1 Introduction

This chapter is designed to take up where Chapter 1, entitled “Laboratory Induction,” left off. In that document, the concepts of voltage, current, energy and power have been explained, along with Ohm’s law and its various implications. Through instruction and laboratory exercises, you are also being introduced to diodes and bipolar junction transistors.

In Section 2, we first introduce you to some common amplifier configurations, involving one or two transistors. You may find these useful in your project, particularly when driving an output device with significant amounts of current — 100’s of milli-Amps through to multiple Amps.

In Section 3, we introduce operational amplifiers and some simple circuits which use them. To build amplifiers which are highly linear, you will find opamps provide a much simpler solution than directly constructing a circuit based on individual transistors. This is because opamps are packaged integrated circuits, which already wrap up sophisticated networks of transistors.

2 Simple Transistor Amplifiers

2.1 Emitter Follower (Buffer)

We begin by considering a simple and very useful configuration known as the “emitter follower,” depicted in Figure 1. Recall that so long as current flows into the base of an NPN transistor, its base-emitter junction exhibits a roughly constant voltage drop of $V_{BE} \approx 0.6V$. That is, the base voltage is higher (more positive) than the emitter voltage by about 0.6V.

What actually happens is that the current through the base-emitter diode junction is a very strong function of $V_{BE}$. The relationship (you don’t really need to know this yet) is given by

$$I_B = I_0 \left( e^{\frac{V_{BE}}{kT}} - 1 \right)$$
Figure 1: Emitter follower configuration.

where $I_0$ is a characteristic current, determined by the device’s construction, $k = 8.617385 \times 10^{-5}$ eV K$^{-1}$ is Boltzmann’s constant and $T$ is the absolute temperature, measured in degrees Kelvin. At voltages below 0.5V, essentially no current flows. Thereafter, each increment of 0.1V in the value of $V_{BE}$ causes the current $I_B$ to be multiplied by a factor of about 50 at room temperature (300° K).

Since the transistor is a current gain device, large changes in $I_B$ produce even larger changes in the collector current, $I_C$. These act to maintain the output voltage from the emitter follower close to 0.6V below $V_{in}$, largely irrespective of how much current must be delivered to the output load to make this happen.

If $V_{out}$ is too small (say 0.1V too small), $I_B$ will be increased by a large amount (say by a factor of 50) and hence $I_C$ will be increased by an even larger amount (say $\beta \times 50 = 300 \times 50$) and this serves to increase $V_{out}$. Similarly, if $V_{out}$ is too large, $I_B$ and $I_C$ will decrease, causing the output voltage to decrease, since the output load has less current flowing in it. These adjustments take place almost instantaneously, so that for most practical purposes $V_{out}$ appears to track perfectly with $V_{in}$.

The value of the emitter follower lies in the fact that almost all of the load current is supplied from the collector, so that the base current can be quite small. This allows the input to drive the voltage across a large load, without having to supply much current (and hence much power). Specifically, we have

$$\beta I_B = I_C = I_E - I_B$$

$$\Rightarrow \quad I_B = \frac{I_E}{\beta + 1} \approx I_E/\beta$$

We say that the emitter follower acts as a “buffer” between the input and the output. It serves to amplify current and power, but not voltage.

If the load resistance becomes very small, large currents will flow through the transistor, in an attempt to maintain $V_{out}$. In the extreme case where the load is shorted to ground, these currents would ideally be unbounded. This reveals a
practical weakness of the circuit in Figure 1, in that the transistor could easily be destroyed by an adverse load. To avoid this problem, it is common to include a small resistor $R_C$ between $V_{CC}$ and the collector. This limits the current through the collector to strictly less than $V_{CC}/R_C$. If the input is capable of supplying large currents, then a small resistor may also need to be inserted between $V_{in}$ and the base for the same reason.

Figure 2 depicts an interesting application of the emitter follower configuration, in the construction of a simple voltage regulator device. The load limiting collector resistor $R_C$ is shown explicitly here. Also, the value of $V_{in}$ is fixed to 5.6V, with the aid of a device known as a Zener diode. Notice that the zener diode is reverse biased. In the forward direction, it conducts current just like an ordinary diode, with a voltage drop on the order of 0.6V like other silicon diodes. Unlike regular diodes, however, zeners are designed to allow current flow in the reverse direction. In regular diodes, this process is known as breakdown, and is undesirable. Zener diodes are designed to provide a stable, well-defined reverse breakdown voltage. In the figure, this reverse breakdown voltage is 5.6V – a common Zener diode voltage. So long as current flows in the diode, the voltage across the diode will be roughly constant. This is true roughly irrespective of the current flowing through the Zener, so long as you don’t exceed the Zener diode’s maximum power rating. Larger Zener diodes can dissipate 1W of power, while smaller devices might not be able to dissipate more than $\frac{1}{8}$W. As the load current $I_E$ varies, $I_B = I_E/(1 + \beta)$ also varies. The resistor $R_B$ needs to be selected so that the maximum expected load can be met without taking all of the current flowing in $R_B$ away from the Zener diode. Writing $V_{\text{min}}$ for the minimum voltage provided by the unregulated power supply, we have

$$I_{R_B} \geq (V_{\text{min}} - 5.6) / R_B$$

so that we can safely support loads of

$$I_{E,\text{max}} = (1 + \beta) I_{B,\text{max}} \leq (1 + \beta) (V_{\text{min}} - 5.6) / R_B$$

Figure 2 shows that the voltage regulator may be understood as a 3-terminal device, where the terminals are indicated by shaded circles. One terminal is attached to ground, the reference voltage against which the regulated voltage is to be measured. A second terminal is connected to the positive supply rail, and the third terminal supplies the regulated output voltage (5V in this case). Three terminal voltage regulators can be purchased as individual encapsulated circuits. The “7805” voltage regulator, for example, is a traditional choice for 5V power supplies. These monolithic voltage regulators are a good deal more sophisticated than the one illustrated in Figure 2, but the principle is essentially the same. You might need a well-defined regulated 5V power supply if your project involves the powering of digital logic IC’s. The input power supply could be provided by a battery if you like, so long as the positive supply terminal’s voltage is sufficiently high relative to the regulated output voltage. You can determine the necessary supply voltage by consulting the regulator’s data sheets.
2.2 Common Emitter (Current Source)

In this section, we consider a very different type of amplifier: one whose role is best understood as converting voltage into current. Specifically, the circuit in Figure 3 is designed to produce a collector current $I_C$ which is proportional to the input voltage, $V_{in}$. If this current is drained through a fixed load resistor $R_{load}$, then the voltage across $R_{load}$ will of course be proportional to $I_C$ and hence $V_{in}$. Changing the value of $R_{load}$ will change the output voltage, but not the current $I_C$. Thus, this type of amplifier is best understood as implementing a current source, whose output current is proportional to $V_{in}$.

We could begin by omitting the emitter resistor $R_E$ shown in Figure 3. Without $R_E$ we can say that the current in the base is given by

$$I_B = \frac{V_{in} - V_{BE}}{R_B}$$

so that

$$I_C = \beta \frac{V_{in} - V_{BE}}{R_B} = V_{in} \frac{\beta V_{BE}}{R_B}$$

is indeed essentially proportional to the input voltage (apart from an offset). The chief problem with this approach is that the collector current depends heavily upon $\beta$, the transistor’s HFE. Unfortunately, $\beta$ is not a parameter that can be accurately controlled during manufacturing. Two transistors which are manufactured by exactly the same process may have $\beta$ values which vary by a factor of 2 or more. Also, $\beta$ tends to vary with temperature and also with the magnitude of the collector current itself. That is, the current gain property of the transistor is actually as linear as we have been implying. All we can really count on is that $\beta$ should be a decently large number (typically $\geq 100$ for small signal transistors).
Figure 3: Common emitter configuration, with current gain determined primarily by the emitter resistor, \( R_E \).

Now suppose we re-introduce the emitter resistor, but do away altogether with \( R_B \). In this case, the voltage across \( R_E \) must be \( V_{in} - V_{BE} \), from which we conclude that

\[
I_E = \frac{V_{in} - V_{BE}}{R_E}
\]

This means that the input voltage controls \( I_E \) directly. Moreover, \( I_E \) is dominated by the collector component \( I_C \). Specifically, we have \( I_E = I_C + I_B = I_C (1 + 1/\beta) \), so that

\[
I_C = \frac{V_{in}}{1 + 1/\beta} \left( \frac{1}{R_E} - \frac{V_{BE}/R_E}{1 + 1/\beta} \right)_{\text{constant}}
\]

Evidently, the collector current is again proportional to the \( V_{in} \), apart from a constant offset. In this case, however, the coefficient of proportionality is largely independent of the actual value of \( \beta \), so long as it is large.

If the circuit of Figure 3 is to be used as a current source, with unknown load resistance \( R_{load} \), it is possible that \( R_{load} \) will be so large that the \( I_C \) given by equation (1) cannot flow without producing a voltage drop in excess of the supply voltage. When this happens, the transistor ceases to amplify current and most of \( I_E = V_{in} - V_{BE}/R_E \) must be supplied via the base. This can lead to destructive base currents flowing in the transistor. The main reason for including an input resistor \( R_{in} \) in Figure 3, is to limit the magnitude of these base currents. The resistor should be small enough that \( I_B R_B \) is small compared with \( I_E R_E \). Otherwise the properties of the amplifier will again come to depend strongly on \( \beta \). With \( R_{in} \) included, we can derive the complete expression for \( I_C \) by noting that \( V_{in} \) is the sum of three voltages: the relatively constant \( V_{BE} \); the voltage drop \( I_E R_E \) across the emitter resistor; and the voltage drop \( I_B R_B \) across the input resistor. Thus, we have

\[
V_{in} = V_{BE} + I_C \left( 1 + \frac{1}{\beta} \right) R_E + I_C \frac{1}{\beta} R_B
\]
from which we conclude that

\[ I_C = \frac{V_{in} - V_{BE}}{R_E (1 + 1/\beta) + R_B/\beta} \]

Thus, to preserve our resilience to variations in \( \beta \), we need to ensure that \( R_B/\beta \ll R_E \), while also keeping \( R_B \) large enough to protect the transistor from excessive base currents.

Before concluding this section, we point out that the offset due to \( V_{BE} \) in the above equations can be eliminated by using a circuit of the form shown in Figure 4. In this case, a separate diode is used to establish a voltage drop which is comparable to \( V_{BE} \). So long as some current flows through the diode in the forward direction, this voltage drop will be maintained. Some elementary manipulation then yields the linear V-I characteristic

\[ I_C = \frac{V_{in}}{R_E} \]

This relationship holds so long as

\[ \frac{V_{CC} - V_{in} - 0.6}{R_B} > I_B = \frac{V_{in}}{\beta R_E} \]

over the range of input voltages \( V_{in} \), which are of interest. The diode used in Figure 4 is typically a small signal diode such as the cheap-as-chips “1n4004.”

### 2.3 Darlington Pair

Without straying far from our goal of a quick introduction to simple transistor amplifiers, it is worth introducing the “Darlington pair” configuration, named after Lord Darlington. As shown in Figure 5, a new 3-terminal device is constructed by cascading two NPN transistors, or two PNP transistors. This new
3-terminal device behaves like a single transistor whose current gain is essentially the product of the two individual transistor gains. To see this, let $I_{B1}$ and $I_{C1}$ be the base and collector currents in the first transistor, while $I_{B2} = I_{E1}$ and $I_{C2}$ refer to the second transistor. The properties of the Darlington pair can then be expressed as

$$I_C = I_{C1} + I_{C2} = I_{C1} + \beta_2 I_{E1}$$
$$= I_{C1} + \beta_2 I_{B1} + \beta_2 I_{C1}$$
$$= I_{B1} \times (\beta_1 + \beta_2 + \beta_1 \beta_2)$$

You can use the darlington pair where you need a transistor with very high current gain. In fact, darlington pairs are sometimes packaged together. One important example is the “L-51P3C” photo-transistor with which you have been supplied for detecting infrared signals; this device is actually a photo-darlington transistor. In the case of photo-transistors, the base is not usually exposed directly through leads on the device; the base current is generated internally through the conversion of photons into electrons.

### 2.4 Push-Pull Driver

For our final look at discrete transistor amplifiers, we will consider something which seems a good deal more complex. Do not be put off, however, by the apparent complexity of the circuit in Figure 6; not all of the elements are strictly necessary. The purpose of this circuit is to drive a heavy load (i.e., large currents), where current may flow either into or out of the load, depending on the output voltage. To this end, we use a double-ended power supply: the positive rail $V_{CC}$ has a potential above the ground (e.g., $+12\text{V}$); the negative rail $V_{EE}$ has a potential below the ground (e.g., $-12\text{V}$). The load has one terminal connected to ground. The amplifier pulls the other terminal either toward the positive rail

![Figure 5: NPN and PNP darlington pair configurations.](image-url)
or toward the negative rail, causing current to flow either toward or away from ground, respectively.

The circuit of Figure 6 consists essentially of two emitter followers. Ignore, for the moment, the input diodes $D_1$ and $D_2$ (imagine they are just wires) and suppose all resistors are replaced by pure conductors. In this case, if $V_{in} > V_{BE}$ the NPN transistor is “on” and its emitter voltage will follow $V_{in}$, yielding

$$V_{out} = V_{in} - V_{BE} > 0.$$  

In this case, the NPN transistor is the source of the load current, which flows toward ground. If, on the other hand, $V_{in} < -V_{BE}$, the PNP transistor will be in the “on” state and its emitter voltage will follow $V_{in}$, yielding

$$V_{out} = V_{in} + V_{BE} < 0.$$  

In this case, the PNP transistor is the sink for the load current, which flows from ground. Of course, the difficulty with this is that there is a dead zone when $-V_{BE} < V_{in} < V_{BE}$, during which both transistors are off and no current flows in the load.

The solution to this problem is to remove the voltage offset, using essentially the same method described in connection with Figure 4. Resistors $R_{B1}$ and $R_{B2}$ are chosen as large as possible while still small enough to guarantee that a forward bias current flows through diodes $D_1$ and $D_2$, so that voltage drops comparable to $V_{BE}$ appear across each of them. Ignoring $R_{E1}$, $R_{E2}$ and $R_{adj}$ for the moment, this means that the NPN transistor should be on whenever $V_{in} > 0$, while the PNP transistor should be on whenever $V_{in} < 0$. In both cases, $V_{out}$ will follow $V_{in}$ without any voltage offset. While this seems to have solved the offset problem, it raises the possibility that both transistors might be on at the same time when $V_{in}$ is very close to 0, if there is any mismatch between the base-emitter and diode voltage drops — indeed there will always be some mismatch. This can cause very large currents to flow directly from $V_{CC}$ to $V_{EE}$ through the two transistors. Emitter resistors $R_{E1}$ and $R_{E2}$ provide the
solution to this problem. These are typically very small (e.g., $1\, \Omega$). If significant current flows through both transistors, an additional voltage drop will appear across the emitter resistors, increasing the potential difference between the two transistors' emitters. The potential difference between the two bases is not affected, however, so at least one of the transistor’s base-emitter voltage must decrease to the point where it is essentially off.

The $R_{adj}$ resistor is typically very much smaller than $R_{B1}$ and $R_{B2}$. This is a variable resistor whose value may be tuned to ensure that the potential difference between the bases of the transistors is every so slightly larger than $2V_{BE}$. This guarantees that there will be no dead zone, so that $V_{out} = V_{in}$ over the entire range of usable input voltages.

3 Operational Amplifiers

In this section, we introduce to a most useful electronic component known as the operational amplifier (Opamp). An opamp is actually a monolithic integrated circuit (IC), which embodies a sophisticated high gain transistor amplifier, based on principles similar to those introduced in the previous section. The transistors in the opamp IC are very tiny, so they cannot usually drive large currents. However, many transistors can be arrayed on a small piece of silicon, allowing for sophisticated amplifier designs.

Most opamps act as very high gain voltage to voltage amplifiers, with gains $K$ of $10^6$ or more. The exact value of $K$ is usually not well defined and may vary significantly from batch to batch within the same manufacturing process. However, we will use this high gain operational amplifier to construct amplifying circuits whose gain is well defined, regardless of the actual value of $K$, so long as $K$ is very large. This is similar in spirit to the way in which we constructed common emitter current sources in Section 2.2, whose characteristics are largely insensitive to the current gain $\beta$ of the transistor itself.

One particularly valuable feature of the opamp is that it acts as a differential voltage amplifier. There are two input voltages $V_+$ and $V_-$, rather than one, and the output voltage $V_O$ is proportional only to the difference between $V_+$ and $V_-$. That is,

$$V_O = K \cdot (V_+ - V_-)$$

(2)

For an introductory treatment, it is helpful to restrict our attention to double-ended power supplies, with positive rail $V_{CC}$ and negative rail $V_{EE}$, both different from the reference ground potential, as shown in Figure 7. Ideally, $V_O$ can take any value in the range $V_{EE} < V_O < V_{CC}$, although practical opamps typically cannot produce output voltages smaller than $V_{EE}$ plus 1 or 2 volts or greater than $V_{CC}$ minus 1 or 2 volts. Of course, we cannot expect equation (2) to hold unless $V_{EE} < V_+, V_- < V_{CC}$. Interestingly, though, some opamps can function correctly even when the input voltages lie slightly above $V_{CC}$ or slightly below $V_{EE}$, depending on the internal design. All of this information may be gleaned from data sheets. Indeed, it is very important to consult the data sheet for an opamp before designing circuits around it. You might start with a popular
cheap-as-chips opamp such as the “741.” The School’s electronics laboratory stocks a number of inexpensive opamps and can provide you with the relevant data sheets.

It is important to realize that equation (2) holds regardless of whether $V_+ > V_-$ or $V_+ < V_-$. In the former case $V_O$ will be positive, while in the latter case $V_O$ will be negative. It is also worth knowing that, in addition to its voltage amplifying properties, the opamp also provides massive current amplification. By this, we mean that the current flowing into or out of the two inputs is usually extremely small—anywhere from $\mu A$ down to $p A$ (pico-Amps), depending on the internal design. Meanwhile the current sourced or sunk by the opamp output can be significant—typically on the order of milli-Amps (or even 100’s of milli-Amps).

In view of these remarkable properties, circuits involving opamps are generally designed and analyzed in the first instance with the assumption that the gain $K$ is infinite and the input “leakage” current is 0. Only in special circumstances, need the actual gain or leakage current limitations of the practical opamp be taken specifically into account. We will find this perspective helpful in understanding the simple amplifying circuits presented in the ensuing sub-sections.

### 3.1 Inverting Amplifier

To understand the behaviour of the circuit in Figure 8, we assume that no current flows into or out of the opamp inputs. Then all of the current $I$ in resistor $R_O$ must flow in $R_N$. Thus,

$$\frac{V_- - V_{in}}{R_N} = \frac{V_O - V_-}{R_O}$$

Figure 7: Operational amplifier, shown here with positive and negative supply rails.
Noting also that the $V_+ = 0$ (grounded), we find that

$$V_O = K (V_+ - V_-) = -KV_-$$

Substituting $V_- = -V_O/K$ into equation (3), we obtain

$$\frac{V_{in}}{R_N} = \frac{V_O - V_-}{R_O} - \frac{V_-}{R_N} = \frac{V_O}{R_O} \cdot \left(1 + \frac{1}{K} + \frac{R_O}{KR_N}\right)$$

(4)

For very large $K$, this simplifies to

$$\frac{V_{in}}{R_N} \approx \frac{V_O}{R_O}$$

(5)

or

$$V_O \approx -V_{in} \frac{R_O}{R_N}$$

(6)

What is actually going on is this. If $V_- \text{ becomes even slightly negative, the opamp’s huge gain serves to produce a large output voltage } V_O, \text{ which pulls } V_- \text{ back up again via } R_O. \text{ Similarly, if } V_- \text{ becomes even slightly positive, } V_O \text{ falls below ground and pulls } V_- \text{ back down again. This is called “negative feedback.” Systems which have negative feedback reach a stable equilibrium. In this case, the negative feedback causes } V_- \text{ to remain extremely close to the ground potential all the time.}

Armed with the observation that the presence of negative feedback keeps $V_- \approx V_+ = 0$, we analyze opamp circuits to first order precision by simply assuming that $V_- = V_+$. To summarize, our ideal model of the opamp has just two properties:

**Property 1:** Input currents $I_+$ and $I_-$ may be safely taken to be 0; and

**Property 2:** Input voltages $V_-$ and $V_+$ may be safely taken to be identical.
These two properties are sufficient to derive equation (5) much more simply. In particular, the current flowing from $V_-$ to $V_{in}$ must be $-V_{in}/R_N$ and this must be identical to the current flowing from $V_O$ to $V_-$, which must be $V_O/R_O$.

It is worth noting that the above properties hold only so long as the circuit provides negative feedback. We could construct another circuit, identical to the first, except that the roles of $V_+$ and $V_-$ are reversed. Applying the conclusions of properties 1 and 2 to this circuit would produce exactly the same conclusion, that $-V_{in}/R_N = V_O/R_O$. However, this would not work at all in practice, because such a circuit would have positive feedback: small positive input voltages at $V_+$ would produce large output voltages $V_O$, which would serve to pull $V_+$ to even higher voltages, until $V_+$ reaches the limit imposed by the power supply rails. The properties, then, do not hold for circuits with positive feedback.

According to equation (5), the absolute values of resistors $R_N$ and $R_O$ are not important; only their ratio matters. This suggests that we could use very large resistors, so that the input to the amplifier at $V_{in}$ need only supply very small currents, $-I$. Another way to understand this is that the (upstream) circuit which drives the amplifier at $V_{in}$ effectively sees just a simple resistive load $R_N$, connected between $V_{in}$ and ground. This is because the opamp works to keep $V_-$ at the ground potential. By making $R_N$ very large, the amplifier will present very little load on the circuit which is supplying its input; this might be a microphone or sensor which produces an appreciable voltage but cannot supply much current (i.e., not much input power). There is, however, a limit to how large we can make $R_N$ before the microscopic currents flowing into or out of $V_-$ have a significant impact on the amplifier’s behaviour. This, of course, depends on the particular type of opamp you select—remember the data sheets.

### 3.2 Non-Inverting Amplifier

The circuit in Figure 9 is an excellent choice where positive, rather than a negative voltage gain is required. You can easily verify that this circuit provides negative feedback; in fact, the circuit is essentially identical to that in Figure 8, except that $V_+$ is now held at the input voltage level instead of ground. Since we have negative feedback, our two properties (0 leakage current and 0 differential input voltage) may be applied, from which we conclude that $V_- = V_{in}$ and hence

$$\frac{V_{in}}{R_N} = I = \frac{V_O - V_{in}}{R_O}$$

Equivalently,

$$\frac{V_O}{R_O} = \frac{V_{in}}{R_N} \cdot \left( 1 + \frac{R_N}{R_O} \right)$$

The voltage gain of the amplifier is thus

$$G = \frac{V_O}{V_{in}} = \frac{R_O}{R_{in}} \left( 1 + \frac{R_N}{R_O} \right) = 1 + \frac{R_O}{R_N}$$  \hspace{1cm} (7)$$

The input resistor $R_P$, shown in Figure 9 is not strictly necessary. Its role is to protect the opamp’s delicate input circuitry against excessive input currents.
which might flow if $V_{in}$ were to exceed the opamp’s operating range (see data sheets). The circuit which drives the amplifier at $V_{in}$ sees an effective resistance (also called “input impedance”) which is much greater than $R_P$, assuming the input leakage current is sufficiently small.

Apart from its positive gain, one potential benefit of the circuit in Figure 9 over that in Figure 8 is that the amplifier’s input impedance does not depend on the values of $R_N$ and $R_O$, which are used to set the gain. This allows the use of smaller resistances for $R_N$ and $R_O$. Lower valued resistors tend to be less affected by parasitic capacitive and inductive components (you will understand these in later years of the course).

### 3.3 Simple Differential Amplifier

It is a very small step indeed now to combine the inverting and non-inverting properties of the amplifiers described in Sections 3.1 and 3.2 into the single circuit of Figure 10. Again, we have negative feedback so our two magic properties apply ($0$ leakage current and $V_+ = V_-$. It follows that $V_- = V_+ = V_{in1}$ and hence we obtain

$$\frac{V_{in1} - V_{in2}}{R_N} = I = \frac{V_O - V_{in1}}{R_O}$$

Simplifying these relations yields

$$V_O = V_{in1} \left( 1 + \frac{R_O}{R_N} \right) - V_{in2} \frac{R_O}{R_N}$$

That is, $V_O$ is just the sum of the inverting and non-inverting contributions described by equations (6) and (7). For large values of $R_O/R_N$ (i.e., large gains), the two channels experience gains of almost identical magnitude and we can write

$$V_O \approx \frac{R_O}{R_N} (V_{in1} - V_{in2})$$
A classic way to compensate for the small difference in gains associated with the non-inverting input $V_{in1}$, whose gain is $1 + \frac{R_O}{R_N}$, and the inverting input $V_{in2}$, whose gain is $\frac{R_O}{R_N}$, is to add an additional resistor $R_G$ between $V_+$ and ground, as shown in Figure 11. In this case, the signal at $V_+$ is the output of a voltage divider, taking the value

$$V_+ = V_{in1} \frac{R_G}{R_G + R_P}$$

For the two channels to have gains of equal magnitudes, both equal to $\frac{R_O}{R_N}$, we have only to ensure that

$$\frac{R_G}{R_G + R_P} = \frac{R_O}{R_N} \Rightarrow 1 + \frac{R_N}{R_O} = 1 + \frac{R_P}{R_G}$$

In most designs, we simply choose $R_N = R_P$ and $R_O = R_G$. 

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Figure 10: Combination of the inverting and non-inverting amplifiers from Figures 8 and 9.

Figure 11: Opamp-based differential amplifier.
3.4 Practical Considerations

Real opamps exhibit a wide range of non-idealities which can become important in certain circumstances. We have already mentioned the fact that the input currents at \( V_+ \) and \( V_- \) are not exactly 0. In fact, they cannot be, since measurements must always have some effect (even if small) on the signals which are being measured. These currents may become important where amplifiers with high input impedances are required. Of course, the gain of the opamp also cannot be truly infinite, and this may impact the overall gain of the amplifier, as suggested by equation (4). Some other non-idealities which may need to be taken into account are listed below. As always, data sheets are the key to understanding the impact of these non-idealities on your circuit and to selecting an appropriate opamp.

**Input offset voltage:** Ideally, \( V_0 = 0 \) when \( V_+ = V_- \). In practice, however, this condition is reached when \( V_+ - V_- = V_{\text{off}} \), where \( V_{\text{off}} \) is typically on the order of milli-volts. The value of \( V_{\text{off}} \) can vary from device to device and with temperature and other conditions. The presence of non-zero \( V_{\text{off}} \) is only important if you want precise amplification of very small signals. This problem arises in measurement applications and also when amplifying the signals produced by certain types of sensors. For these applications, you may require a precision opamp. Precision opamps are designed to have offset voltages in the 10’s of micro-volts; they also usually come with additional pins to which you can attach an offset trimming circuit.

**Common mode rejection:** Common mode rejection refers to the ability of the opamp to ignore the absolute values of \( V_+ \) and \( V_- \), amplifying only the difference \( V_+ - V_- \). The common mode gain of the opamp is the amount by which changes in the absolute value of \( V_+ = V_- \) are amplified in the output \( V_O \). That is, setting \( V_+ = V_- = V_C \), the common mode gain is \( K_C = \Delta V_O / \Delta V_C \). A figure of merit for opamps is the common mode rejection ratio, defined as \( K / K_C \). This value is typically very large, so it is quoted on a logarithmic scale known as decibels (dB). The value of \( K / K_C \), expressed in dB is \( 10 \log_{10} K / K_C \). Many opamps have common mode rejection ratios (CMRR) in excess of 100 dB.