 Loop analysis example

Using the figure to the left :

if V_2 and $Z_2 = 0$ then $\mathcal{E}^+ = 0 \therefore \mathcal{E}^- = 0$

We know that no current flows into an input terminal so $I^+ = I^- = 0 \therefore I_1 = I_f$ and $I_2 = 0$

$I_1 = \frac{V_1 - \mathcal{E}^-}{Z_1} = \frac{V_1}{Z_1}$ and $I_f = \frac{\mathcal{E}^- - V_o}{Z_f} = -\frac{V_o}{Z_f}$

so $I_1 = I_f$ gives us $\frac{V_1}{Z_1} = -\frac{V_o}{Z_f}$

hence $\frac{V_o}{V_1} = -\frac{Z_f}{Z_1}$

The closed loop transfer function is in the frequency domain, $A_v(s) = \frac{V_o(s)}{V_1(s)} = -\frac{Z_f}{Z_1}$.

For the remainder of the course, we'll use the shorter notation as, $A_v = \frac{V_o}{V_1} = -\frac{Z_f}{Z_1}$.

Also note : in AC network analysis, impedances are represented as phasors and do not vary with time but with frequency. So there is no time domain representation or time variation of an impedance.

In the following sections, the same method is used for application specific circuits where the voltages and impedances are arbitrary.

[1] To get the output in the time domain $v_o(t)$ we would have to multiply $V_i(s)$ by $A_V(s)$ and then take the inverse Laplace Transform ; $v_o(t)=\mathcal{L}^{-1}[\, V_i(s) \cdot A_V(s)\,]$.

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3 The Ideal Voltage Feedback Op Amp

The voltage feedback op amp is a discrete device that has 2 input terminals and one output terminal. Without feedback, the output is the difference between the input voltages, multiplied by the open-loop gain (transfer function) of the op amp.

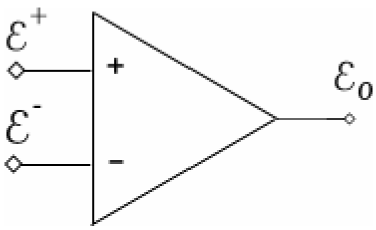


Figure 3.1-1 Ideal operational amplifier

The open loop equations :

$$\left(\mathcal{E}^+ - \mathcal{E}^-\right) \cdot A_{vol} \equiv \mathcal{E}_o \quad \text{where} \quad A_{vol} = \infty$$

$$\left(\mathcal{E}^+ - \mathcal{E}^-\right) \equiv \frac{\mathcal{E}_o}{A_{vol}} \equiv 0 \quad \text{so} \quad \mathcal{E}^+ - \mathcal{E}^- \equiv 0$$

$$\therefore \quad \mathcal{E}^+ \equiv \mathcal{E}^-$$

$\mathcal{E}^+$ the non-inverting terminal

$\mathcal{E}^-$ the inverting terminal

$\mathcal{E}_o$ the output terminal

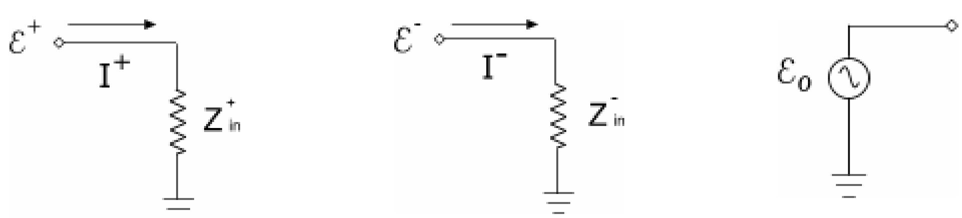
A_{vol} is the open loop gain (transfer function) of the op amp itself, without feedback

The key is in maintaining $\mathcal{E}^+ = \mathcal{E}^-$ otherwise the output will saturate:

If we apply a voltage at $\mathcal{E}^+$ and ground $\mathcal{E}^-$ $\mathcal{E}_o$ will saturate to positive supply voltage.

If we apply a voltage at $\mathcal{E}^-$ and ground $\mathcal{E}^+$ $\mathcal{E}_o$ will saturate to negative supply voltage.

In the open loop model, each input terminal has infinite impedance so no current can flow into an input terminal even with a voltage source or a ground applied. The output terminal has zero output impedance.



$\mathcal{E}^+$ has $Z^+_{in} = \infty$ so $I^+ = 0$ $\quad$ $\mathcal{E}^-$ has $Z^-_{in} = \infty$ so $I^- = 0$ $\quad$ $\mathcal{E}_o$ has $Z_{out} \equiv 0$

Figure 3.1-2 Ideal op amp input impedances

The ideal characteristics are summarized in Table 3-1 below

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3.1 Ideal Characteristics

Summary Of Ideal Characteristics	
$Z_{in} = \infty$	the input impedance at each terminal is infinite
$Z_{out} = 0$	the output impedance is zero
$I^{\pm} = 0$	no current flows into either of the input terminals
$A_{vol} = \infty$	the open loop gain is infinite
Bandwidth = ∞	the bandwidth is infinitely wide
No temperature drift	
$\mathcal{E}_o = 0$	the output voltage is zero when $\mathcal{E}^+ = \mathcal{E}^-$

Table 3-1 Summary of ideal characteristics

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3.2 Ideal Model with Feedback

With feedback, all or a portion of the output is tied to either or both input terminals. The difference between the voltages at the input terminals is still equal to zero and again no current flows into either of the input terminals.

The voltage at each input terminal is treated as a node and is determined by the loop equations. The terminal voltages are equated and we get a transfer function or closed loop gain which provides the output over the input as a ratio (Laplace Transform).

In the following subsections, we'll look at the feedback models : the non-zero reference and the zero reference feedback models. As the names imply, the non-zero reference model is characterized by the voltage at either input terminal, $\mathcal{E}^{\pm} \neq 0$, and the zero-reference model is where the voltage at either input terminal is referenced to 0 volts or ground; $\mathcal{E}^{\pm} = 0$.

3.2.1 The non-zero reference model

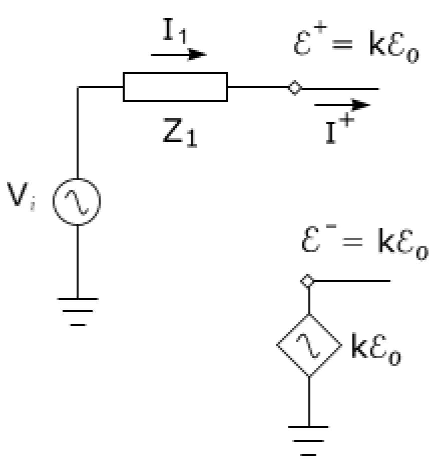

Figure 3.2-1 Ideal op amp feedback model 1

The figure to the left models a source V_i connected to the $\mathcal{E}^+$ terminal through a series impedance Z_1 and feedback $k\,\mathcal{E}_o$ tied to $\mathcal{E}^-$; with k being a voltage divider $= \dfrac{Z_2}{Z_2 + Z_f}$

The voltage at each input terminal balances out to $k\,\mathcal{E}_o$

So $\mathcal{E}^+ = \mathcal{E}^- = k\,\mathcal{E}_o$ and we see that $I^+ = I_1$.

But $I^+ = 0$, giving us $I_1 = \dfrac{V_i - \mathcal{E}^+}{Z_1} = \dfrac{V_i - k\,\mathcal{E}_o}{Z_1} = 0$

Hence $\dfrac{\mathcal{E}_o}{V_i} = \dfrac{1}{k} = \dfrac{Z_2 + Z_f}{Z_2} = 1 + \dfrac{Z_f}{Z_2}$

The transfer function $A_v = \dfrac{\mathcal{E}_o}{V_i} = \left(1 + \dfrac{Z_f}{Z_2}\right)$

3.2.1.1 The input impedance

The input impedance seen by V_i is $Z_{in} = \dfrac{V_i}{I_{in}}$

The input current $I_{in} = I_1$, so $I_{in} = \dfrac{V_i - k\,\mathcal{E}_o}{Z_1}$ Since $kA_v = 1$, $V_i \cdot \left(\dfrac{1 - kA_v}{Z_1}\right) = 0$

With $I_{in} = 0$, $Z_{in} = \infty$, so the input impedance seen by V_i is infinite or an open circuit.

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3.2.2 The zero-reference model

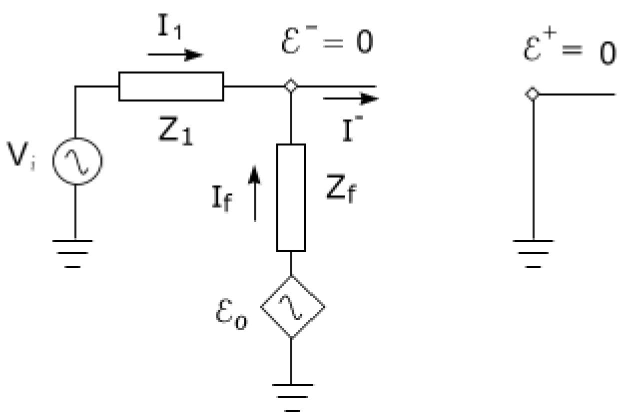

Figure 3.2-2 Ideal op amp feedback model 2

The figure to the left models a source V_i connected to the $\mathcal{E}^-$ terminal through a series impedance Z_1, with $\mathcal{E}^+$ tied to ground.

The voltage at each input must balance to 0 volts; thus $\mathcal{E}^{\pm} = 0$

We can see that $I_1 + I_f = I^-$; with $I^- = 0$, $I_1 = -I_f$

Substituting $I_1 = \dfrac{V_i - \mathcal{E}^-}{Z_1}$, $I_f = \dfrac{\mathcal{E}_o - \mathcal{E}^-}{Z_f}$ and $\mathcal{E}^- = 0$

into $I_1 = -I_f$ gives us $\dfrac{V_i}{Z_1} = -\dfrac{\mathcal{E}_o}{Z_f}$

So the transfer function $A_v = \dfrac{\mathcal{E}_o}{V_i} = -\dfrac{Z_f}{Z_1}$

3.2.2.1 The input impedance

The input impedance seen by V_i is $Z_{in} = \dfrac{V_i}{I_{in}}$; with $I_{in} = I_1 = \dfrac{V_i}{Z_1}$, $Z_{in} = \dfrac{V_i}{\frac{V_i}{Z_1}} = Z_1$

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3.3 Inverting or Non-inverting

The determining factors for whether the output is inverted or not, are the circuit configuration and the loop equations for the terminal currents and voltages. With voltage feedback op amp circuits, which terminal the signal is connected to, by itself, does not determine whether the output is inverted or not.

For example, the 2 circuits below are both inverting amplifier/attenuators circuits. The only difference is the op amp input terminals are reversed but both provide an output that is a linear amplifier or attenuator with 180° phase shift or inversion:

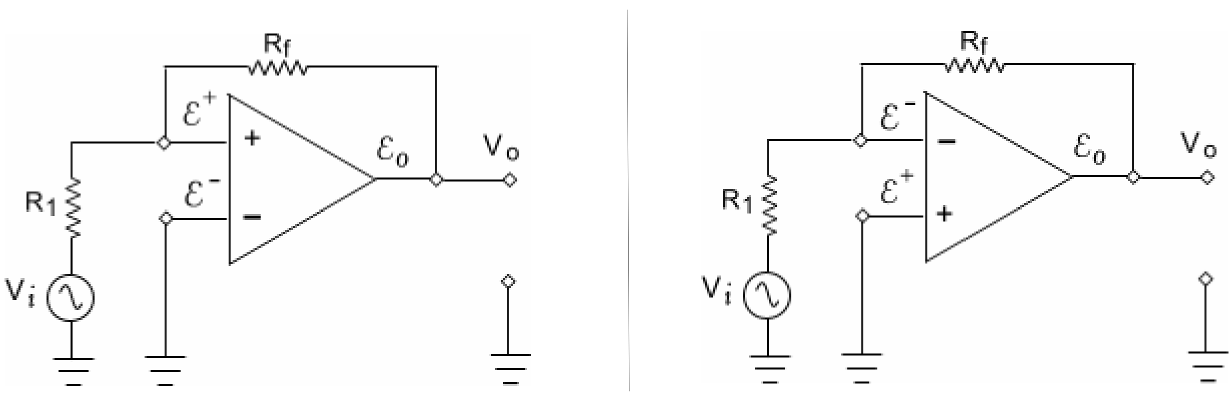

Figure 3.3-1 Non-inverting examples

$$\mathcal{E}^- = 0 \;\therefore\; \mathcal{E}^+ = 0$$

$$I_1 + I_f = I^+ \quad I^+ = 0 \;\therefore\; I_1 = -I_f$$

$$I_1 = \frac{V_i}{R_1} \qquad I_f = \frac{V_o}{R_f}$$

$$\therefore \frac{V_i}{R_1} = -\frac{V_o}{R_f} \text{ and } A_V = -\frac{R_f}{R_1}$$

$$\mathcal{E}^+ = 0 \;\therefore\; \mathcal{E}^- = 0$$

$$I_1 + I_f = I^- \quad I^- = 0 \;\therefore\; I_1 = -I_f$$

$$I_1 = \frac{V_i}{R_1} \qquad I_f = \frac{V_o}{R_f}$$

$$\therefore \frac{V_i}{R_1} = -\frac{V_o}{R_f} \text{ and } A_V = -\frac{R_f}{R_1}$$

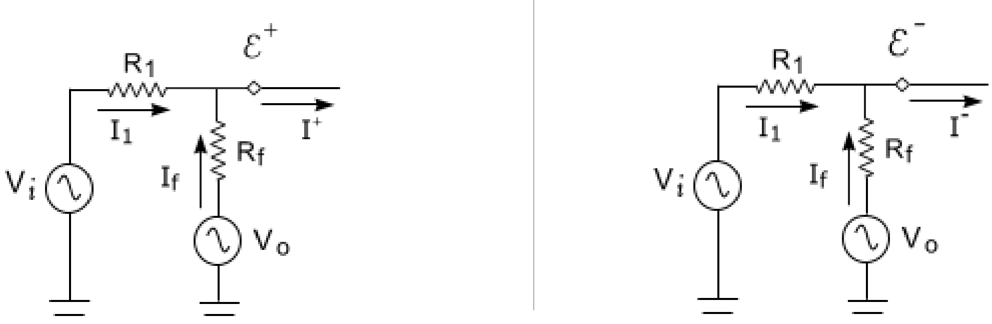

Figure 3.3-2 Loop equations for Fig 3.3-1

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3.4 About the Transfer Function

The transfer function gives you the output over the input expressed as a ratio. To get the output for a specific input, multiply the Laplace Transform of the input by the transfer function. To convert to the time domain, take the inverse Laplace Transform [2]. Using Laplace Transform tables available on the internet or in printed textbooks, is a very useful tool in op amp circuit design. Some online references are listed in Appendix B.

Working in the s domain (using Laplace Transforms) is advantageous and readily gives any transients that might exist

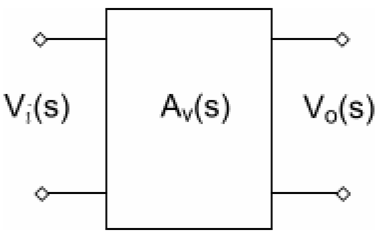
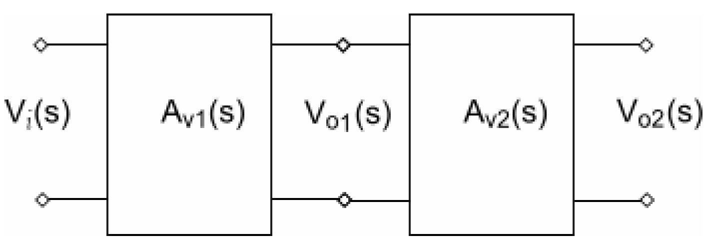


Figure 3.4-1 Black box diagram



$$A_V(s)\frac{V_{o2}(s)}{V_i(s)} = A_{V1}(s)\cdot A_{V2}(s)$$

Figure 3.4-2 Cascaded transfer functions

$$A_V(s) \equiv \frac{V_o(s)}{V_i(s)} \quad \text{so } V_o(s) = V_i(s)\cdot A_V(s)$$

to get the time domain representation

$$v_o(t) = \mathcal{L}^{-1}\left[\, V_i(s)\cdot A_V(s)\,\right] \quad \text{or}$$

by using convolution :

$$v_o(t) = \int_0^t v_i(\tau)\cdot a_V(t-\tau)d\tau$$

When cascading circuits, the overall transfer function is the product of all the transfer functions in the cascade and the overall transfer function can be viewed as a single ratio or Laplace transform

In the time domain, you could not express the transfer function as a ratio. You would have to solve differential equations for the terminal voltages and currents and use the convolution integral to get the output as a function of time, because superposition does not apply.

[2] The time domain representation of Av (that is $a_v(t) = \mathcal{L}^{-1}[A_V(s)]$) is actually the unit impulse response of the circuit. The output $v_O(t)$ after applying an arbitrary input $v_i(t)$ is found by convolving $v_i(t)$ with $a_V(t)$ { $v_O(t) \equiv v_i(t)*a_V(t)$ } .

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3.5 About Input Impedances

An input source can be either a directly connected zero impedance source, or a thevenin equivalent source with an internal impedance. We need to be aware if a source does have internal impedance. This needs to be considered in determining both the input impedance seen by the voltage source and the transfer function of the circuit.

3.5.1 Directly connected source

Consider a directly-connected source, with zero output impedance, connecting to an op amp input terminal, through a series impedance Z_1 :

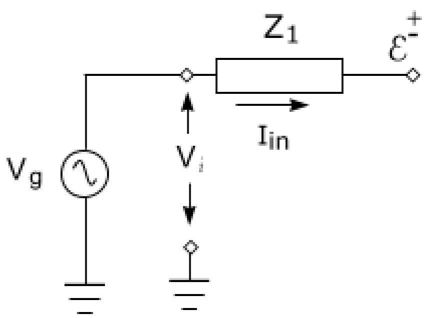

Figure 3.5-1 Zero impedance source

The input to the circuit is $V_i = V_g$.

In general, the transfer function A_v gives us $\frac{V_o}{V_g}$

When $\mathcal{E}^{\pm} = 0$, the input impedance seen by V_g is

$$Z_{in} = \frac{V_g}{I_{in}} \text{ ; with } I_{in} = \frac{V_g - \mathcal{E}^{\pm}}{Z_1} = \frac{V_g - 0}{Z_1}, \quad Z_{in} = \frac{V_g}{\frac{V_g}{Z_1}} = Z_1 .$$

When $\mathcal{E}^{\pm} \neq 0$, the input impedance seen by V_i is $Z_{in} = \frac{V_i}{I_{in}}$; with $I_{in} = \frac{V_i - \mathcal{E}^{\pm}}{Z_1}$, $Z_{in} = Z_1\left(\frac{V_i}{V_i - \mathcal{E}^{\pm}}\right)$

* It is interesting to note that in a case like this, we have a capability of synthesizing a voltage dependent input impedance that varies with V_i and $\mathcal{E}^{\pm}$.

3.5.2 Source with built-in internal impedance

Consider the case where the input is the end of a cable span or an input from a previous circuit stage in the cascade:

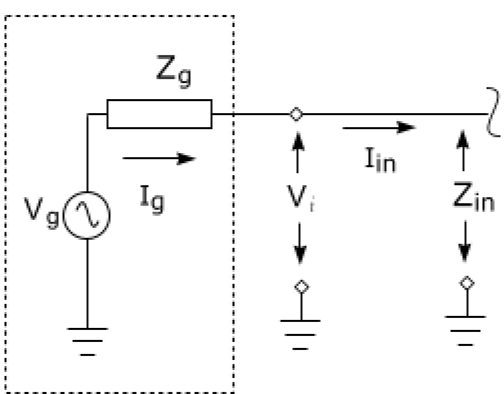

Figure 3.5-2 Source with internal impedance

Now we have a source V_g with internal impedance Z_g.

We have to modify our calculations because $V_i = V_g\left(\frac{Z_{in}}{Z_{in} + Z_g}\right)$.

Substituting into the transfer function A_v, we get

$$\frac{V_o}{V_i} = \frac{V_o}{V_g}\left(\frac{Z_{in} + Z_g}{Z_{in}}\right) = \frac{V_o}{V_g}\left(1 + \frac{Z_g}{Z_{in}}\right)$$

the total impedance seen by V_i is Z_{in}

the total impedance seen by V_g is $Z_g + Z_{in}$

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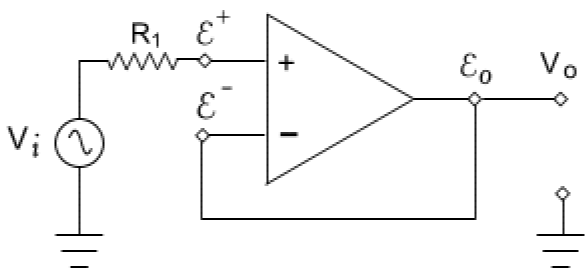
4 Linear Amplifiers and Attenuators

The most basic circuit configurations are linear amplifiers and attenuators. Ideally, they provide flat gain or loss across the device-rated bandwidth.

4.1 Voltage Follower

The voltage follower circuit provides unity gain with no inversion. This circuit is used to isolate a high impedance source input and provide a buffered output.



Circuit 4-1 Voltage Follower

For Circuit 4 - 1, the terminal voltages are:

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = V_i$

$\mathcal{E}^- = \mathcal{E}_o$

Since $\mathcal{E}^+ = \mathcal{E}^-$, $V_i = V_o$ giving us $\frac{V_o}{V_i} = 1$.

So $A_V = 1$ is the transfer function for Circuit 4.1.

4.1.1 The Input impedance

The input impedance seen by source V_i is $Z_{in} = \frac{V_i}{I_{in}}$ where $I_{in} = \frac{V_i - \mathcal{E}^+}{R_1}$

Since $\mathcal{E}^+ = V_i$, $I_{in} = 0$, making $Z_{in} = \frac{V_i}{0} = \infty$

So the input impedance seen by V_i for Circuit 4 - 1 is an open circuit.

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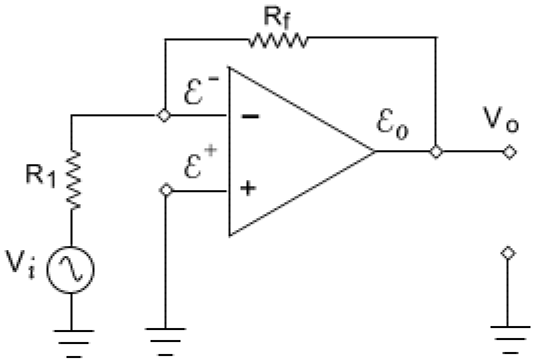
4.2 Inverting Amplifier or Attenuator

The inverting configuration provides flat gain or attenuation with constant 180° phase shift.



Circuit 4-2 Inverting Amplifier or Attenuator

For Circuit 4 - 2, the terminal voltages are :

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = 0$

$\mathcal{E}^- = \mathcal{E}^+ = 0$

* the terminal voltages balance out to zero

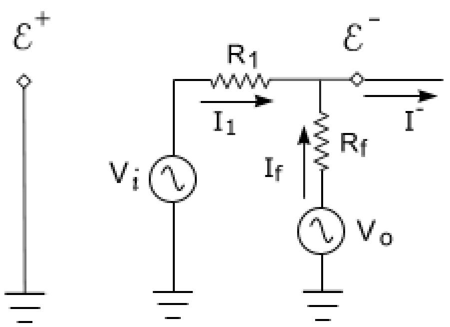


Figure 4.2-1 $\mathcal{E}^-$ and $\mathcal{E}^+$ loops for Circuit 4-2

From Figure 4.2 - 1 , for the $\mathcal{E}^-$ loop we have :

$$I_1 = \frac{V_i - \mathcal{E}^-}{R_1} \quad \text{and} \quad I_f = \frac{V_o - \mathcal{E}^-}{R_f}$$

With $I_1 + I_f = I^-$ and $I^- = 0$, we have $I_1 = -I_f$

Since the terminal voltage at $\mathcal{E}^- = 0$,

$$I_1 = \frac{V_i}{R_1} \quad \text{and} \quad I_f = \frac{V_o}{R_f} \quad \text{and with } I_1 = -I_f \text{ ,} \quad \frac{V_o}{V_i} = -\frac{R_f}{R_1}$$

So $A_V = -\dfrac{R_f}{R_1}$ is the transfer function for Circuit 4.2

4.2.1 The Input impedance

The input impedance seen by source V_i is , $Z_{in} = \dfrac{V_i}{I_{in}}$; where $I_{in} = I_1 = \dfrac{V_i}{R_1}$

Substituting I_{in} into the expression for Z_{in} gives us $Z_{in} = \dfrac{V_i}{\frac{V_i}{R_1}} = R_1$

So the input impedance seen by V_i for Circuit 4 - 2 is R_1

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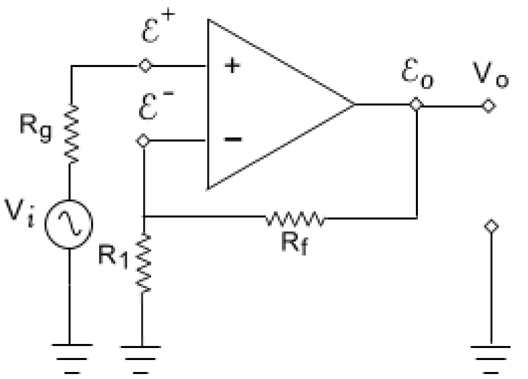
4.3 Non-Inverting Amplifier

The non-inverting configuration provides flat gain with no phase shift.



Circuit 4-3 Non-inverting Amplifier

For Circuit 4 - 3, the terminal voltages are :

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = V_i$

$\mathcal{E}^- = kV_o$; where k is a voltage divider $= \dfrac{R_1}{R_1 + R_f}$

Since $\mathcal{E}^+ = \mathcal{E}^-$, $kV_o = V_i$; readily giving us $\dfrac{V_o}{V_i} = \dfrac{1}{k}$

So by inspection, we see that $A_V = \dfrac{1}{k} = \left(1 + \dfrac{R_f}{R_1}\right)$.

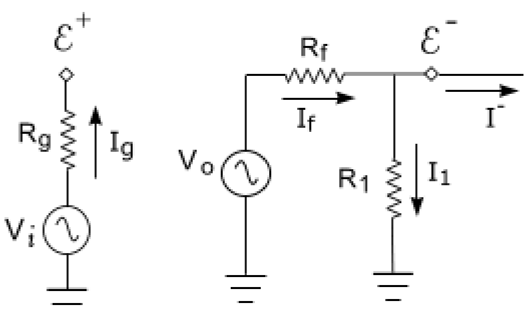


Figure 4.3-1 $\mathcal{E}^-$ and $\mathcal{E}^+$ loops for Circuit 4-3

Analysis of the $\mathcal{E}^-$ loop gives us the same result :

From Fig 4.3 - 1 we see that $I_1 = \dfrac{\mathcal{E}^-}{R_1}$ and $I_f = \dfrac{V_o - \mathcal{E}^-}{R_f}$

With $I_f = I_1 + I^-$ and $I^- = 0$, $I_f = I_1$

Since $\mathcal{E}^- = kV_o$, $\mathcal{E}^+ = V_i$ and $\mathcal{E}^- = \mathcal{E}^+$,

we substitute $\mathcal{E}^- = V_i$, into the expressions for I_1 and I_f

giving us $I_1 = \dfrac{V_i}{R_1}$ and $I_f = \dfrac{V_o - V_i}{R_f}$

With $I_f = I_1$ we get $\dfrac{V_i}{R_1} = \dfrac{V_o - V_i}{R_f}$ $\therefore$ $V_i \cdot \left(\dfrac{1}{R_1} + \dfrac{1}{R_f}\right) \cdot R_f = V_o$; giving us

$A_V = \dfrac{V_o}{V_i} = \left(1 + \dfrac{R_f}{R_1}\right)$ as the transfer function for Circuit 4 - 3.

4.3.1 The Input impedance

The input impedance seen by source V_i is $Z_{in} = \dfrac{V_i}{I_{in}}$ From Figure 4.3 - 1, $I_{in} = I_g$ where $I_g = \dfrac{V_i - \mathcal{E}^+}{R_g}$

Substituting $\mathcal{E}^+ = kV_o$ into I_{in} gives us $I_{in} = \dfrac{V_i - kV_o}{R_g}$ With $kV_o = V_i$, $I_{in} = 0$ making $Z_{in} = \dfrac{V_i}{0} = \infty$

So the input impedance seen by V_i for Circuit 4.3 is an open circuit.

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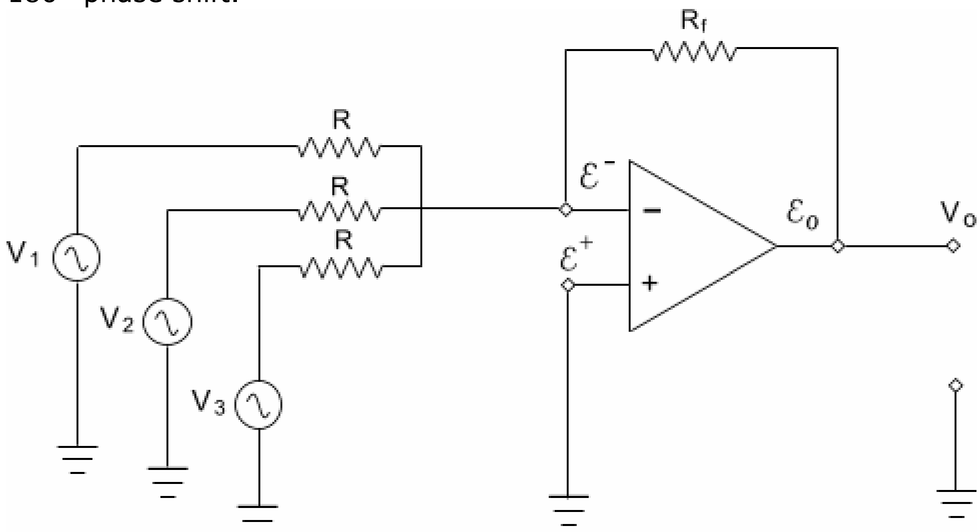
5 Summation Amplifier

The summation amplifier provides flat gain or loss and can be configured in an inverting or non-inverting configuration. Typical uses are signal combiner, voltage comparator or summing junction.

5.1 Inverting Summation Amplifier

The inverting summation amplifier provides flat gain or loss across the rated bandwidth with constant 180° phase shift.



For Circuit 5 - 1,
the terminal voltages are:

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = 0$ so $\mathcal{E}^- = 0$

* the terminal voltages balance out to zero

Circuit 5-1 Inverting Summation Amplifier

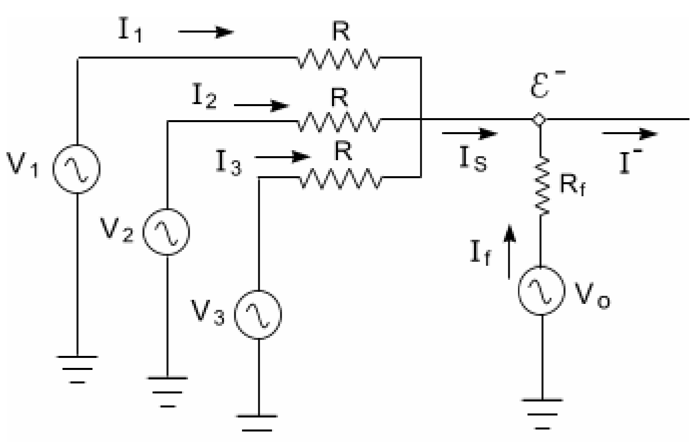


Figure 5.1-1 $\mathcal{E}^-$ loop for Circuit 5-1

From Figure 5.1 - 1, for the $\mathcal{E}^-$ loop we have:

$I_S = I_1 + I_2 + I_3$ where

$$I_1 = \frac{V_1 - \mathcal{E}^-}{R}\ ,\ I_2 = \frac{V_2 - \mathcal{E}^-}{R}\ ,\ I_3 = \frac{V_3 - \mathcal{E}^-}{R}$$

and $I_f = \dfrac{V_o - \mathcal{E}^-}{R_f}$

Since the terminal voltage at $\mathcal{E}^- = 0$

$$I_1 = \frac{V_1}{R},\ I_2 = \frac{V_2}{R},\ I_3 = \frac{V_3}{R} \text{ and } I_f = \frac{V_o}{R_f}$$

With $I^- = I_S + I_f$ and $I^- = 0$, $I_S = -I_f$

With $I_S = -I_f$, we have $\dfrac{V_1}{R} + \dfrac{V_2}{R} + \dfrac{V_3}{R} = -\dfrac{V_o}{R_f}$

Simplifying the expression for V_o , we get $V_o = -\dfrac{R_f}{R} \cdot (V_1 + V_2 + V_3)$

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To get the transfer function as a single ratio, we have to thevenize the 3 sources into one equivalent source V_i with thevenin impedance Z_{eq} ; giving us $V_i = \frac{V_1 + V_2 + V_3}{3}$ and $Z_{eq} = \frac{R}{3}$ as illustrated in the figure below.

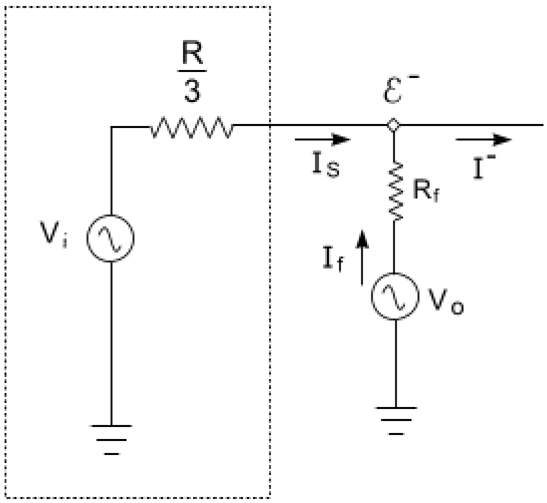


Figure 5.1-2 Thevenized input for Circuit 5-1

Analyzing the $\mathcal{E}^-$ loop in Figure 5.1 - 2 we have :

$$I_s = \frac{V_i - \mathcal{E}^-}{\frac{R}{3}} \quad \text{and} \quad I_f = \frac{V_o - \mathcal{E}^-}{R_f}$$

Substituting $\mathcal{E}^- = 0$ into I_s and I_f we have,

$$I_s = \frac{3V_i}{R} \quad \text{and} \quad I_f = \frac{V_o}{R_f}$$

With $I_s + I_f = I^-$ and $I^- = 0$, $I_s = -I_f$

$$\frac{3V_i}{R} = -\frac{V_o}{R_f} \quad \therefore \quad \frac{V_o}{V_i} = -\frac{3R_f}{R}$$

So $A_v = -\frac{3R_f}{R}$ is the transfer function for Circuit 5 - 1.

5.1.1 The Input impedance

By inspection, the input impedance seen by our thenevized source V_i is $Z_{in} = 0$ because, as Figure 5.1 - 2 illustrates, we are connecting a source with an internal impedance of $\frac{R}{3}$ directly to the $\mathcal{E}^-$ terminal.

However, the input impedance seen by each of the sources V_1 , V_2 and V_3 is $Z_n = \frac{V_n}{I_n}$; where

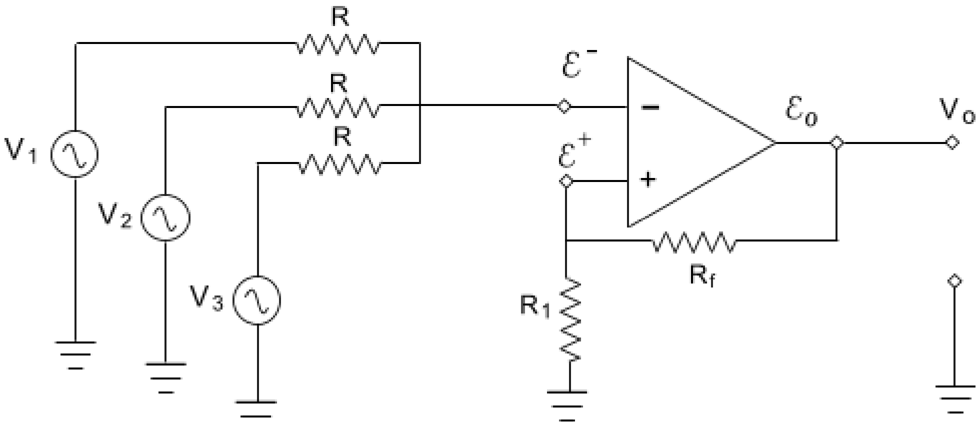
$I_n = \frac{V_n - \mathcal{E}^-}{Z_n}$ With $\mathcal{E}^- = 0$, Z_n is going to be whatever impedance is in series with each source.

For Circuit 5 - 1, each of the sources V_n sees $Z_n = R$ as the input impedance.

TOP

5.2 Non-inverting Summation Amplifier

The non-inverting summation amplifier provides flat gain across the rated bandwidth with no phase shift.



For Circuit 5 - 2,
the terminal voltages are:

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = k\mathcal{E}_o$

$k = \dfrac{R_1}{R_1 + R_f}$; a voltage divider

$\mathcal{E}^- = \mathcal{E}^+ = k\mathcal{E}_o$

Circuit 5-2 Non-inverting Summation Amplifier

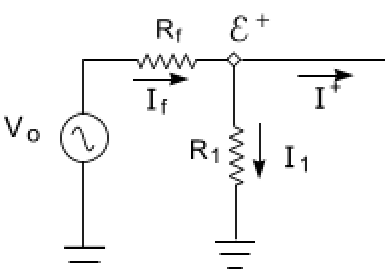


Figure 5.2-1 $\mathcal{E}^+$ loop for Circuit 5-2

From Figure 5.2 - 1, we see that the voltage at the $\mathcal{E}^+$ terminal is $\mathcal{E}_o$ through a voltage divider.

So $\mathcal{E}^+ = k\mathcal{E}_o$

Also, $I_f = I_1 + I^+$ and with $I^+ = 0$, $I_1 = I_f$

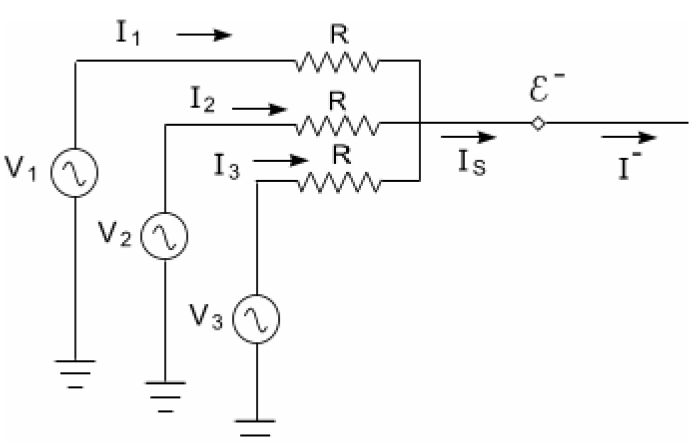


Figure 5.2-2 $\mathcal{E}^-$ loop for Circuit 5-2

From Figure 5.2 - 2, for the $\mathcal{E}^-$ loop we have:
$I_S = I_1 + I_2 + I_3$ where

$I_1 = \dfrac{V_1 - \mathcal{E}^-}{R}$, $I_2 = \dfrac{V_2 - \mathcal{E}^-}{R}$, $I_3 = \dfrac{V_3 - \mathcal{E}^-}{R}$

so $I_S = \dfrac{V_1}{R} + \dfrac{V_2}{R} + \dfrac{V_3}{R} - \dfrac{3 \cdot \mathcal{E}^-}{R}$

With $I^- = I_S$ and $I^- = 0$, $I_S = 0$

$$\therefore \frac{V_1}{R} + \frac{V_2}{R} + \frac{V_3}{R} = \frac{3 \cdot \mathcal{E}^-}{R}$$

TOP

Substituting $\mathcal{E}^- = kV_o$, we get $\frac{V_1}{R} + \frac{V_2}{R} + \frac{V_3}{R} = \frac{3 \cdot kV_o}{R}$ Substituting $k = \frac{R_1}{R_1 + R_f}$ we get,

$\left(\frac{V_1}{R} + \frac{V_2}{R} + \frac{V_3}{R}\right) = \frac{3V_o}{R} \cdot \left(\frac{R_1}{R_1 + R_f}\right)$ Simplifying we get, $V_o = \frac{1}{3}(V_1 + V_2 + V_3) \cdot \left(1 + \frac{R_f}{R_1}\right)$

To get the transfer function in the form of a ratio, we have to thevenize the 3 sources into one equivalent source V_i with thevenin impedance Z_{eq} ; so $V_i = \frac{V_1 + V_2 + V_3}{3}$ and $Z_{eq} = \frac{R}{3}$

From Figure 5.2 - 3 below, we look at the loop equations for the $\mathcal{E}^-$ terminal.

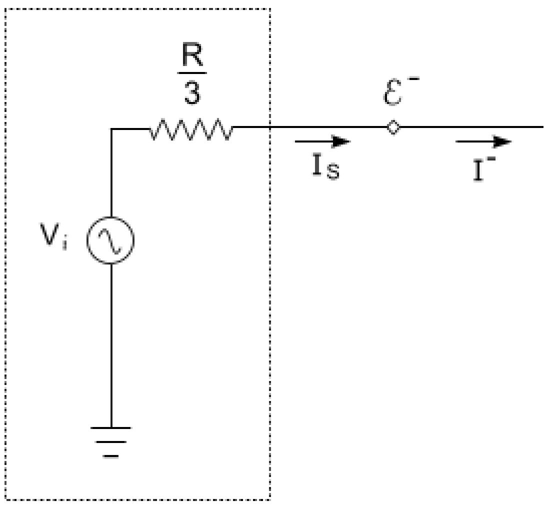


Figure 5.2-3 Thevenized input for Circuit 5-2

We have :

$I_S = I^- = 0$

with $I_S = \frac{V_i - \mathcal{E}^-}{\frac{R}{3}}$, substitute $\mathcal{E}^- = kV_o$

Expanding I_S we have ; $I_S = \frac{V_i - kV_o}{\frac{R}{3}} = 3 \cdot \left(\frac{V_i - kV_o}{R}\right) = 0$

giving us $V_i = kV_o$

Substituting $k = \left(\frac{R_1}{R_1 + R_f}\right)$, we get $\frac{V_o}{V_i} = \left(1 + \frac{R_f}{R_1}\right)$, so

$A_v = \left(1 + \frac{R_f}{R_1}\right)$ is the transfer function for Circuit 5 - 2.

5.2.1 The Input impedance

The input impedance seen by V_i is $Z_{in} = \frac{V_i}{I_{in}}$ where $I_{in} = I_S = 3 \cdot \left(\frac{V_i - kV_o}{R}\right)$

Substituting $kV_o = V_i$ into I_{in} gives us $I_{in} = 0$ making $Z_{in} = \frac{V_i}{0} = \infty$

The input impedance for Circuit 5 - 2 is an open circuit.

The input impedance seen by each source is going to be voltage - dependent.

We know that $V_o = \frac{1}{3k}(V_1 + V_2 + V_3)$, where $k = \left(\frac{R_1}{R_1 + R_f}\right)$

TOP

We also know that the voltage at $\mathcal{E}^{-} = kV_o$. The input impedance seen by each source is $Z_{in_n} = \frac{V_n}{I_n}$

where $I_n = \frac{V_n - \mathcal{E}^{-}}{Z_n}$ and Z_n is the series impedance of each source.

For example;

In Circuit 5 - 2, source V_1 has a series impedance of R so the input current $I_1 = \frac{V_1 - \mathcal{E}^{-}}{R}$

Substituting $\mathcal{E}^{-} = \frac{1}{3}(V_1 + V_2 + V_3)$, $I_1 = \frac{V_1 - \frac{1}{3}(V_1 + V_2 + V_3)}{R} = \frac{2V_1 - (V_2 + V_3)}{3R}$

so the input impedance seen by source V_1 is $Z_{in_1} = \frac{V_1}{I_1} = \frac{V_1}{\frac{2V_1 - (V_2 + V_3)}{3R}}$ which simplifies to

$$Z_{in_1} = 3R \cdot \left(\frac{V_1}{2V_1 - (V_2 + V_3)} \right)$$

So Z_{in_1} is an open circuit when $V_1 - \frac{(V_2 + V_3)}{2}$, and is a negative impedance when $V_1 < \frac{(V_2 + V_3)}{2}$

A negative impedance means V_1 is drawing current from the other sources and the output.

To get a nominal value for Z_{in_1}, we have to know something about the harmonic content and magnitude coefficients of the other sources so we can calculate an average.

The same calculations apply to getting the input impedances seen by the other sources at the input.

$$Z_{in_n} = 3R \cdot \left(\frac{V_n}{2V_n - \sum V_m} \right)$$

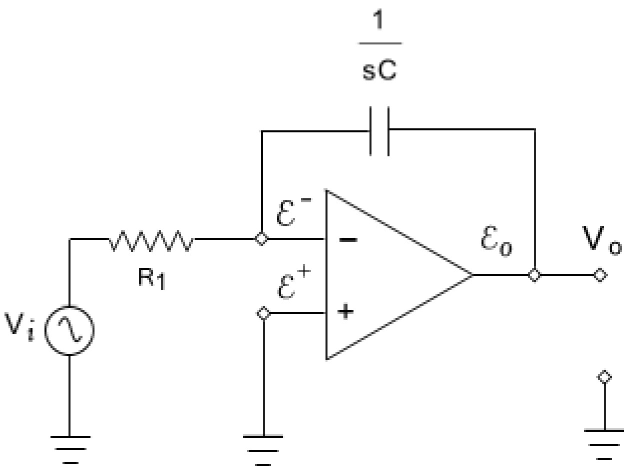
TOP

6 The Integrator

Typical uses of the integrator circuit are: noise reduction, simulation of a first order RC low-pass circuit, low pass filter, calculating an integral or just phase compensation.

6.1 Inverting Integrator

The inverting integrator provides low-pass filtering with constant +90° phase shift. The roll-off starts at DC. Typical use is direct integration of a time domain function, noise reduction, or low pass filter.



Circuit 6-1 Inverting Integrator

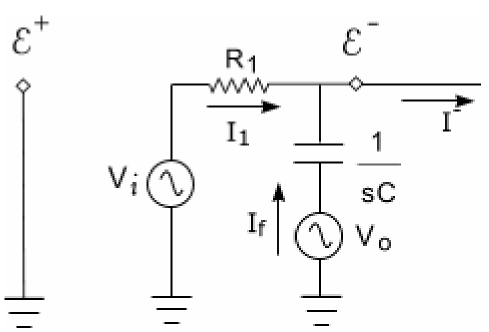


Figure 6.1-1 $\mathcal{E}^-$ and $\mathcal{E}^+$ loops for Circuit 6-1

For Circuit 6 - 1, the terminal voltages are:

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = 0$

$\mathcal{E}^- = \mathcal{E}^+ = 0$

* the terminal voltages balance out to zero

From Figure 6.1 - 1, for the $\mathcal{E}^-$ loop we have:

$$I_1 = \frac{V_i - \mathcal{E}^-}{R_1}\ ,\text{ and }\ I_f = \frac{V_o - \mathcal{E}^-}{\frac{1}{sC}} = sC\left(V_o - \mathcal{E}^-\right)$$

Substituting $\mathcal{E}^- = 0$ into I_1 and I_f gives us

$$I_1 = \frac{V_i}{R_1}\ ,\ \text{ and }\ I_f = sCV_o$$

With $I_f + I_1 = I^-$ and $I^- = 0$, $I_f = -I_1$ so

$$sCV_o = -\frac{V_i}{R_1}\ \therefore\ \frac{V_o}{V_i} = -\frac{1}{sCR_1}$$

$A_V = -\frac{1}{sCR_1}$ is the transfer function for Circuit 6 - 1.

6.1.1 The Input impedance

The input impedance Z_{in} seen by V_i is $Z_{in} = \frac{V_i}{I_{in}}$ in this case $I_{in} = I_1 = \frac{V_i}{R_1}$

With $I_{in} = \frac{V_i}{R_1}$ $\quad Z_{in} = \frac{V_i}{\frac{V_i}{R_1}} = R_1$

The input impedance seen by source V_i for Circuit $6 - 1$ is R_1.

TOP

6.1.2 Bandwidth Considerations

For Circuit 6 – 1, $A_v = -\frac{1}{sCR_1}$; substituting $s = j\omega$ gives us the value of $A_V(s)$ at $s = j\omega$.

We see at $s = 0$ (DC), $A_v = -\infty$ which means the output of the circuit will be saturated at the negative supply voltage, or 0 volts if we are using a single supply. We may want to offset this if we are working with input signals that have DC components.

See Section 8.3.2 for offsetting the baseline.

We can rewrite $A_v = -\frac{1}{j\omega CR_1}$ as $A_v = |A_v| \cdot e^{j\frac{\pi}{2}}$ where $|A_v| = \frac{1}{\omega CR_1}$; noting that the constant 90° phase shift is independent of ω. If we start the design at a fundamental frequency ω_0 , then we would select CR_1 to give us the desired gain magnitude at ω_0; i.e $CR_1 = \frac{1}{\omega_0 \cdot |A_v|}$

Our - 3dB bandwidth is the ω at which $|A_v(\omega)| = \frac{1}{2}|A_v(\omega_0)|$

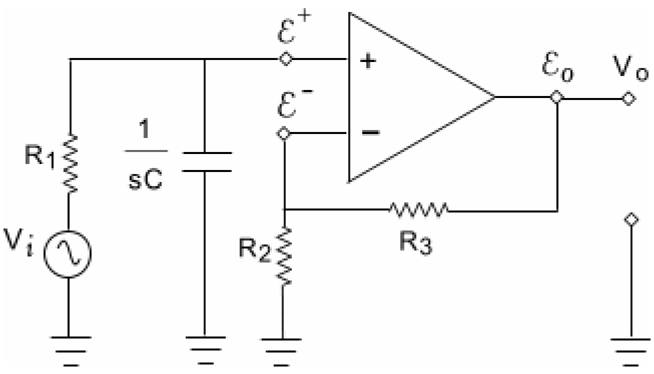
This gives us the direct relationship $\frac{|A_v(\omega)|}{|A_v(\omega_0)|} = \frac{\omega_0}{\omega} = \frac{1}{2}$

So our - 3dB bandwidth occurs at $\omega = 2\omega_0$

TOP

6.2 Non-inverting Low-pass Filter

The non-inverting low-pass filter provides filtering with phase shift that varies with the complex transfer function.



Circuit 6-2 The Non-inverting Integrator

For Circuit 6 - 2, the terminal voltages are:

$V_o = \mathcal{E}_o$

$$\mathcal{E}^+ = V_i\left(\frac{\frac{1}{sC}}{R_1 + \frac{1}{sC}}\right) = V_i\left(\frac{1}{sCR_1 + 1}\right)$$

$\left(\frac{1}{sCR_1+1}\right)$ being a voltage divider

$\mathcal{E}^- = kV_o$ where $k = \frac{R_2}{R_2 + R_3}$

k being a voltage divider

With $\mathcal{E}^+ = \mathcal{E}^-$, $V_i\left(\frac{1}{sCR_1+1}\right) = kV_o$ so $\frac{V_o}{V_i} = \frac{1}{k}\left(\frac{1}{sCR_1+1}\right) = \left(1+\frac{R_3}{R_2}\right)\cdot\left(\frac{1}{sCR_1+1}\right)$

$A_v = \frac{V_o}{V_i} = \left(1+\frac{R_3}{R_2}\right)\cdot\left(\frac{1}{sCR_1+1}\right)$ is the transfer function for Circuit 6 - 2.

6.2.1 The input impedance

To calculate the input impedance seen by V_i we analyze the $\mathcal{E}^+$ loop.

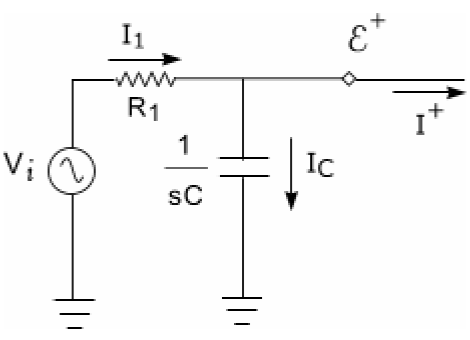


Figure 6.2-1 $\mathcal{E}^+$ loop for Circuit 6-2

The input impedance seen by source V_i is $Z_{in} = \frac{V_i}{I_{in}}$

For Circuit 6 - 2, $I_{in} = I_1$

From Figure 6.2 - 1, we have $I_1 = \frac{V_i - \mathcal{E}^+}{R_1}$

Recall $\mathcal{E}^+ = \frac{V_i}{sCR_1 + 1}$ so $I_1 = \frac{V_i - \frac{V_i}{sCR_1+1}}{R_1}$

I_1 simplifies to $V_i \cdot \left(\frac{sC}{sCR_1+1}\right)$ so $Z_{in} = \frac{V_i}{V_i \cdot \left(\frac{sC}{sCR_1+1}\right)} = \frac{sCR_1+1}{sC} = R_1 + \frac{1}{sC}$

The input impedance seen by source V_i for Circuit 6 - 2 is $R_1 + \frac{1}{sC}$.

TOP

6.2.2 Bandwidth considerations

Circuit 6 - 2 is a first order low - pass filter with a fundamental frequency of $\omega_0 = 0$.

Previously, we found the transfer function to be $A_v = \frac{V_o}{V_i} = \left(1 + \frac{R_3}{R_2}\right) \cdot \left(\frac{1}{sCR_1 + 1}\right)$; let's replace $\left(1 + \frac{R_3}{R_2}\right)$

with $\frac{1}{k}$ so $A_v = \frac{1}{k} \cdot \left(\frac{1}{sCR_1 + 1}\right)$ Rewriting A_v in phasor form, and substituing $s = j\omega$ gives us $A_v = |A_v| \cdot e^{-j\theta}$

where $|A_v| = \frac{1}{k} \cdot \left(\frac{1}{\sqrt{(\omega CR_1)^2 + 1}}\right)$ and $\theta = \tan^{-1} \omega CR_1$

We see that at ω_0, $|A_v(\omega_0)| = \frac{1}{k}$ and $\theta = 0$ radians

The - 3dB bandwidth is where $|A_v(\omega)| = \frac{1}{2} \cdot |A_v(\omega_0)|$ or $\frac{|A_v(\omega)|}{|A_v(\omega_0)|} = \frac{1}{2}$

Since $|A_v(\omega_0)| = \frac{1}{k}$ and $|A_v(\omega)| = \frac{1}{k} \cdot \left(\frac{1}{\sqrt{(\omega CR_1)^2 + 1}}\right)$, $\frac{|A_v(\omega)|}{|A_v(\omega_0)|} = \frac{1}{\sqrt{(\omega CR_1)^2 + 1}} = \frac{1}{2}$

The - 3dB point is where $\frac{1}{\sqrt{(\omega CR_1)^2 + 1}} = \frac{1}{2}$ or where $\omega = \frac{\sqrt{3}}{CR_1}$

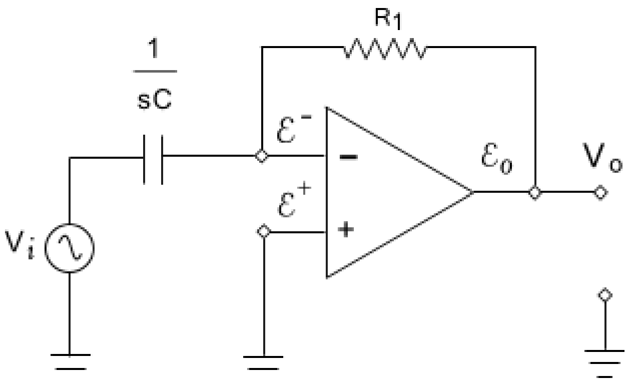
TOP

7 The Differentiator

Typical uses of the differentiator circuit are: simulation of a first order RC high-pass circuit, high pass filter, calculating a derivative or just phase compensation.

7.1 The Inverting Differentiator

The inverting differentiator provides high-pass filtering with constant - 90° phase shift. The roll-off starts at DC. Typical use is direct differentiation of a time domain function, or high pass filter.



Circuit 7-1 Inverting Differentiator

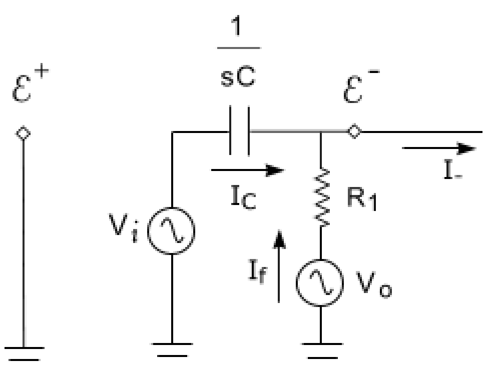


Figure 7.1-1 $\mathcal{E}^-$ and $\mathcal{E}^+$ loops for Circuit 7-1

For Circuit 7 - 1, the terminal voltages are:

$V_o = \mathcal{E}_o$

$\mathcal{E}^+ = 0$

$\mathcal{E}^- = \mathcal{E}^+ = 0$

From Figure 7.1 - 1, for the $\mathcal{E}^-$ loop we have:

$I_C + I_f = I^-$; with $I^- = 0$, $I_C = -I_f$ where

$$I_C = \frac{V_i - \mathcal{E}^-}{\frac{1}{sC}} \quad \text{and} \quad I_f = \frac{V_o - \mathcal{E}^-}{R_1}$$

With the voltage at $\mathcal{E}^- = 0$,

$$I_C = \frac{V_i}{\frac{1}{sC}} \quad \text{and} \quad I_f = \frac{V_o}{R_1}$$

Substituting into $I_C = -I_f$,

we get $sCV_i = -\frac{V_o}{R_1} \therefore \frac{V_o}{V_i} = -sCR_1$

$A_V = \frac{V_o}{V_i} = -sCR_1$ is the transfer function for Circuit $7-1$.

7.1.1 The input impedance

The input impedance seen by V_i is $Z_{in} = \frac{V_i}{I_{in}}$ where $I_{in} = I_C$

with $I_{in} = sCV_i$ $Z_{in} = \frac{V_i}{sCV_i}$

The input impedance seen by source V_i for Circuit 7 - 1 is $\frac{1}{sC}$

TOP

7.1.2 Bandwidth considerations

For Circuit $7-1$, $A_v = -sCR_1$; substituting $s = j\omega$ gives us the value of $A_V(s)$ at $s = j\omega$.
We see at $s = 0$ (DC), $A_v = 0$ which means the output of the circuit will be 0 volts

We can rewrite $A_v = -j\omega CR_1$ as $A_v = |A_v| \cdot e^{-j\frac{\pi}{2}}$ where $|A_v| = \omega CR_1$; noting that the constant - 90° phase shift is independent of ω. If we start the design at a fundamental frequency ω_0 , we would select CR_1 to give us the desired gain magnitude at ω_0; i.e $CR_1 = \frac{|A_v|}{\omega_0}$

Our - 3dB bandwidth is the ω at which $|A_v(\omega)| = 2 \cdot |A_v(\omega_0)|$

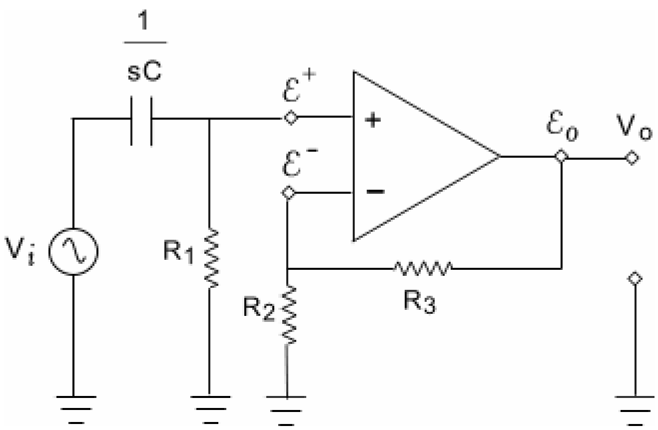
This gives us the direct relationship $\frac{|A_v(\omega)|}{|A_v(\omega_0)|} = \frac{\omega}{\omega_o} = 2$

So our - 3dB bandwidth occurs at $\omega = 2\omega_0$

TOP

7.2 Non-inverting High-pass Filter

The non-inverting differentiator provides high-pass filtering with frequency-dependent phase shift. Typical use is direct integration of a time function, or noise reduction. The total phase shift is overall frequency dependent (with an imaginary zero and complex pole).



Circuit 7-2 Non-inverting Differentiator

For Circuit 7 - 2, the terminal voltages are:

$$V_o \equiv \mathcal{E}_o$$

$$\mathcal{E}^+ = V_i\left(\frac{R_1}{R_1 + \frac{1}{sC}}\right) = V_i\left(\frac{sCR_1}{sCR_1 + 1}\right)$$

$$\mathcal{E}^- = kV_o \quad \text{where } k = \frac{R_2}{R_2 + R_3}$$

$$\mathcal{E}^+ = \mathcal{E}^- \quad \text{so} \quad V_i\left(\frac{sCR_1}{sCR_1 + 1}\right) = kV_o$$

$$\frac{V_o}{V_i} = \frac{1}{k}\left(\frac{sCR_1}{sCR_1 + 1}\right) = \left(1 + \frac{R_3}{R_2}\right)\cdot\left(\frac{sCR_1}{sCR_1 + 1}\right)$$

$A_V = \frac{V_o}{V_i} = \left(1 + \frac{R_3}{R_2}\right)\cdot\left(\frac{sCR_1}{sCR_1 + 1}\right)$ is the transfer function for Circuit 7 - 2.

7.2.1 The input impedance

To calculate the input impedance seen by V_i, we look at the $\mathcal{E}^+$ loop

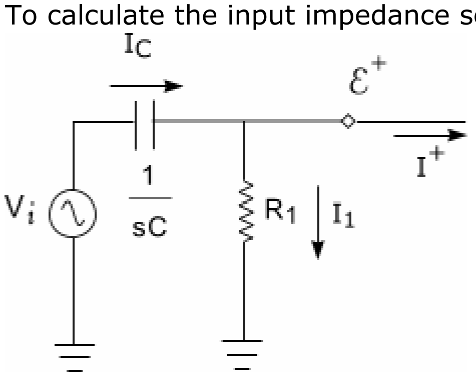


Figure 7.2-1 $\mathcal{E}^+$ loop for Circuit 7-2

The input impedance seen by source V_i is $Z_{in} = \frac{V_i}{I_{in}}$

For Circuit 7 - 2, $I_{in} = I_C$

Form Figure 7.2 - 1, we have $I_c = \frac{V_i - \mathcal{E}^+}{\frac{1}{sC}} = sC\left(V_i - \mathcal{E}^+\right)$

Recall $\mathcal{E}^+ = V_i \cdot \left(\frac{sCR_1}{sCR_1 + 1}\right)$ so $I_c = sC\left(V_i - V_i\left(\frac{sCR_1}{sCR_1 + 1}\right)\right)$

Simplifying, $I_c = sCV_i\left(1 - \left(\frac{sCR_1}{sCR_1 + 1}\right)\right) = V_i\left(\frac{sC}{sCR_1 + 1}\right)$

Substituting $I_{in} = I_C$ into $Z_{in} = \frac{V_i}{I_{in}}$ gives us $Z_{in} = \frac{V_i}{V_i \cdot \left(\frac{sC}{sCR_1 + 1}\right)} = R_1 + \frac{1}{sC}$

The input impedance seen by V_i for Circuit 7.2 is $R_1 + \frac{1}{sC}$

TOP

7.2.2 Bandwidth considerations

Circuit 7 - 2 is a first order high - pass filter and will not pass DC. If a unit step function is applied as an input signal the output will be a unit impulse $\partial(t)$.

So the fundamental frequency ω_0, is a low frequency just beyond DC and cannot be 0.
The + 3dB bandwidth will be approximated.

Previously, we found the transfer function to be $A_v = \frac{V_o}{V_i} = \left(1 + \frac{R_3}{R_2}\right) \cdot \left(\frac{sCR_1}{sCR_1 + 1}\right)$; let's replace $\left(1 + \frac{R_3}{R_2}\right)$ with $\frac{1}{k}$ so $A_v = \frac{1}{k} \cdot \left(\frac{sCR_1}{sCR_1 + 1}\right)$ Rewriting A_v in phasor form, and substituing $s = j\omega$ gives us $A_v = |A_v| \cdot e^{j\phi}$

where $|A_v| = \frac{1}{k} \cdot \left(\frac{\omega CR_1}{\sqrt{(\omega CR_1)^2 + 1}}\right)$, $\theta = \tan^{-1}\omega CR_1$ and $\phi = \left(\frac{\pi}{2} - \theta\right)$

The + 3dB bandwidth is approximately where $|A_v(\omega)| = 2$

So $|A_v(\omega)| = \frac{1}{k} \cdot \left(\frac{1}{\sqrt{(\omega CR_1)^2 + 1}}\right) = 2$ is approximately at $\omega = \frac{1}{R_1 C \cdot \sqrt{\left(4k^2 - 1\right)}}$

TOP

8 Practical Considerations

Device selection, component tolerances and power supply stability should be the priorities when designing and prototyping a circuit.

Practical considerations are device and application specific. Various types of op amps ranging from general purpose, low and high frequency, to application specific devices have their own specifications and recommendations for optimal performance.

8.1 Device Parameters

Some commonly used terms and their textbook definition are listed in Appendix A. Manufacturers may use the same terms, call them by other names or introduce new parameters to thoroughly characterize their op amp. Manufacturers may also include application notes on how best to make offset or compensation adjustments for voltage, current and phase. Most of these parameters are centered on an open loop mode.

So a circuit designer may not need to be concerned with all of the specifications when designing a circuit. It's important to select a device that exceeds expectations.

For example, the common mode rejection ratio is a value based on open loop gain and small, simultaneous signal variations on larger input levels; essentially nolise. A widely accepted minimum for the CMRR is 70dB. This value implies a level of stability that applies to a closed loop circuit and other parameters of the op amp. Less than 70dB indicates a noisier or less stable device.

Below are data sheets for 2 op amps that can be used for comparison.
[* links open in new browser window]

Device	Description	Overview and Specs PDF
LM741 - Operational Amplifier	General purpose low frequency	http://www.national.com/mpf/LM/LM741.html http://www.national.com/ds/LM/LM741.pdf
LMH6609 - 900MHz Voltage Feedback Op Amp	High speed high frequency	http://www.national.com/pf/LM/LMH6609.html http://www.national.com/ds/LM/LMH6609.pdf

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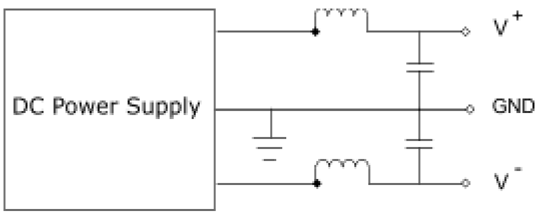
8.2 Power Supplies

The decision to use single or dual voltage power supplies may be arbitrary or a design constraint. The better choice is to use a dual balanced supply to minimize circuitry required to offset the DC baseline for AC inputs. If there is no choice and a design must be a single voltage supply, each of the circuits above needs to be modified.

8.2.1 Bypassing the power supply

Use a well regulated power supply and bypass with electrolytic capacitors, choke coils or a combination of the 2 to further filter the supply voltage to the circuits. Keep lead lengths or printed circuit board layouts as short as possible.



The best case for filtering out power supply connections is a series choke coil shunted by a capacitor (electrolytic or mylar) to minimize if not eliminate any noise and ringing on the supply rails.

Choke coils may be expensive and impose additional space requirements in a final product, but they provide excellent filtering for a prototype.

Figure 8.2-1 Filtering the Power Supply

Large electrolytic capacitors shunt out high frequency components and help keep the supply voltage regulated by absorbing voltage spikes or instantaneous load changes. Filtering power supplies helps keep the op amp operating in the range of its rated PSRR

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8.3 Offsetting and Stabilizing

The op amp data sheet may have specifications for offsetting and stabilizing. Below are cases for avoiding common-mode noise and shifting the DC baseline.

8.3.1 Common mode noise

When the intention to apply a constant 0 volts to either input terminal, tie the terminal directly to ground rather than through a resistor. This would prevent any voltage appearing due to leakage currents and having to offset or null it with additional biasing. Recommendations for offsetting mentioned in the op amp data sheet should also be followed.

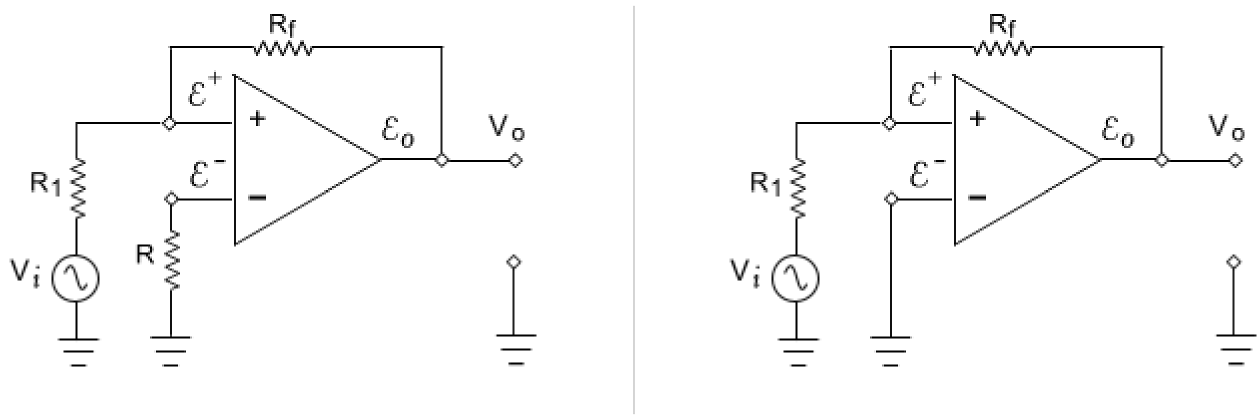

Figure 8.3-1 Avoiding common mode noise

8.3.2 Shifting the DC baseline

Working with AC sources that alternate between positive and negative voltages, and being constrained to using single supply op amps, requires shifting the DC baseline to avoid clipping of the output signal.

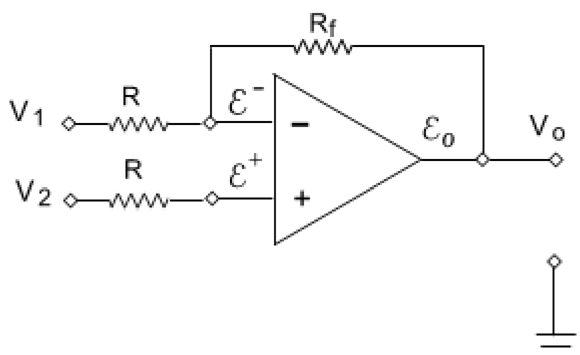

Figure 8.3-2 DC baseline shift example

Take Circuit 4 2 Inverting amplifier or attenuator as an example. If we are using a single supply and apply a sine wave as an input, the output produced would only be the positive portion of the input signal.

Modifying the circuit as in the figure to the left will provide the DC shift that is needed to provide the expected output; an amplified or attenuated sine wave alternating about a DC baseline

With no signal but the DC bias applied, $V_1 = V_2 = V_B$ and R is the thevenin equivalent of voltage dividers as shown in Figs 8.3-3.

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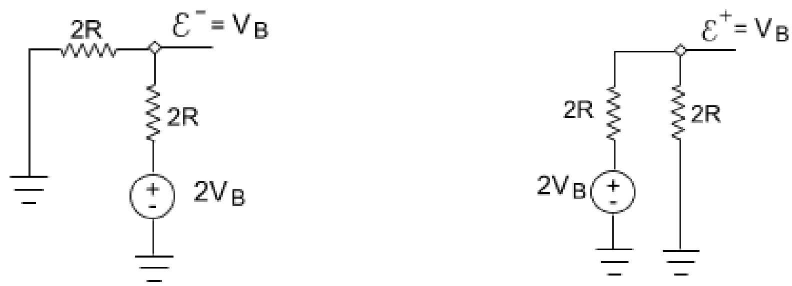

Figure 8.3-3 $\mathcal{E}^-$ loop with $V_S = 0$ and $\mathcal{E}^+$ loop with bias V_B

With no signal applied, the output $V_o = V_B$ which is now the baseline for this circuit.
Note that we are working in the frequency domain so V_B is $V_B(s) = \frac{V_{DC}}{s}$ the inverse Laplace being V_{DC} which is a constant voltage in the time domain.

If we remove the ground from the first resistor in the left - hand figure, and apply a signal V_S, V_1 becomes $V_1 = V_S + V_B$ and the output of the op amp becomes $V_o = V_B - V_S \frac{R_f}{2R}$

If V_B has a magnitude of V_{DC} and our input $v_s(t) = V_m \sin(\omega t)$, our output $v_o(t) = V_{DC} - V_m \sin(\omega t) \cdot \frac{R_f}{2R}$

The design constraint here is to keep the output voltage peaks in the range of the op amp specs and the power supply voltage.

positive peak being $V_{DC} + V_m$

negative peak being $V_{DC} - V_m$

We just applied a DC baseline shift to the inverting amplifier / attenuator circuit, the same can be done for the inverting integrator or differentiator circuits

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8.4 Impedance Matching and Phase Compensation

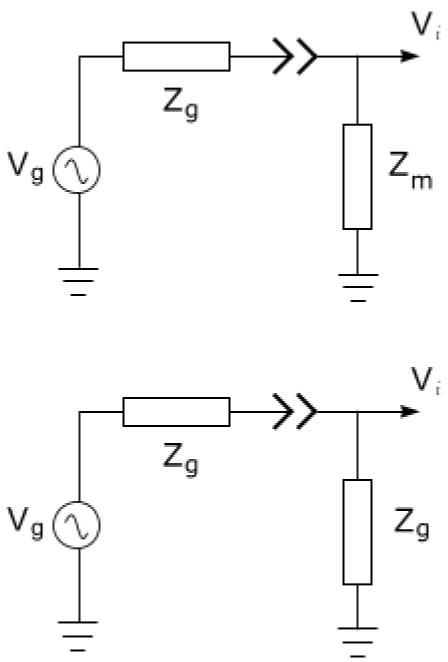

Figure 8.4-1 Impedance matching

An input voltage source may have an internal impedance that needs to be matched to achieve:
maximum power transfer
cancel out any reactances
pre-condition the phase

The figure to the top left represents either a directly connected voltage source with an internal impedance or a downstream thevenin equivalent source.

Shunting the input source with an impedance Z_m using RLC components will satisfy the requirements.

V_i then becomes $\frac{V_g \cdot Z_m}{Z_g + Z_m}$

Matching an impedance where $Z_g = Z_m$

when $Z_g = R_g$ or $\pm jX_g$ or $R_g + jX_g$

$V_i = \frac{V_g}{2}$

without introducing any phase components

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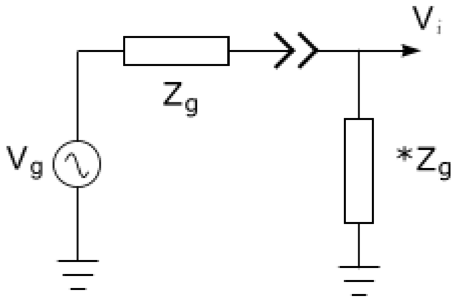

Figure 8.4-2 Canceling reactance

To cancel the reactance portion,

use an impedance of $Z_m = Z_g^*$

then with $Z_g = R_g \pm jX_g \quad Z_g^* = R_g \mp jX_g$

and the input will be $V_i = \dfrac{V_g\left(R_g \mp jX_g\right)}{2R_g}$

Z_{in} seen by V_g becomes

$Z_{in} = 2R_g$ thereby cancelling the reactance

and introducing a phase angle of $\mp\theta = \tan^{-1}\dfrac{X_g}{R_g}$

$*$ if Z_g is purely a reactance, terminating with Z_g^* will present a short circuit to the source

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Appendix A. Commonly Used Terms

Common-Mode Voltage Range
Typically the range of voltages on the input terminals for which the amplifier’s performance is specified

Common-Mode Rejection Ratio
The ratio of differential voltage amplification to common-mode voltage amplification. It is measured by determining the ratio of a change in input common-mode voltage to the resulting change in input offset voltage change.

Gain Bandwidth Product
The product of a given input frequency and the op-amp open loop gain at that frequency (usually specified in MHz, voltage feedback amplifiers only.)

Input bias current
The Input Current specification is the average of the currents drawn by the two input pins. Input current is also often called "bias current"

Input Offset Current
The difference of the currents entering the two input terminals of a balanced amplifier

Input Offset Voltage
The DC error voltage which exists between the input terminals due to non-ideal balancing of the input stage to the output. It is multiplied by the closed loop gain

Offset Current Temperature Coefficient
The average rate of change in offset current for junction temperature variation over a specified temperature range

Offset Voltage Temperature Coefficient
The average rate of change in offset voltage for the junction temperature variation over a specified temperature range

Output Offset Voltage
The output voltage when the 2 input terminals are grounded.

Output Voltage Swing
The maximum peak-to-peak output voltage swing under specified load and supply voltages

Power Supply Rejection Ratio
Power Supply Rejection Ratio (PSRR) can be one of two specifications. DC PSRR is the ratio of the change in a specified parameter (e.g., Full Scale Error) that results from a specified change in the power supply voltage. AC PSRR is measured with a signal of specified frequency and amplitude riding upon the power supply and is the ratio of the output amplitude of that signal at the output to its amplitude on the power supply pin. PSRR is usually specified in dB

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Slew Rate
The rate that an amplifier output changes from one voltage level to another, usually given in V/μsec, when a step or square wave input is applied. Typically it is the average rate measured from 10% to 90% of the total output voltage change

Unity Gain Bandwidth
The frequency where the amplifier open loop gain equals to one. It equals GBW if the op amp has a single pole roll-off in its frequency response

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Appendix B. Bibliography

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Online Reference List [* links open in new browser window]

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