Positive feedback

In Chapter 7 we showed how negative feedback can be applied to an amplifier to form the basis of a stage which has a precisely controlled gain. An alternative form of feedback, where the output is fed back in such a way as to reinforce the input (rather than to subtract from it), is known as positive feedback.

Fig. 9.1 shows the block diagram of an amplifier stage with positive feedback applied. Note that the amplifier provides a phase shift of 180° and the feedback network provides a further 180°. Thus the overall phase shift is 0°. The overall voltage gain, G, is given by:

$$\text{Overall gain, } G = \frac{V_{out}}{V_{in}}$$

By applying Kirchhoff's Voltage Law

$$V_{in}' = V_{in} + \beta V_{out}$$

thus

$$V_{in} = V_{in}' - \beta V_{out}$$

and

$$V_{out} = A_v \times V_{in}$$

where Av is the internal gain of the amplifier. Hence:

$$\text{Overall gain, } G = \frac{A_v \times V_{in}'}{V_{in}' - \beta V_{out}} = \frac{A_v \times V_{in}'}{V_{in}' - \beta(A_v \times V_{in}')}$$

$$\text{Thus, } G = \frac{A_v}{1 - \beta A_v}$$

Now consider what will happen when the loop gain, βA_v, approaches unity (i.e. when the loop gain is just less than 1). The denominator $(1 - \beta A_v)$ will become close to zero. This will have the effect of *increasing* the overall gain, i.e. the overall gain with positive feedback applied will be *greater* than the gain without feedback.

It is worth illustrating this difficult concept using some practical figures. Assume that you have an amplifier with a gain of 9 and one-tenth of the output is fed back to the input (i.e. $\beta = 0.1$). In this case the loop gain $(\beta \times A_v)$ is 0.9.

With negative feedback applied (see Chapter 7) the overall voltage gain will be:

$$G = \frac{A_v}{1 + \beta A_v} = \frac{9}{1 + (0.1 \times 9)} = \frac{9}{1 + 0.9} = \frac{9}{1.9} = 4.7$$

With positive feedback applied the overall voltage gain will be:

$$G = \frac{A_v}{1 - \beta A_v} = \frac{10}{1 - (0.1 \times 9)} = \frac{10}{1 - 0.9} = \frac{10}{0.1} = 90$$

Now assume that you have an amplifier with a gain of 10 and, once again, one-tenth of the output is fed back to the input (i.e. $\beta = 0.1$). In this example the loop gain $(\beta \times A_v)$ is exactly 1.

With negative feedback applied (see Chapter 7) the overall voltage gain will be:

$$G = \frac{A_v}{1 + \beta A_v} = \frac{10}{1 + (0.1 \times 10)} = \frac{10}{1 + 1} = \frac{10}{2} = 5$$

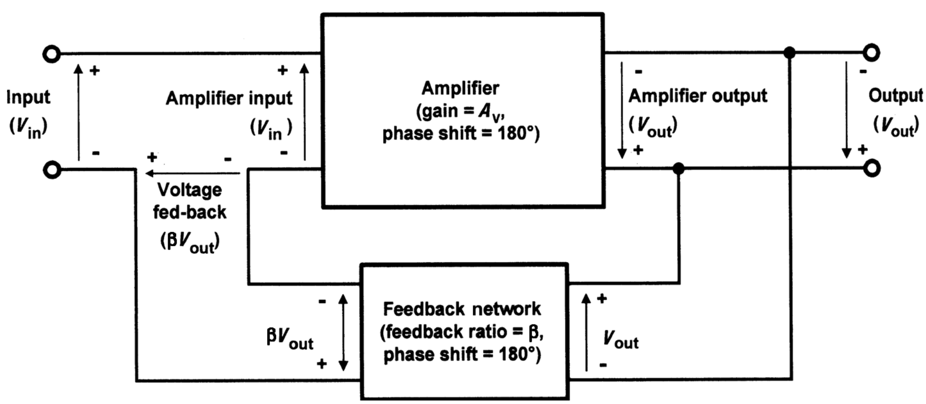

Figure 9.1 Amplifier with positive feedback applied

With positive feedback applied the overall voltage gain will be:

$$G = \frac{A_v}{1-\beta A_v} = \frac{10}{1-(0.1\times 10)} = \frac{10}{1-1} = \frac{10}{0} = \infty$$

This simple example shows that a loop gain of unity (or larger) will result in infinite gain and an amplifier which is unstable. In fact, the amplifier will oscillate since any disturbance will be amplified and result in an output.

Clearly, as far as an amplifier is concerned, positive feedback may have an undesirable effect – instead of reducing the overall gain the effect is that of reinforcing any signal present and the output can build up into continuous oscillation if the loop gain is 1 or greater. To put this another way, oscillator circuits can simply be thought of as amplifiers that generate an output signal without the need for an input!

Conditions for oscillation

From the foregoing we can deduce that the conditions for oscillation are:

(a) the feedback must be positive (i.e. the signal fed back must arrive back in-phase with the signal at the input);

(b) the overall loop voltage gain must be greater than 1 (i.e. the amplifier's gain must be sufficient to overcome the losses associated with any frequency selective feedback network).

Hence, to create an oscillator we simply need an amplifier with sufficient gain to overcome the losses of the network that provide positive feedback. Assuming that the amplifier provides 180° phase shift, the frequency of oscillation will be that at which there is 180° phase shift in the feedback network.

A number of circuits can be used to provide 180° phase shift, one of the simplest being a three-stage *C*–*R* ladder network that we shall meet next. Alternatively, if the amplifier produces 0° phase shift, the circuit will oscillate at the frequency at which the feedback network produces 0° phase shift. In both cases, the essential point is that the feedback should be positive so that the output signal arrives back at the input in such a sense as to reinforce the original signal.

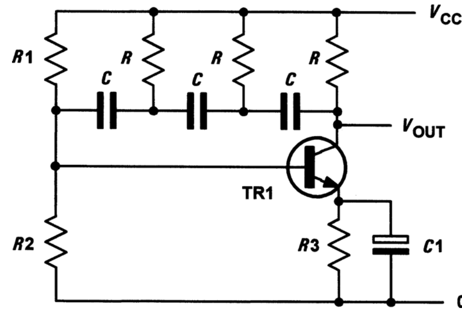

Figure 9.2 Sine wave oscillator based on a three-stage *C*–*R* ladder network

Ladder network oscillator

A simple phase-shift oscillator based on a three-stage *C*–*R* ladder network is shown in Fig. 9.2. TR1 operates as a conventional common-emitter amplifier stage with *R*1 and *R*2 providing base bias potential and *R*3 and *C*1 providing emitter stabilization.

The total phase shift provided by the *C*–*R* ladder network (connected between collector and base) is 180° at the frequency of oscillation. The transistor provides the other 180° phase shift in order to realize an overall phase shift of 360° or 0° (note that these are the same).

The frequency of oscillation of the circuit shown in Fig. 9.2 is given by:

$$f = \frac{1}{2\pi \times \sqrt{6}CR}$$

The loss associated with the ladder network is 29, thus the amplifier must provide a gain of *at least* 29 in order for the circuit to oscillate. In practice this is easily achieved with a single transistor.

Example 9.1

Determine the frequency of oscillation of a three-stage ladder network oscillator in which C = 10 nF and R = 10 kΩ.

Solution

Using

$$f = \frac{1}{2\pi \times \sqrt{6}CR}$$

gives

$$f = \frac{1}{6.28 \times 2.45 \times 10 \times 10^{-9} \times 10 \times 10^{3}}$$

from which

$$f = \frac{1}{6.28 \times 2.45 \times 10^{-4}} = \frac{10^{4}}{15.386} = 647 \text{ Hz}$$

Wien bridge oscillator

An alternative approach to providing the phase shift required is the use of a Wien bridge network (Fig. 9.3). Like the *C–R* ladder, this network provides a phase shift which varies with frequency. The input signal is applied to A and B while the output is taken from C and D. At one particular frequency, the phase shift produced by the network will be exactly zero (i.e. the input and output signals will be in-phase). If we connect the network to an amplifier producing 0° phase shift which has sufficient gain to overcome the losses of the Wien bridge, oscillation will result.

The minimum amplifier gain required to sustain oscillation is given by:

$$A_v = 1 + \frac{C1}{C2} + \frac{R2}{R1}$$

In most cases, $C1 = C2$ and $R1 = R2$, hence the minimum amplifier gain will be 3.

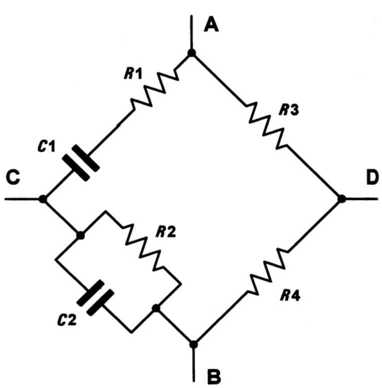

Figure 9.3 A Wien bridge network

The frequency at which the phase shift will be zero is given by:

$$f = \frac{1}{2\pi \times \sqrt{C1\,C2\,R1\,R2}}$$

When $R1 = R2$ and $C1 = C2$ the frequency at which the phase shift will be zero will be given by:

$$f = \frac{1}{2\pi \times \sqrt{C^2R^2}} = \frac{1}{2\pi CR}$$

where $R = R1 = R2$ and $C = C1 = C2$.

Example 9.2

Fig. 9.4 shows the circuit of a Wien bridge oscillator based on an operational amplifier. If $C1 = C2 = 100$ nF, determine the output frequencies produced by this arrangement (a) when $R1 = R2 = 1$ kΩ and (b) when $R1 = R2 = 6$ kΩ.

Solution

(a) When $R1 = R2 = 1$ kΩ

$$f = \frac{1}{2\pi CR}$$

where $R = R1 = R1$ and $C = C1 = C2$.

Thus

$$f = \frac{1}{6.28 \times 100 \times 10^{-9} \times 1 \times 10^{3}}$$

$$f = \frac{10^{4}}{6.28} = 1.59 \text{ kHz}$$

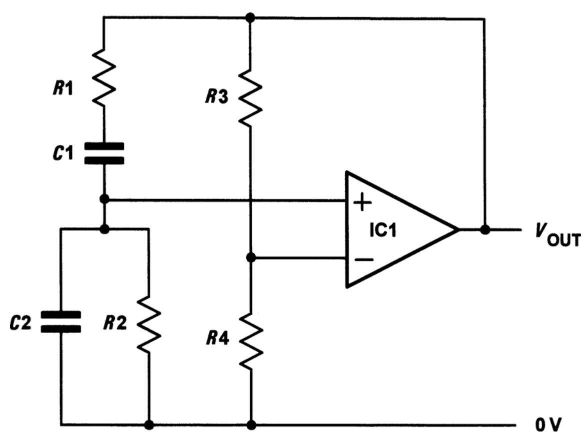

Figure 9.4 Sine wave oscillator based on a Wien bridge network (see Example 9.2)

(b) When $R1 = R1 = 6\ k\Omega$

$$f = \frac{1}{2\pi CR}$$

where $R = R1 = R1$ and $C = C1 = C2$.
Thus

$$f = \frac{1}{6.28 \times 100 \times 10^{-9} \times 6 \times 10^{3}}$$

$$f = \frac{10^{4}}{37.68} = 265\ \text{Hz}$$

Multivibrators

There are many occasions when we require a square wave output from an oscillator rather than a sine wave output. Multivibrators are a family of oscillator circuits that produce output waveforms consisting of one or more rectangular pulses. The term 'multivibrator' simply originates from the fact that this type of waveform is rich in harmonics (i.e. 'multiple vibrations').

Multivibrators use regenerative (i.e. positive) feedback; the active devices present within the oscillator circuit being operated as switches, being alternately cut-off and driven into saturation.

The principal types of multivibrator are:

(a) **astable multivibrators** that provide a continuous train of pulses (these are sometimes also referred to as free-running multivibrators);

(b) **monostable multivibrators** that produce a single output pulse (they have one stable state and are thus sometimes also referred to as 'one-shot');

(c) **bistable multivibrators** that have two stable states and require a trigger pulse or control signal to change from one state to another.

The astable multivibrator

Fig. 9.6 shows a classic form of astable multivibrator based on two transistors. Fig. 9.7 shows how this circuit can be redrawn in an arrangement that more closely resembles a two-stage common-emitter amplifier with its output connected back to its input. In Fig. 9.6, the values of the base resistors, $R3$ and $R4$, are such that

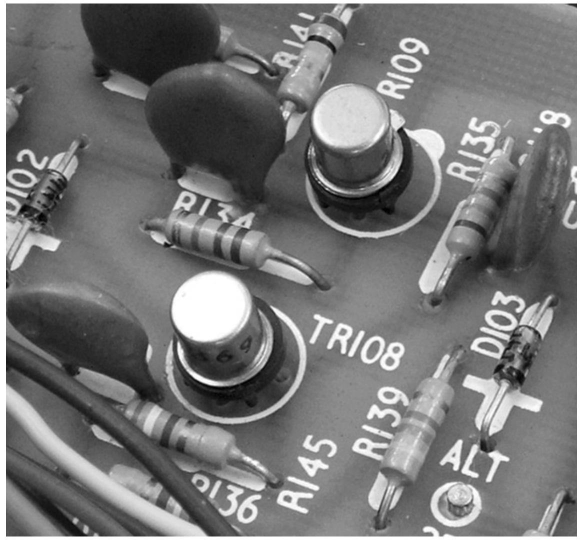

Figure 9.5 This high-speed bistable multivibrator uses two general-purpose silicon transistors and works at frequencies of up to 1 MHz triggered from an external signal

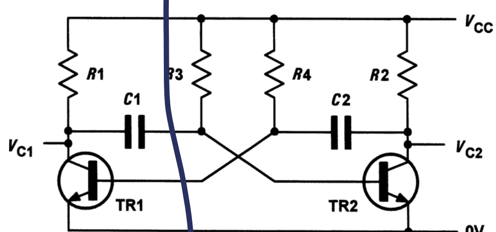

Figure 9.6 Astable multivibrator using BJTs

the sufficient base current will be available to completely saturate the respective transistor. The values of the collector load resistors, $R1$ and $R2$, are very much smaller than $R3$ and $R4$. When power is first applied to the circuit, assume that TR2 saturates before TR1 when the power is first applied (in practice one transistor would always saturate before the other due to variations in component tolerances and transistor parameters).

As TR2 saturates, its collector voltage will fall rapidly from $+V_{CC}$ to 0 V. This drop in voltage will be transferred to the base of TR1 via $C1$. This negative-going voltage will ensure that TR1 is initially placed in the non-conducting state. As long as TR1 remains cut-off, TR2 will continue to be saturated. During this time, $C1$ will charge via $R4$ and TR1's base voltage will rise exponentially

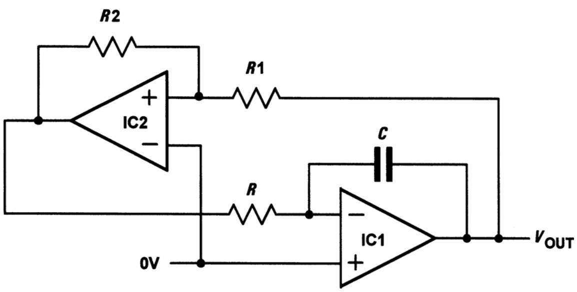

Figure 9.9 Astable oscillator using operational amplifiers

Assume that the output from IC2 is initially at, or near, $+V_{CC}$ and capacitor, *C*, is uncharged. The voltage at the output of IC2 will be passed, via *R*, to IC1. Capacitor, *C*, will start to charge and the output voltage of IC1 will begin to fall.

Eventually, the output voltage will have *fallen* to a value that causes the polarity of the voltage at the non-inverting input of IC2 to change from positive to negative. At this point, the output of IC2 will rapidly fall to $-V_{CC}$. Again, this voltage will be passed, via *R*, to IC1. Capacitor *C* will then start to charge in the other direction and the output voltage of IC1 will begin to rise.

Some time later, the output voltage will have *risen* to a value that causes the polarity of the non-inverting input of IC2 to revert to its original (positive) state and the cycle will continue indefinitely.

The upper threshold voltage (i.e. the maximum positive value for V_{out}) will be given by:

$$V_{UT} = V_{CC} \times \left(\frac{R1}{R2}\right)$$

The lower threshold voltage (i.e. the maximum negative value for V_{out}) will be given by:

$$V_{LT} = -V_{CC} \times \left(\frac{R1}{R2}\right)$$

Single-stage astable oscillator

A simple form of astable oscillator that produces a square wave output can be built using just one operational amplifier, as shown in Fig. 9.10. The circuit employs positive feedback with the output fed back to the non-inverting input via the potential divider formed by *R*1 and *R*2. This circuit can make a very simple square wave source with a frequency that can be made adjustable by replacing *R* with a variable or preset resistor.

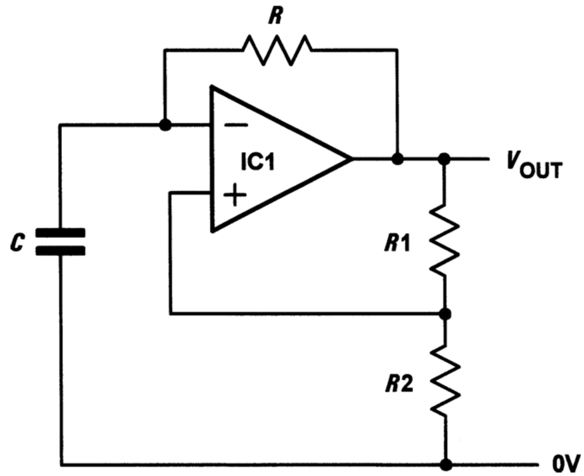

Figure 9.10 Single-stage astable oscillator using an operational amplifier

Assume that *C* is initially uncharged and the voltage at the inverting input is slightly less than the voltage at the non-inverting input. The output voltage will rise rapidly to $+V_{CC}$ and the voltage at the inverting input will begin to rise exponentially as capacitor *C* charges through *R*.

Eventually the voltage at the inverting input will have reached a value that causes the voltage at the inverting input to exceed that present at the non-inverting input. At this point, the output voltage will rapidly fall to $-V_{CC}$. Capacitor *C* will then start to charge in the other direction and the voltage at the inverting input will begin to fall exponentially.

Eventually, the voltage at the inverting input will have reached a value that causes the voltage at the inverting input to be less than that present at the non-inverting input. At this point, the output voltage will rise rapidly to $+V_{CC}$ once again and the cycle will continue indefinitely.

The upper threshold voltage (i.e. the maximum positive value for the voltage at the inverting input) will be given by:

$$V_{UT} = V_{CC} \times \left(\frac{R2}{R1+R2}\right)$$

The lower threshold voltage (i.e. the maximum negative value for the voltage at the inverting input) will be given by:

$$V_{LT} = -V_{CC} \times \left(\frac{R2}{R1+R2}\right)$$

Finally, the time for one complete cycle of the output waveform produced by the astable oscillator is given by:

$$T = 2CR\ln\left(1+2\left(\frac{R2}{R1}\right)\right)$$

Crystal controlled oscillators

A requirement of some oscillators is that they accurately maintain an exact frequency of oscillation. In such cases, a quartz crystal can be used as the frequency determining element. The quartz crystal (a thin slice of quartz in a hermetically sealed enclosure, see Fig. 9.11) vibrates whenever a potential difference is applied across its faces (this phenomenon is known as the piezoelectric effect). The frequency of oscillation is determined by the crystal's 'cut' and physical size.

Most quartz crystals can be expected to stabilize the frequency of oscillation of a circuit to within a few parts in a million. Crystals can be manufactured for operation in **fundamental mode** over a frequency range extending from 100 kHz to around 20 MHz and for **overtone** operation from 20 MHz to well over 100 MHz. Fig. 9.12 shows a simple crystal oscillator circuit in which the crystal provides feedback from the drain to the source of a junction gate FET.

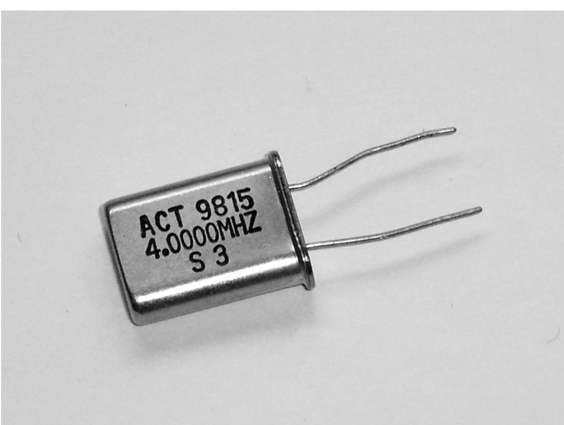

Figure 9.11 A quartz crystal (this crystal is cut to be resonant at 4 MHz and is supplied in an HC18 wire-ended package)

Practical oscillator circuits

Fig. 9.13 shows a practical sine wave oscillator based on a three-stage *C–R* ladder network. The circuit provides an output of approximately 1 V peak–peak at 1.97 kHz.

A practical Wien bridge oscillator is shown in Fig. 9.14. This circuit produces a sine wave output at 16 Hz. The output frequency can easily be varied by making *R*1 and *R*2 a 10 kΩ dual-gang potentiometer and connecting a fixed resistor of 680 Ω in series with each. In order to adjust the loop gain for an optimum sine wave output it may be necessary to make *R*3/*R*4 adjustable. One way of doing this is to replace both components with a 10 kΩ multi-turn potentiometer with the sliding contact taken to the inverting input of IC1.

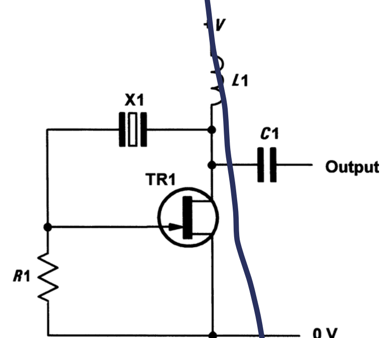

Figure 9.12 A simple JFET oscillator

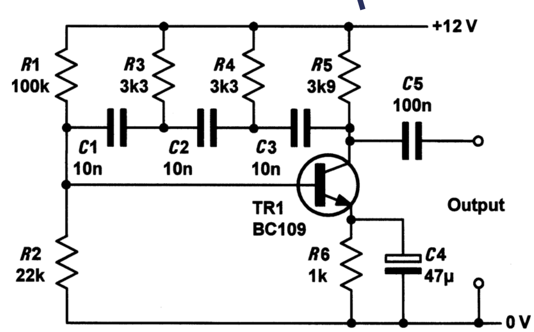

Figure 9.13 A practical sine wave oscillator based on a phase shift ladder network

Symbols and connections

The symbol for an operational amplifier is shown in Fig. 8.2. There are a few things to note about this. The device has two inputs and one output and no common connection. Furthermore, we often don't show the supply connections – it is often clearer to leave them out of the circuit altogether!

In Fig. 8.2, one of the inputs is marked '–' and the other is marked '+'. These polarity markings have nothing to do with the supply connections – they indicate the overall phase shift between each input and the output. The '+' sign indicates zero phase shift while the '–' sign indicates 180° phase shift. Since 180° phase shift produces an inverted waveform, the '–' input is often referred to as the **inverting input**. Similarly, the '+' input is known as the **non-inverting** input.

Most (but not all) operational amplifiers require a symmetrical supply (of typically ±6 V to ±15 V) which allows the output voltage to swing both positive (above 0 V) and negative (below 0 V). Fig. 8.3 shows how the supply connections would appear if we decided to include them. Note that we usually have two separate supplies; a positive supply and an equal, but opposite, negative supply. The common connection to these two supplies (i.e. the 0 V supply connection) acts as the **common rail** in our circuit. The input and output voltages are usually measured relative to this rail.

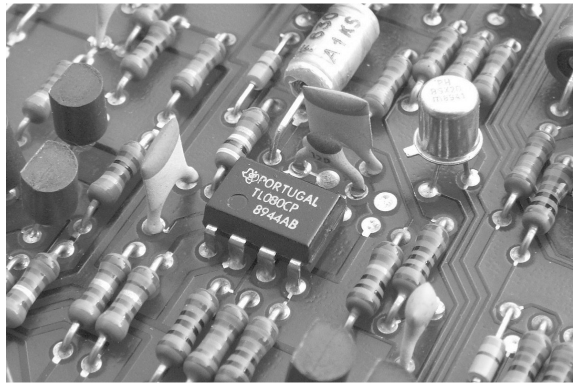

Figure 8.1 A typical operational amplifier. This device is supplied in an eight-pin dual-in-line (DIL) package. It has a JFET input stage and produces a typical open-loop voltage gain of 200,000

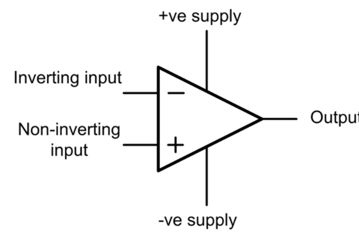

Figure 8.2 Symbol for an operational amplifier

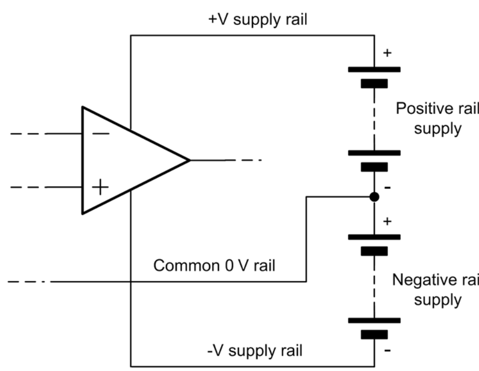

Figure 8.3 Supply connections for an operational amplifier

Operational amplifier parameters

Before we take a look at some of the characteristics of 'ideal' and 'real' operational amplifiers it is important to define some of the terms and parameters that we apply to these devices.

Open-loop voltage gain

The open-loop voltage gain of an operational amplifier is defined as the ratio of output voltage to input voltage measured with no feedback applied. In practice, this value is exceptionally high (typically greater than 100,000) but is liable to considerable variation from one device to another.

Open-loop voltage gain may thus be thought of as the 'internal' voltage gain of the device, thus:

$$A_{V(OL)} = \frac{V_{OUT}}{V_{IN}}$$

where $A_{V(OL)}$ is the open-loop voltage gain, V_{OUT} and V_{IN} are the output and input voltages, respectively, under open-loop conditions.

In linear voltage amplifying applications, a large amount of negative feedback will normally be applied and the open-loop voltage gain can be thought of as the internal voltage gain provided by the device.

The open-loop voltage gain is often expressed in **decibels** (**dB**) rather than as a ratio. In this case:

$$A_{V(OL)} = 20\log_{10}\frac{V_{OUT}}{V_{IN}}$$

Most operational amplifiers have open-loop voltage gains of 90 dB or more.

Closed-loop voltage gain

The closed-loop voltage gain of an operational amplifier is defined as the ratio of output voltage to input voltage measured with a small proportion of the output fed-back to the input (i.e. with feedback applied). The effect of providing negative feedback is to reduce the loop voltage gain to a value that is both predictable and manageable. Practical closed-loop voltage gains range from one to several thousand but note that high values of voltage gain may make unacceptable restrictions on bandwidth (see later).

Closed-loop voltage gain is once again the ratio of output voltage to input voltage but with negative feedback applied, hence:

$$A_{V(CL)} = \frac{V_{OUT}}{V_{IN}}$$

where $A_{V(CL)}$ is the open-loop voltage gain, V_{OUT} and V_{IN} are the output and input voltages, respectively, under closed-loop conditions. The closed-loop voltage gain is normally very much less than the open-loop voltage gain.

Example 8.1

An operational amplifier operating with negative feedback produces an output voltage of 2 V when supplied with an input of 400 μV. Determine the value of closed-loop voltage gain.

Solution

Now:

$$A_{V(CL)} = \frac{V_{OUT}}{V_{IN}}$$

Thus:

$$A_{V(CL)} = \frac{2}{400\times10^{-6}} = \frac{2\times10^{6}}{400} = 5{,}000$$

Expressed in decibels (rather than as a ratio) this is:

$$A_{V(CL)} = 20\log_{10}(5{,}000) = 20\times3.7 = 74\text{ dB}$$

Input resistance

The input resistance of an operational amplifier is defined as the ratio of input voltage to input current expressed in ohms. It is often expedient to assume that the input of an operational amplifier is purely resistive, though this is not the case at high frequencies where shunt capacitive reactance may become significant. The input resistance of operational amplifiers is very much dependent on the semiconductor technology employed. In practice values range from about 2 MΩ for common bipolar types to over 10^{12} Ω for FET and CMOS devices.

Input resistance is the ratio of input voltage to input current:

$$R_{IN} = \frac{V_{IN}}{I_{IN}}$$

where R_{IN} is the input resistance (in ohms), V_{IN} is the input voltage (in volts) and I_{IN} is the input current (in amps). Note that we usually assume that the input of an operational amplifier is purely resistive though this may not be the case at high frequencies where shunt capacitive reactance may become significant.

The input resistance of operational amplifiers is very much dependent on the semiconductor technology employed. In practice, values range from about 2 MΩ for bipolar operational amplifiers to over 10^{12} Ω for CMOS devices.

Example 8.2

An operational amplifier has an input resistance of 2 MΩ. Determine the input current when an input voltage of 5 mV is present.

Solution

Now:

$$R_{IN} = \frac{V_{IN}}{I_{IN}}$$

thus

$$I_{IN} = \frac{V_{IN}}{R_{IN}} = \frac{5 \times 10^{-3}}{2 \times 10^{6}} = 2.5 \times 10^{-9}\,\text{A} = 2.5\ \text{nA}$$

Output resistance

The output resistance of an operational amplifier is defined as the ratio of open-circuit output voltage to short-circuit output current expressed in ohms. Typical values of output resistance range from less than 10 Ω to around 100 Ω, depending upon the configuration and amount of feedback employed.

Output resistance is the ratio of open-circuit output voltage to short-circuit output current, hence:

$$R_{OUT} = \frac{V_{OUT(OC)}}{I_{OUT(SC)}}$$

where R_{OUT} is the output resistance (in ohms), $V_{OUT(OC)}$ is the open-circuit output voltage (in volts) and $I_{OUT(SC)}$ is the short-circuit output current (in amps).

Input offset voltage

An ideal operational amplifier would provide zero output voltage when 0 V difference is applied to its inputs. In practice, due to imperfect internal balance, there may be some small voltage present at the output. The voltage that must be applied differentially to the operational amplifier input in order to make the output voltage exactly zero is known as the input offset voltage.

Input offset voltage may be minimized by applying relatively large amounts of negative feedback or by using the offset null facility provided by a number of operational amplifier devices. Typical values of input offset voltage range from 1 mV to 15 mV. Where a.c. rather than d.c. coupling is employed, offset voltage is not normally a problem and can be happily ignored.

Full-power bandwidth

The full-power bandwidth for an operational amplifier is equivalent to the frequency at which the maximum undistorted peak output voltage swing falls to 0.707 of its low-frequency (d.c.) value (the sinusoidal input voltage remaining constant). Typical full-power bandwidths range from 10 kHz to over 1 MHz for some high-speed devices.

Slew rate

Slew rate is the rate of change of output voltage with time, when a rectangular step input voltage is applied (as shown in Fig. 8.4). The slew rate of an operational amplifier is the rate of change of output voltage with time in response to a perfect step-function input. Hence:

$$\text{Slew rate} = \frac{\Delta V_{OUT}}{\Delta t}$$

where ΔV_{OUT} is the change in output voltage (in volts) and Δt is the corresponding interval of time (in seconds).

Slew rate is measured in V/s (or V/μs) and typical values range from 0.2 V/μs to over 20 V/μs. Slew rate imposes a limitation on circuits in which large amplitude pulses rather than small amplitude sinusoidal signals are likely to be encountered.

Operational amplifier characteristics

Having defined the parameters that we use to describe operational amplifiers we shall now

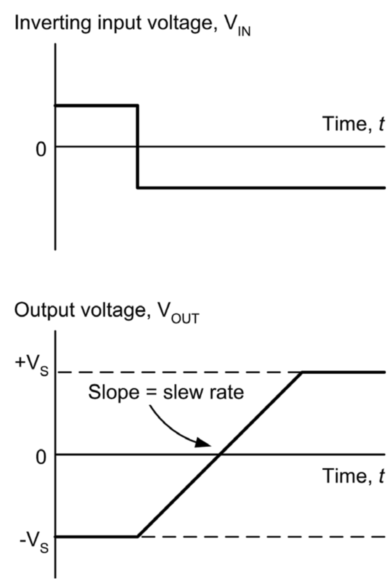

Figure 8.4 Slew rate for an operational amplifier

consider the desirable characteristics for an 'ideal' operational amplifier. These are:

(a) The open-loop voltage gain should be very high (ideally infinite).
(b) The input resistance should be very high (ideally infinite).
(c) The output resistance should be very low (ideally zero).
(d) Full-power bandwidth should be as wide as possible.
(e) Slew rate should be as large as possible.
(f) Input offset should be as small as possible.

The characteristics of most modern integrated circuit operational amplifiers (i.e. 'real' operational amplifiers) come very close to those of an 'ideal' operational amplifier, as witnessed by the data shown in Table 8.1.

Table 8.1 Comparison of operational amplifier parameters for 'ideal' and 'real' devices

Parameter	Ideal	Real
Voltage gain	Infinite	100,000
Input resistance	Infinite	100 MΩ
Output resistance	Zero	20 Ω
Bandwidth	Infinite	2 MHz
Slew rate	Infinite	10 V/μs
Input offset	Zero	Less than 5 mV

Example 8.3

A perfect rectangular pulse is applied to the input of an operational amplifier. If it takes 4 μs for the output voltage to change from –5 V to +5 V, determine the slew rate of the device.

Solution

The slew rate can be determined from:

$$\text{Slew rate} = \frac{\Delta V_{OUT}}{\Delta t} = \frac{10\ \text{V}}{4\ \mu\text{s}} = 2.5\ \text{V}/\mu\text{s}$$

Example 8.4

A wideband operational amplifier has a slew rate of 15 V/μs. If the amplifier is used in a circuit with a voltage gain of 20 and a perfect step input of 100 mV is applied to its input, determine the time taken for the output to change level.

Solution

The output voltage change will be 20 × 100 = 2,000 mV (or 2 V). Re-arranging the formula for slew rate gives:

$$\Delta t = \frac{\Delta V_{OUT}}{\text{Slew rate}} = \frac{2\ \text{V}}{15\ \text{V}/\mu\text{s}} = 0.133\ \mu\text{s}$$

Operational amplifier applications

Table 8.2 shows abbreviated data for some common types of integrated circuit operational amplifier together with some typical applications.

Example 8.5

Which of the operational amplifiers in Table 8.2 would be most suitable for each of the following applications:

(a) amplifying the low-level output from a piezoelectric vibration sensor
(b) a high-gain amplifier that can be used to faithfully amplify very small signals

Table 8.2 Some common examples of integrated circuit operational amplifiers

Device	Type	Open-loop voltage gain (dB)	Input bias current	Slew rate (V/μs)	Application
AD548	Bipolar	100 min.	0.01 nA	1.8	Instrumentation amplifier
AD711	FET	100	25 pA	20	Wideband amplifier
CA3140	CMOS	100	5 pA	9	Low-noise wideband amplifier
LF347	FET	110	50 pA	13	Wideband amplifier
LM301	Bipolar	88	70 nA	0.4	General-purpose operational amplifier
LM348	Bipolar	96	30 nA	0.6	General-purpose operational amplifier
TL071	FET	106	30 pA	13	Wideband amplifier
741	Bipolar	106	80 pA	0.5	General-purpose operational amplifier

(c) a low-frequency amplifier for audio signals.

Solution

(a) AD548 (this operational amplifier is designed for use in instrumentation applications and it offers a very low input offset current which is important when the input is derived from a piezoelectric transducer).
(b) CA3140 (this is a low-noise operational amplifier that also offers high gain and fast slew rate).
(c) LM348 or LM741 (both are general-purpose operational amplifiers and are ideal for non-critical applications such as audio amplifiers).

Gain and bandwidth

It is important to note that the product of gain and bandwidth is a constant for any particular operational amplifier. Hence, an increase in gain can only be achieved at the expense of bandwidth, and vice versa.

Fig. 8.5 shows the relationship between voltage gain and bandwidth for a typical operational amplifier (note that the axes use logarithmic rather than linear scales). The open-loop voltage gain (i.e. that obtained with no feedback applied) is 100,000 (or 100 dB) and the bandwidth obtained in this condition is a mere 10 Hz. The effect of applying increasing amounts of negative feedback (and consequently reducing the gain to a more manageable amount) is that the bandwidth increases in direct proportion.

The frequency response curves in Fig. 8.5 show the effect on the bandwidth of making the closed-loop gains equal to 10,000, 1,000, 100, and 10. Table 8.3 summarizes these results. You should also note that the (gain × bandwidth) product for this amplifier is 1×10^6 Hz (i.e. 1 MHz).

We can determine the bandwidth of the amplifier when the closed-loop voltage gain is set to 46 dB by constructing a line and noting the intercept point on the response curve. This shows that the bandwidth will be 10 kHz (note that, for this operational amplifier, the (gain × bandwidth) product is 2×10^6 Hz (or 2 MHz).

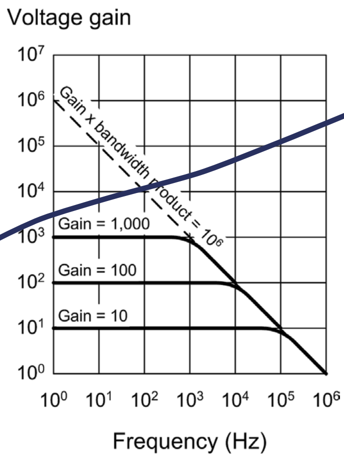

Figure 8.5 Frequency response curves for an operational amplifier

Table 8.3 Corresponding values of voltage gain and bandwidth for an operational amplifier with a gain × bandwidth product of 1×10^6

Voltage gain (A_V)	Bandwidth
1	d.c. to 1 MHz
10	d.c. to 100 kHz
100	d.c. to 10 kHz
1,000	d.c. to 1 kHz
10,000	d.c. to 100 Hz
100,000	d.c. to 10 Hz

Inverting amplifier with feedback

Fig. 8.6 shows the circuit of an inverting amplifier with negative feedback applied. For the sake of our explanation we will assume that the operational amplifier is 'ideal'. Now consider what happens when a small positive input voltage is applied. This voltage (V_{IN}) produces a current (I_{IN}) flowing in the input resistor $R1$.

Since the operational amplifier is 'ideal' we will assume that:

(a) the input resistance (i.e. the resistance that appears between the inverting and non-inverting input terminals, R_{IC}) is infinite
(b) the open-loop voltage gain (i.e. the ratio of V_{OUT} to V_{IN} with no feedback applied) is infinite.

As a consequence of (a) and (b):

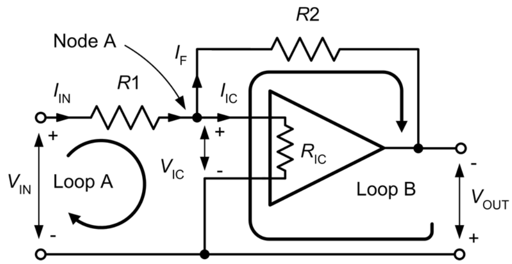

Figure 8.6 Operational amplifier with negative feedback applied

(i) the voltage appearing between the inverting and non-inverting inputs (V_{IC}) will be zero, and

(ii) the current flowing into the chip (I_{IC}) will be zero (recall that $I_{IC} = V_{IC}/R_{IC}$ and R_{IC} is infinite)

Applying Kirchhoff's Current Law at node A gives:

$I_{IN} = I_{IC} + I_F$ but $I_{IC} = 0$ thus $I_{IN} = I_F$ (1)

(this shows that the current in the feedback resistor, $R2$, is the same as the input current, I_{IN}).

Applying Kirchhoff's Voltage Law to loop A gives:

$V_{IN} = (I_{IN} \times R1) + V_{IC}$

but $V_{IC} = 0$ thus $V_{IN} = I_{IN} \times R1$ (2)

Using Kirchhoff's Voltage Law in loop B gives:

$V_{OUT} = -V_{IC} + (I_F \times R2)$

but $V_{IC} = 0$ thus $V_{OUT} = I_F \times R2$ (3)

Combining (1) and (3) gives:

$V_{OUT} = I_{IN} \times R2$ (4)

The voltage gain of the stage is given by:

$$A_v = \frac{V_{OUT}}{V_{IN}} \quad (5)$$

Combining (4) and (2) with (5) gives:

$$A_v = \frac{I_{IN} \times R2}{I_{IN} \times R1} = \frac{R2}{R1}$$

To preserve symmetry and minimize offset voltage, a third resistor is often included in series with the non-inverting input. The value of this resistor should be equivalent to the parallel combination of $R1$ and $R2$. Hence:

$$R3 = \frac{R1 \times R2}{R1 + R2}$$

From this point onwards (and to help you remember the function of the resistors) we shall refer to the input resistance as R_{IN} and the feedback resistance as R_F (instead of the more general and less meaningful $R1$ and $R2$, respectively).

Operational amplifier configurations

The three basic configurations for operational voltage amplifiers, together with the expressions for their voltage gain, are shown in Fig. 8.7. Supply rails have been omitted from these diagrams for clarity but are assumed to be symmetrical about 0 V.

All of the amplifier circuits described previously have used direct coupling and thus have frequency response characteristics that extend to d.c. This, of course, is undesirable for many applications, particularly where a wanted a.c. signal may be superimposed on an unwanted d.c. voltage level or when the bandwidth of the amplifier greatly exceeds that of the signal that it is required to amplify. In such cases, capacitors of appropriate value may be inserted in series with the input resistor, R_{IN}, and in parallel with the feedback resistor, R_F, as shown in Fig. 8.8.

The value of the input and feedback capacitors, C_{IN} and C_F respectively, are chosen so as to roll-off the frequency response of the amplifier at the desired lower and upper cut-off frequencies, respectively. The effect of these two capacitors on an operational amplifier's frequency response is shown in Fig. 8.9.

By selecting appropriate values of capacitor, the frequency response of an inverting operational voltage amplifier may be very easily tailored to suit a particular set of requirements.

The lower cut-off frequency is determined by the value of the input capacitance, C_{IN}, and input resistance, R_{IN}. The lower cut-off frequency is given by:

$$f_1 = \frac{1}{2\pi C_{IN} R_{IN}} = \frac{0.159}{C_{IN} R_{IN}}$$

where f_1 is the lower cut-off frequency in hertz, C_{IN} is in farads and R_{IN} is in ohms.

Provided the upper frequency response it not limited by the gain × bandwidth product, the

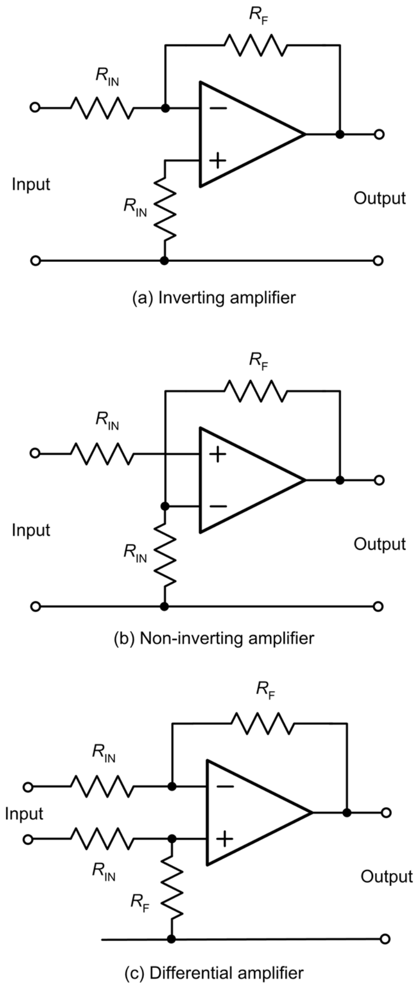

Figure 8.7 The three basic configurations for operational voltage amplifiers

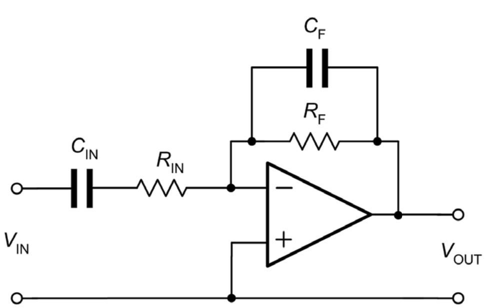

Figure 8.8 Adding capacitors to modify the frequency response of an inverting operational amplifier

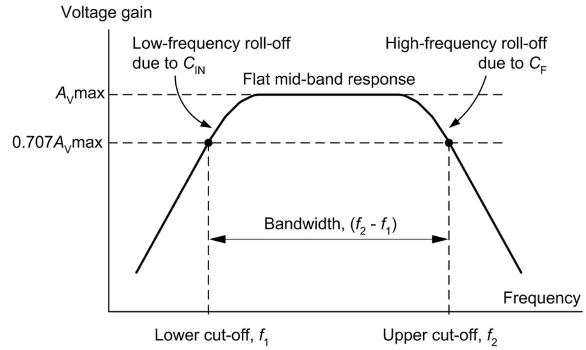

Figure 8.9 Effect of adding capacitors, C_{IN} and C_F, to modify the frequency response of an operational amplifier

upper cut-off frequency will be determined by the feedback capacitance, C_F, and feedback resistance, R_F, such that:

$$f_2 = \frac{1}{2\pi C_F R_F} = \frac{0.159}{C_F R_F}$$

where f_2 is the upper cut-off frequency in hertz, C_F is in farads and R_2 is in ohms.

Example 8.6

An inverting operational amplifier is to operate according to the following specification:

Voltage gain = 100
Input resistance (at mid-band) = 10 kΩ
Lower cut-off frequency = 250 Hz
Upper cut-off frequency = 15 kHz

Devise a circuit to satisfy the above specification using an operational amplifier.

Solution

To make things a little easier, we can break the problem down into manageable parts. We shall base our circuit on a single operational amplifier configured as an inverting amplifier with capacitors to define the upper and lower cut-off frequencies, as shown in Fig. 8.9.

The nominal input resistance is the same as the value for R_{IN}. Thus:

$$R_{IN} = 10\ \text{k}\Omega$$

To determine the value of R_F we can make use of the formula for mid-band voltage gain:

$$A_v = \frac{R2}{R1}$$

thus $R2 = A_V \times R1 = 100 \times 10\ \text{k}\Omega = 100\,\text{k}\Omega$

To determine the value of C_{IN} we will use the formula for the low-frequency cut-off:

$$f_1 = \frac{0.159}{C_{IN}R_{IN}}$$

from which:

$$C_{IN} = \frac{0.159}{f_1 R_{IN}} = \frac{0.159}{250 \times 10 \times 10^3}$$

hence:

$$C_{IN} = \frac{0.159}{2.5 \times 10^6} = 63 \times 10^{-9}\text{F} = 63\ \text{nF}$$

Finally, to determine the value of C_F we will use the formula for high-frequency cut-off:

$$f_2 = \frac{0.159}{C_F R_F}$$

from which:

$$C_F = \frac{0.159}{f_2 R_{IN}} = \frac{0.159}{15 \times 10^3 \times 100 \times 10^3}$$

hence:

$$C_F = \frac{0.159}{1.5 \times 10^9} = 0.106 \times 10^{-9}\text{F} = 106\ \text{pF}$$

For most applications the nearest preferred values (68 nF for C_{IN} and 100 pF for C_F) would be perfectly adequate. The complete circuit of the operational amplifier stage is shown in Fig. 8.10.

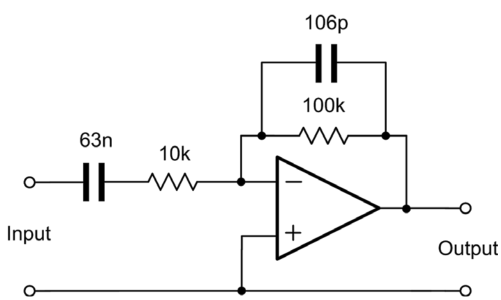

Figure 8.10 See Example 8.6. This operational amplifier has a mid-band voltage gain of 10 over the frequency range 250 Hz to 15 kHz

Operational amplifier circuits

As well as their application as a general-purpose amplifying device, operational amplifiers have a number of other uses, including voltage followers, differentiators, integrators, comparators and summing amplifiers. We shall conclude this section by taking a brief look at each of these applications.

Voltage followers

A voltage follower using an operational amplifier is shown in Fig. 8.11. This circuit is essentially an inverting amplifier in which 100% of the output is fed back to the input. The result is an amplifier that has a voltage gain of 1 (i.e. unity), a very high input resistance and a very high output resistance. This stage is often referred to as a buffer and is used for matching a high-impedance circuit to a low-impedance circuit.

Typical input and output waveforms for a voltage follower are shown in Fig. 8.12. Notice how the input and output waveforms are both in-phase (they rise and fall together) and that they are identical in amplitude.

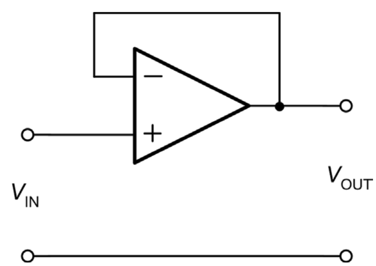

Figure 8.11 A voltage follower

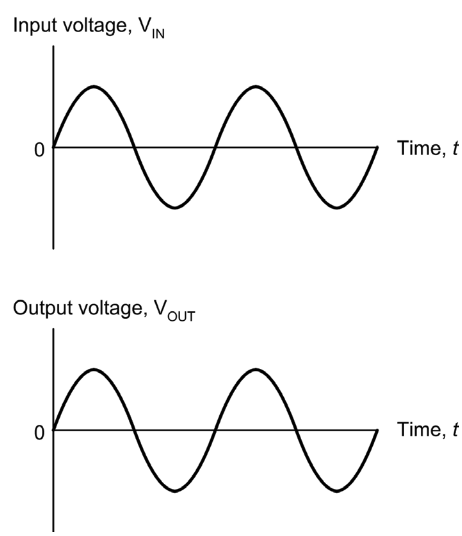

Figure 8.12 Typical input and output waveforms for a voltage follower

Differentiators

A differentiator using an operational amplifier is shown in Fig. 8.13. A differentiator produces an output voltage that is equivalent to the rate of change of its input. This may sound a little complex but it simply means that if the input voltage remains constant (i.e. if it isn't changing) the output also remains constant. The faster the input voltage changes the greater will the output be. In mathematics this is equivalent to the differential function.

Typical input and output waveforms for a differentiator are shown in Fig. 8.14. Notice how the square wave input is converted to a train of short duration pulses at the output. Note also that the output waveform is inverted because the signal has been applied to the inverting input of the operational amplifier.

Integrators

An integrator using an operational amplifier is shown in Fig. 8.15. This circuit provides the opposite function to that of a differentiator (see earlier) in that its output is equivalent to the area under the graph of the input function rather than its rate of change. If the input voltage remains constant (and is other than 0 V) the output voltage will ramp up or down according to the polarity of the input. The longer the input voltage remains at a particular value the larger the value of output voltage (of either polarity) will be produced.

Typical input and output waveforms for an integrator are shown in Fig. 8.16. Notice how the square wave input is converted to a wave that has a triangular shape. Once again, note that the output waveform is inverted.

Comparators

A comparator using an operational amplifier is shown in Fig. 8.17. Since no negative feedback has been applied, this circuit uses the maximum gain of the operational amplifier. The output voltage produced by the operational amplifier will

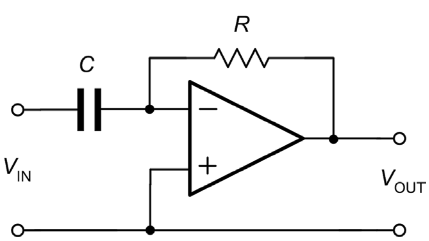

Figure 8.13 A differentiator

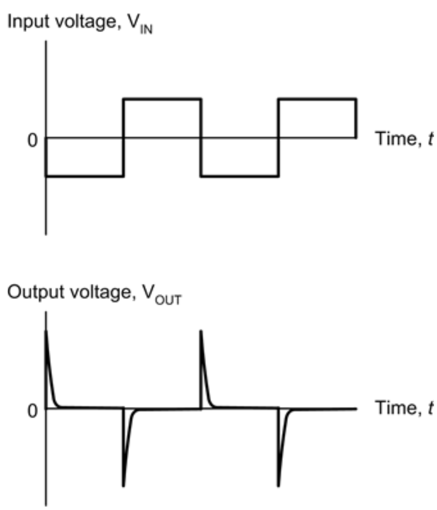

Figure 8.14 Typical input and output waveforms for a differentiator

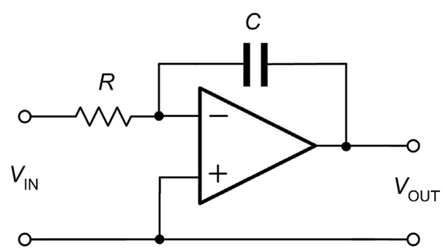

Figure 8.15 An integrator

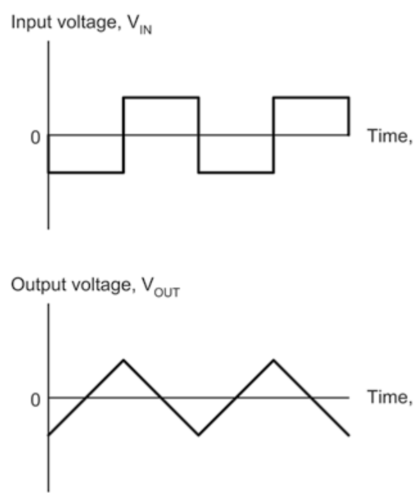

Figure 8.16 Typical input and output waveforms for an integrator

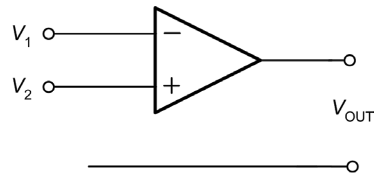

Figure 8.17 A comparator

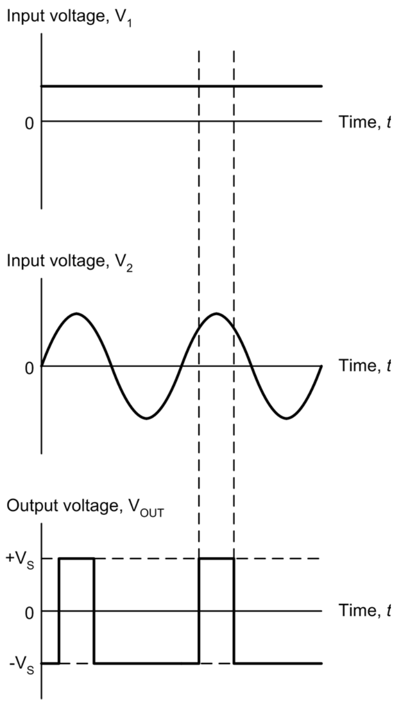

Figure 8.18 Typical input and output waveforms for a comparator

thus rise to the maximum possible value (equal to the positive supply rail voltage) whenever the voltage present at the non-inverting input exceeds that present at the inverting input. Conversely, the output voltage produced by the operational amplifier will fall to the minimum possible value (equal to the negative supply rail voltage) whenever the voltage present at the inverting input exceeds that present at the non-inverting input.

Typical input and output waveforms for a comparator are shown in Fig. 8.18. Notice how the output is either +15 V or –15 V depending on the relative polarity of the two inputs. A typical application for a comparator is that of comparing a signal voltage with a reference voltage. The output will go high (or low) in order to signal the result of the comparison.

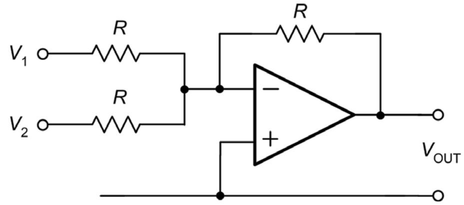

Figure 8.19 A summing amplifier

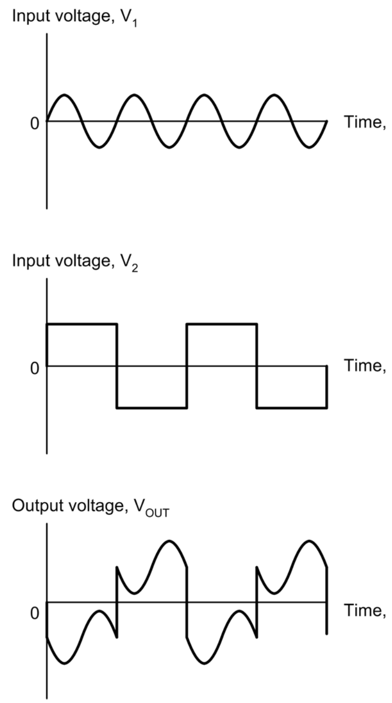

Figure 8.20 Typical input and output waveforms for a summing amplifier

Summing amplifiers

A summing amplifier using an operational amplifier is shown in Fig. 8.19. This circuit produces an output that is the sum of its two input voltages. However, since the operational amplifier is connected in inverting mode, the output voltage is given by:

$$V_{OUT} = -(V_1 + V_2)$$

where V_1 and V_2 are the input voltages (note that all of the resistors used in the circuit have the same value). Typical input and output waveforms for a summing amplifier are shown in Fig. 8.20. A typical application is that of 'mixing' two input signals to produce an output voltage that is the sum of the two.