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Rashtreeya Vidyalaya Institute of Technology and Management (RVITM), Bengaluru

BASIC ELECTRONICS & COMMUNICATION ENGINEERING (21ELN14/24)

SEMESTER-I

Module-1

Electronic Circuits

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DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

Module 1 **Electronic Circuits**	RBT Levels
Power Supplies – Block diagram, Rectifiers, Reservoir and smoothing circuits, Full-wave rectifiers, Bi-phase rectifier circuits, Bridge rectifier circuits, Voltage regulators, Output resistance and voltage regulation, Voltage multipliers. Amplifiers – Types of amplifiers, Gain, Input and output resistance, Frequency response, Bandwidth, Phase shift, Negative feedback, Multi-stage amplifiers. Operational amplifiers - Operational amplifier parameters, Operational amplifier characteristics, Operational amplifier configurations, Operational amplifier circuits. Oscillators – Positive feedback, Conditions for oscillation, Ladder network oscillator, Wein bridge oscillator, Multivibrators, Single-stage astable oscillator, Crystal controlled oscillators. (Only Concepts, working, and waveforms. No mathematical derivations)	L1, L2

POWER SUPPLY

1.1 Power supply

The block diagram of a d.c. power supply is shown in Fig. 1.1. The main input is at a relatively high voltage, a step-down transformer of appropriate turns ratio is used to convert this to a low voltage. The a.c. output from the transformer secondary is then rectified using conventional silicon rectifier diodes to produce an unsmoothed (sometimes referred to as **pulsating d.c.**) output. This is then smoothed and filtered before being applied to a circuit which will **regulate** (or **stabilize**) the output voltage so that it remains relatively constant despite variations in both load current and incoming mains voltage.

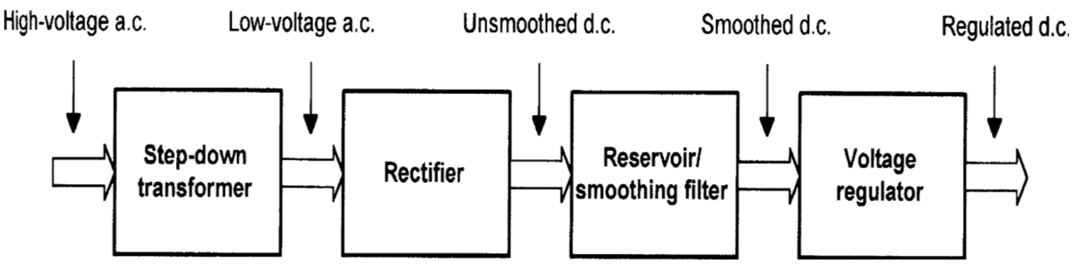


Fig. 1.1 Block diagram of a d.c. power supply

Fig. 1.2 shows how some of the electronic components that we have already met can be used in the realization of the block diagram in Fig. 1.1.

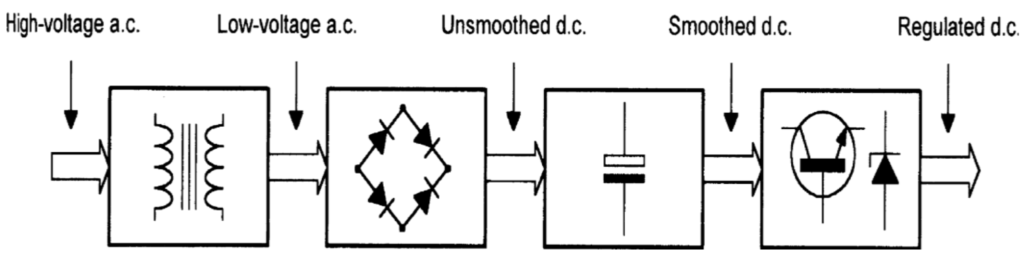


Fig. 1.2 Block diagram of a d.c. power supply showing principal components

The iron-cored step-down transformer feeds a rectifier arrangement (often based on a bridge circuit). The output of the rectifier is then applied to a high-value **reservoir** capacitor. This capacitor stores a considerable amount of charge and is being constantly topped-up by the rectifier arrangement. The capacitor also helps to smooth out the voltage pulses produced by the rectifier. Finally, a stabilizing circuit (often based on a **series transistor regulator** and a zener diode **voltage reference**) provides a constant output voltage.

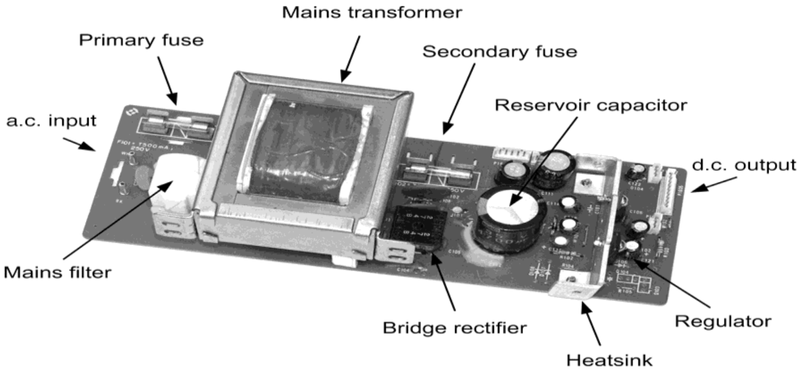


Fig. 1.3 A simple d.c. power supply

1.2 Rectifiers

- Semiconductor diodes are commonly used to convert alternating current (a.c.) to direct current (d.c), in which case they are referred to as **rectifiers**.
- The simplest form of rectifier circuit makes use of a single diode and, since it operates on only either positive or negative half-cycles of the supply, it is known as a **half-wave** rectifier.
- Fig. 1.4 shows a simple half-wave rectifier circuit.Main voltage (220 to 240 V) is applied to the primary of a step-down transformer (T1). The secondary of T1 steps down the 240 V r.m.s. to 12 V r.m.s. (the turns ratio of T1 will thus be 240/12 or 20:1). Diode D1 will only allow the current to flow in the direction shown (i.e. from cathode to anode). D1 will be forward biased during each positive half-cycle (relative to common) and will effectively behave like a closed switch. When the circuit current tries to flow in the opposite direction, the voltage bias across the diode will be reversed, causing the diode to act like an open switch
- The switching action of D1 results in a pulsating output voltage which is developed

across the load resistor (R_L). Since the mains supply is at 50 Hz, the pulses of voltage developed across R_L will also be at 50 Hz even if only half the a.c. cycle is present. During the positive half-cycle, the diode will drop the 0.6 V to 0.7 V forward threshold

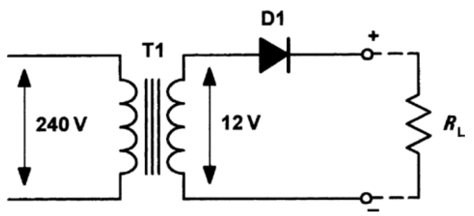


Fig. 1.4 A simple half-wave rectifier circuit

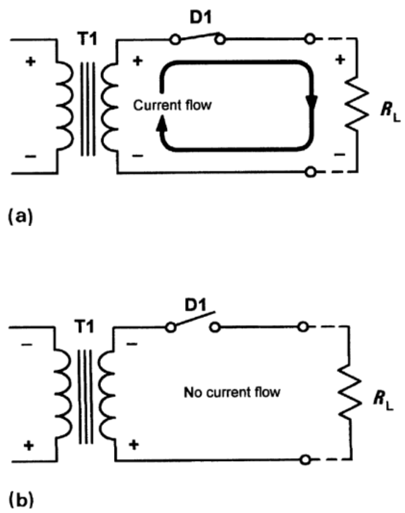


Fig. 1.5 (a) Half-wave rectifier circuit with D1 conducting (positive-going half-cycles of secondary voltage); (b) half-wave rectifier with D1 not conducting (negative-going half-cycles of secondary voltage)

voltage normally associated with silicon diodes. However, during the negative half-cycle the peak a.c. voltage will be dropped across D1 when it is reverse biased. This is an important consideration when selecting a diode for a particular application. Assuming that the secondary of T1 provides 12 V r.m.s., the peak voltage output from the transformer's secondary winding will be given by:

$$V_{pk} = 1.414 \times V_{r.m.s.} = 1.414 \times 12\ V = 16.97\ V \qquad (1)$$

The peak voltage applied to D1 will thus be approximately 17 V. The negative half-cycles are blocked by D1 and thus only the positive half- cycles appear across R_L. Note, however, that the actual peak voltage across R_L will be the 17 V positive peak being supplied from the secondary on T1, *minus* the 0.7 V forward threshold voltage dropped by D1. In other words, positive half-cycle pulses having a peak amplitude of 16.3 V will appear across R_L.

Example 1.1

A mains transformer having a turns ratio of 44:1 is connected to a 220 V r.m.s. mains supply. If the secondary output is applied to a half-wave rectifier, determine the peak voltage that will appear across a load.

Solution

The r.m.s. secondary voltage will be given by:

$$V_S = V_P / 44 = 220 / 44 = 5\ V$$

The peak voltage developed after rectification will be given by:

$$V_{PK} = 1.414 \times 5\ V = 7.07\ V$$

Assuming that the diode is a silicon device with a forward voltage drop of 0.6 V, the actual peak voltage dropped across the load will be:

$$V_L = 7.07\ V - 0.6\ V = 6.47V$$

1.3 Reservoir and smoothing circuits

Fig. 1.6 shows a considerable improvement to the circuit of Fig. 1.4. The capacitor, C 1, has been added to ensure that the output voltage remains at, or near, the peak voltage even when the diode is not conducting. When the primary voltage is first applied to T1, the first positive half-cycle output from the secondary will charge $C1$ to the peak value seen across R_L. Hence $C1$ charges to 16.3 V at the peak of the positive half-cycle. Because $C1$ and R are in parallel, the voltage across R_L will be the same as that across $C1$.

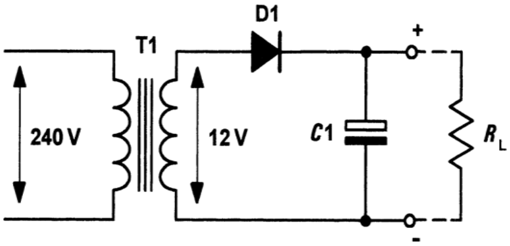


Fig. 1.6 A simple half-wave rectifier circuit with reservoir capacitor

The time required for $C1$ to charge to the maximum (peak) level is determined by the charging circuit time constant (the series resistance multiplied by the capacitance value). In this circuit, the series resistance comprises the secondary winding resistance together with the forward resistance of the diode and the (minimal) resistance of the wiring and connections. Hence $C1$ charges very rapidly as soon as D1 starts to conduct.

The time required for $C1$ to discharge is, in contrast, very much greater. The discharge time constant is determined by the capacitance value and the load resistance, R_L. In practice, R_L is very much larger than the resistance of the secondary circuit and hence $C1$ takes an appreciable time to discharge. During this time, D1 will be reverse biased and will thus be held in its non-conducting state. As a consequence, the only discharge path for $C1$ is through R_L.

$C1$ is referred to as a **reservoir** capacitor. It stores charge during the positive half- cycles of secondary voltage and releases it during the negative half-cycles. The circuit of Fig. 1.6 is thus able to maintain a reasonably constant output voltage across R_L. Even so, *C1* will discharge by a small amount during the negative half-cycle periods from the transformer secondary

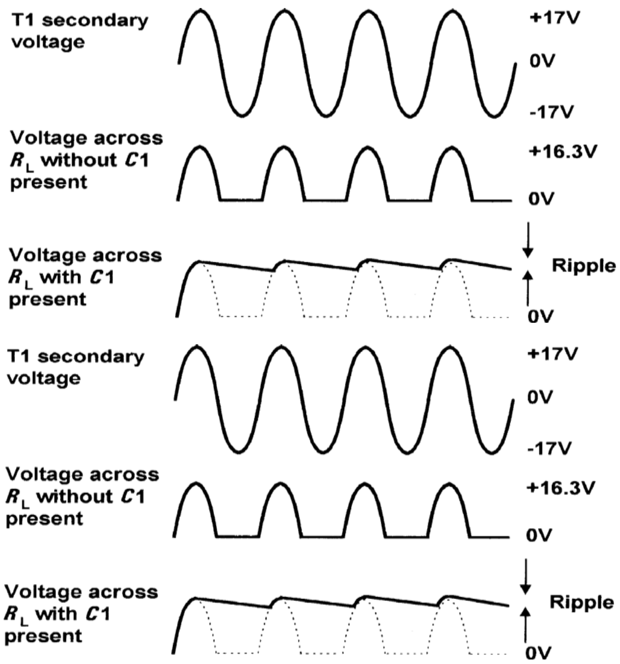


Fig. 1.7 A simple half-wave rectifier circuit with reservoir capacitor

Fig. 1.7 shows the secondary voltage waveform together with the voltage developed across R_L with and without $C1$ present. This gives rise to a small variation in the d.c. output voltage (known as **ripple**). Since ripple is undesirable we must take additional precautions to reduce it. One obvious method of reducing the amplitude of the ripple is that of simply increasing the discharge time constant. This can be achieved either by increasing the value of $C1$ or by increasing the resistance value of R_L. In practice, however, the latter is not really an option because R_L is the effective resistance of the circuit being supplied and we don't usually have the ability to change it! Increasing the value of *C*1 is a more practical alternative and very large capacitor values (often in excess of 4,700 μF) are typical.

Fig. 1.8 shows a further refinement of the simple power supply circuit. This circuit employs two additional components, *R*1 and *C*1, which act as a filter to remove the ripple. The value of *C*1 is chosen so that the component exhibits a negligible reactance at the ripple frequency (50 Hz for a half-wave rectifier or 100 Hz for a full-wave rectifier). In effect, *R*1 and *C*1 act like a potential divider. The amount of ripple is reduced by an approximate factor equal to:

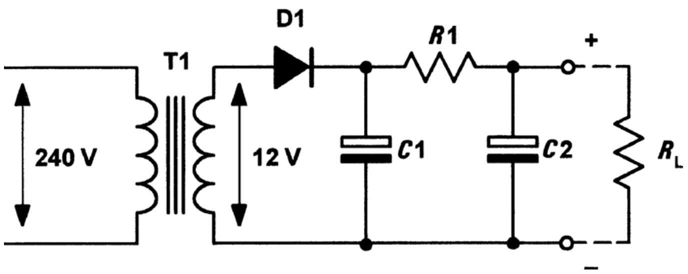


Fig. 1.8 Half-wave rectifier circuit with *R*–*C* smoothing filter

1.4 Full-wave rectifiers

Unfortunately, the half-wave rectifier circuit is relatively inefficient as conduction takes place only on alternate half-cycles. A better rectifier arrangement would make use of both positive *and* negative half-cycles. These **full-wave rectifier** circuits offer a considerable improvement over their half-wave counterparts. They are not only more efficient but are significantly less demanding in terms of the reservoir and smoothing components. There are two basic forms of full- wave rectifier; the bi-phase type and the bridge rectifier type.

1.5 Bi-phase rectifier circuits

Fig. 1.9 shows a simple bi-phase rectifier circuit. Mains voltage (240 V) is applied to the primary of the step-down transformer (T1) which has two identical secondary windings, each providing 12 V r.m.s. (the turns ratio of T1 will thus be 240/12 or 20:1 for *each* secondary winding).

On positive half-cycles, point A will be positive with respect to point B. Similarly, point B will be positive with respect to point C. In this condition D1 will allow conduction (its anode will be positive with respect to its cathode) while D2 will not allow conduction (its anode will be negative with respect to its cathode). Thus D1 alone conducts on positive half-cycles.

On negative half-cycles, point C will be positive with respect to point B. Similarly, point B will be positive with respect to point A. In this condition D2 will allow conduction (its anode will be positive with respect to its cathode) while D1 will not allow conduction (its anode will be negative with respect to its cathode). Thus D2 alone conducts on negative half-cycles.

Fig. 1.10 shows the bi-phase rectifier circuit with the diodes replaced by switches. In Fig. 1.10(a) D1 is shown conducting on a positive half-cycle while in Fig. 1.10(b) D2 is shown conducting. The result is that current is routed through the load *in the same direction* on successive half-cycles.

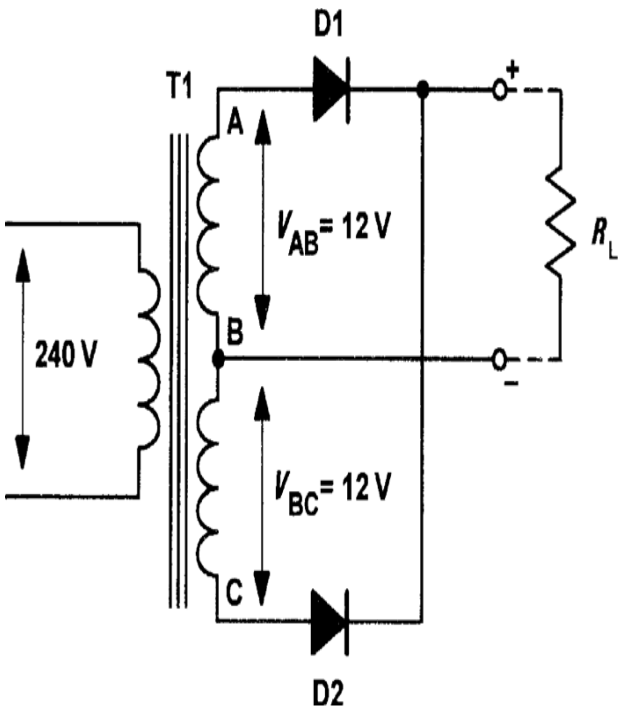


Fig. 1.9 Bi-phase rectifier circuit

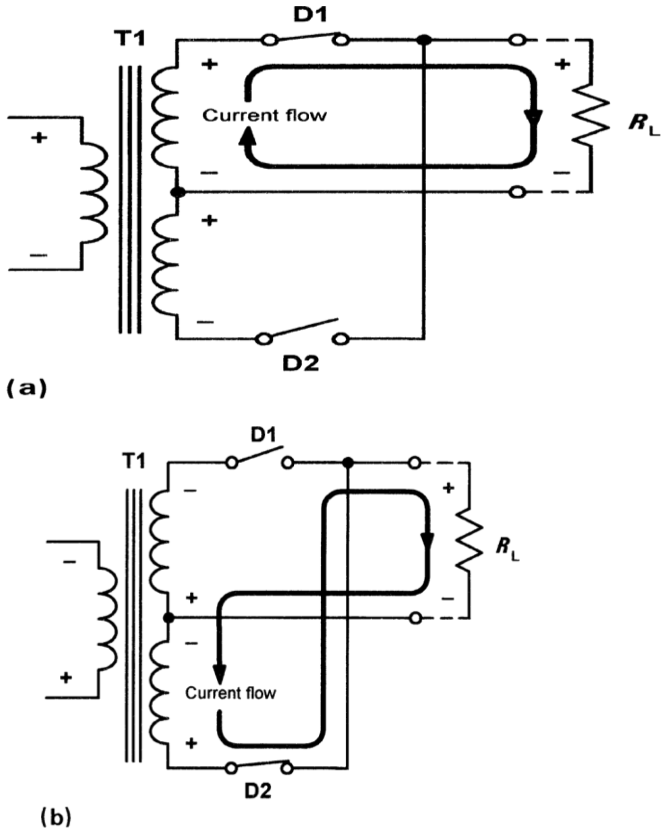


Fig. 1.10 (a) Bi-phase rectifier with D1 conducting and D2 non-conducting

(b) bi-phase rectifier with D2 conducting and D1 non-conducting

Furthermore, this current is derived alternately from the two secondary windings.

As with the half-wave rectifier, the switching action of the two diodes results in a pulsating output voltage being developed across the load resistor (R_L). However, unlike the half-wave circuit the pulses of voltage developed across R_L will occur at a frequency of 100 Hz (*not* 50 Hz). This doubling of the ripple frequency allows us to use smaller values of reservoir and smoothing capacitor to obtain the same degree of ripple reduction (the reactance of a capacitor is reduced as frequency increases).

As before, the peak voltage produced by each of the secondary windings will be approximately 17 V and the peak voltage across R_L will be 16.3 V i.e. 17 V less the 0.7 V forward threshold voltage dropped by the diodes).

Fig. 1.11 shows how a reservoir capacitor (*C*1) can be added to ensure that the output voltage remains at, or near, the peak voltage even when the diodes are not conducting. This component operates in exactly the same way as for the half-wave circuit, i.e. it charges to approximately 16.3 V at the peak of the positive half-cycle and holds the voltage at this level when the diodes are in their non-conducting states. The time required for C1 to charge to the maximum (peak) level is determined by the charging circuit time constant (the series resistance multiplied by the capacitance value). In this circuit, the series resistance comprises the secondary winding resistance together with the forward resistance of the diode and the (minimal) resistance of the wiring and connections. Hence *C*1 charges very rapidly as soon as either D1 or D2 starts to conduct. The time required for *C*1 to discharge is, in contrast, very much greater. The discharge time contrast is determined by the capacitance value and load resistance. In practice R is very much larger than the resistance of the secondary circuit and hence *C*1 takes an appreciable time to discharge. During this time, D1 and D2 will be reverse biased and held in a non-conducting state. As a consequence, the only discharge path for *C*1

is through R_L. Fig. 1.12 shows voltage waveforms for the bi-phase rectifier, with and without *C*1 present. Note that the ripple frequency (100 Hz) is twice that of the half-wave circuit shown previously in Fig. 1.7.

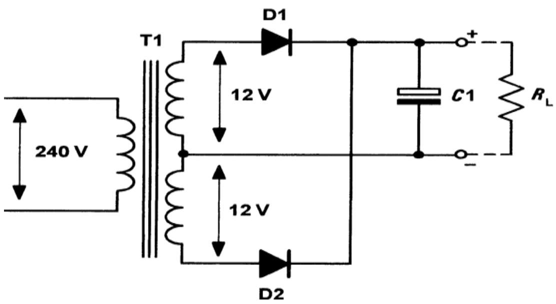


Fig. 1.11 Bi-phase rectifier with reservoir capacitor

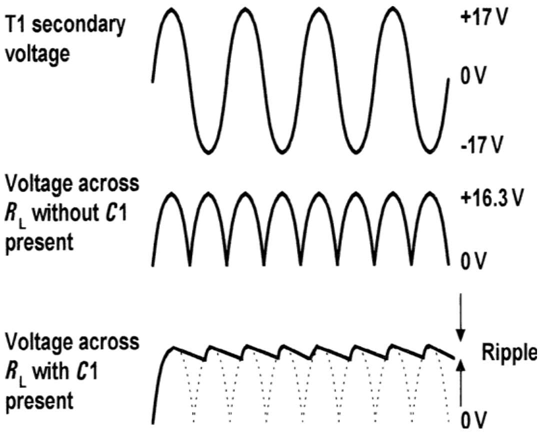


Fig. 1.12 Waveforms for the bi-phase rectifier

1.6 Bridge rectifier circuits

An alternative to the use of the bi-phase circuit is that of using a four-diode bridge rectifier (see Fig. 1.13) in which opposite pairs of diodes conduct on alternate half-cycles. This arrangement avoids the need to have two separate secondary windings. A full-wave bridge rectifier arrangement is shown in Fig. 1.14. Mains voltage (240 V) is applied to the primary of a step-down transformer (T1).

The secondary winding provide 12 v rms (approximately 17 V peak) and has a turns ratio of 20:1, as before. On positive half-cycles, point A will be positive with respect to point B. In this condition D1 and D2 will allow conduction while D3 and D4 will not allow conduction. Conversely, on negative half-cycles, point B will be positive with respect to point A. In this condition D3 and D4 will allow conduction while D1 and D2 will not allow conduction.

Fig. 1.15 shows the bridge rectifier circuit with the diodes replaced by four switches. In Fig. 1.15(a) D1 and D2 are conducting on a positive half-cycle while in Fig. 1.15(b) D3 and D4 are conducting. Once again, the result is that current is routed through the load *in the same direction* on successive half-cycles. As with the bi-phase rectifier, the switching action of the two diodes results in a pulsating output voltage being developed across the load resistor (R_L).

Once again, the peak output voltage is approximately 16.3 V (i.e. 17 V less the 0.7 V forward threshold voltage).

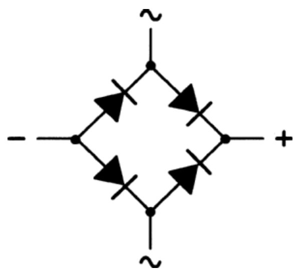


Fig. 1.13 Four diodes connected as a bridge

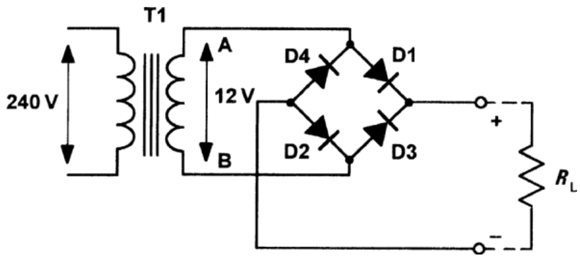


Fig. 1.14 Full-wave bridge rectifier circuit

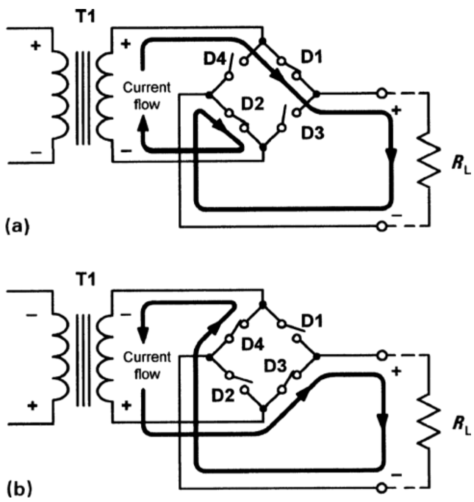


Fig. 1.15 (a) Bridge rectifier with D1 and D2 conducting, D3 and D4 non-conducting (b) bridge rectifier with D1 and D2 non-conducting, D3 and D4 conducting

Fig. 1.16 shows how a reservoir capacitor (*C*1) can be added to maintain the output voltage when the diodes are not conducting. This component operates in exactly the same way as for the bi-phase circuit, i.e. it charges to approximately 16.3 V at the peak of the positive half-cycle and holds the voltage at this level when the diodes are in their non-conducting states. This component operates in exactly the same way as for the bi-phase circuit and the secondary and rectified output waveforms are shown in Fig. 1.17.

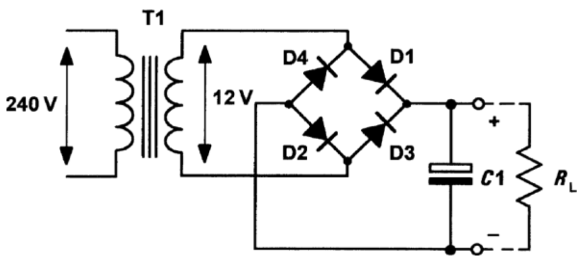


Fig. 1.16 Bridge rectifier with reservoir capacitor

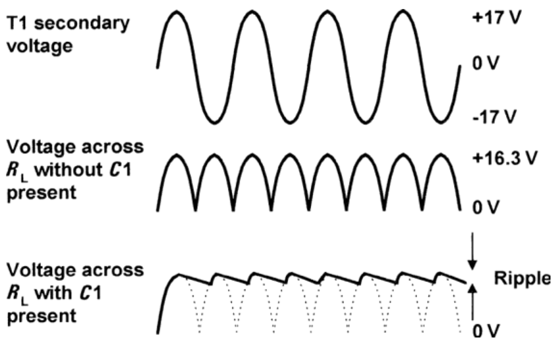


Fig. 1.17 Waveforms for the bridge rectifier

Once again note that the ripple frequency is twice that of the incoming a.c. supply.

Finally, *R*–*C* and *L*–*C* ripple filters can be added to bi-phase and bridge rectifier circuits in exactly the same way as those shown for the half-wave rectifier arrangement (see Figs 1.8 and 1.9)

1.7 Output resistance and voltage regulation

In a perfect power supply, the output voltage would remain constant regardless of the current taken by the load. In practice, however, the output voltage falls as the load current increases. To account for this fact, we say that the power supply has internal resistance (ideally this should be zero). This internal resistance appears at the output of the supply and is defined as the change in output voltage divided by the corresponding change in output current. Hence: R_{out}= change in output voltage/ change in output current is

$$R_{out} = DV_{out} / DI_{out} \qquad (3)$$

in a simple regulator circuit to supply a regulated 5 V to a load having a resistance of 400 Ω, determine a suitable value of series resistor for operation in conjunction with a supply of 9 V.

The regulation of a power supply is given by the relationship:

Regulation = change in output voltage / change in line (input) voltage x 100%

Ideally, the value of regulation should be very small. It is capable of producing values of regulation of 5% to 10%. More sophisticated circuits based on discrete components produce values of between 1% and 5% and integrated circuit regulators often provide values of 1% or less

1.8 Voltage multipliers

By adding a second diode and capacitor, we can increase the output of the simple half-wave rectifier that we met earlier. A voltage doubler using this technique is shown in Fig. 1.18. In this arrangement C1 will charge to the positive peak secondary voltage while C2 will charge to the negative peak secondary voltage. Since the output is taken from C1 and C2 connected in series the resulting output voltage is twice that produced by one diode alone.

The voltage doubler can be extended to produce higher voltages using the cascade

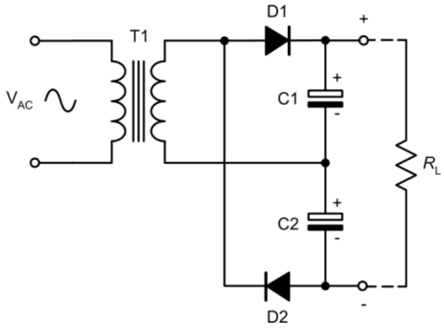


Fig. 1.18 A voltage doubler

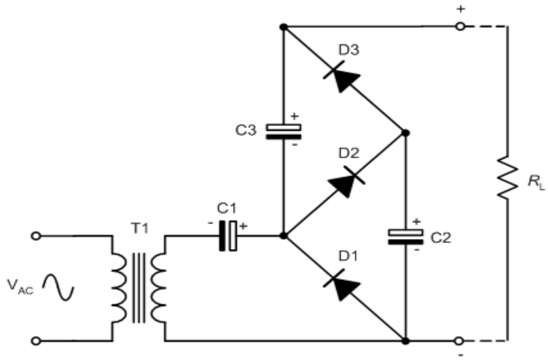


Fig. 1.19 A voltage tripler

arrangement shown in Fig. 1.19. Here C1 charges to the positive peak secondary voltage, while C2 and C3 charge to twice the positive peak secondary voltage. The result is that the output voltage is the sum of the voltages across C1 and C3 which is three times the voltage that would be produced by a single diode. The ladder arrangement shown in Fig. 1.18 can be easily extended to provide even higher voltages but the efficiency of the circuit becomes increasingly impaired and high-order voltage multipliers of this type are only suitable for providing relatively small currents.

Symbols introduced in this chapter

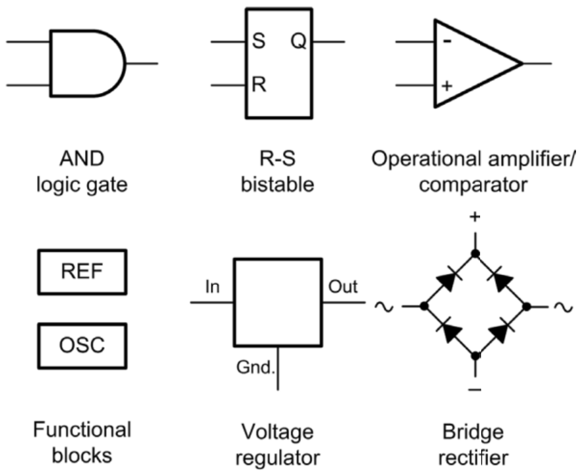


Fig. 1.20 Circuit symbols introduced in this chapter

Problems

1. A half-wave rectifier is fitted with an *R*–*C* smoothing filter comprising $R = 200\ \Omega$ and $C = 50\ \mu F$. If 2 V of 400 Hz ripple appear at the input of the circuit, determine the amount of ripple appearing at the output

2. The *L*–*C* smoothing filter fitted to a 50 Hz mains operated full-wave rectifier circuit consists of $L = 4$ H and $C = 500\ \mu F$. If 4 V of ripple appear at the input of the circuit, determine the amount of ripple appearing at the output.

3. If a 9 V zener diode is to be used in a simple shunt regulator circuit to supply a load having a nominal resistance of 300 Ω, determine the maximum value of series resistor for operation in conjunction with a supply of 15 V.

4. The circuit of a d.c. power supply is shown in Fig. 1.21. Determine the voltages that will appear at test points A, B and C.

5. In Fig. 1.21, determine the current flowing in *R*1 and the power dissipated in D5 when the circuit is operated without any load connected.

6. In Fig. 1.21, determine the effect of each of the following fault conditions:

   (a) *R*1 open-circuit;

(b) D5 open-circuit;

(c) D5 short-circuit.

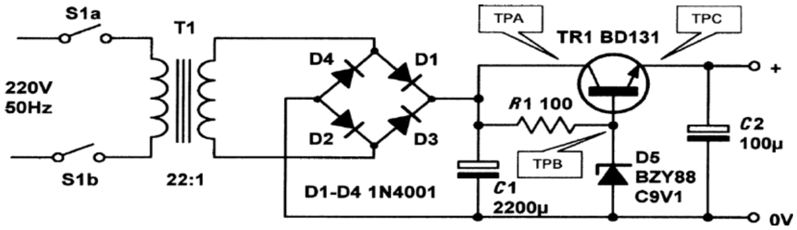


Fig. 1.21

7. A 220 V a.c supply feeds a 20:1 step- down transformer, the secondary of which is connected to a bridge rectifier and reservoir capacitor. Determine the approximate d.c. voltage that will appear

across the reservoir capacitor under 'no- load' conditions.

8. The following data were obtained during a load test carried out on a d.c. power supply:

Output voltage (no-load) = 8.5 V Output voltage (800 mA load) = 8.1 V

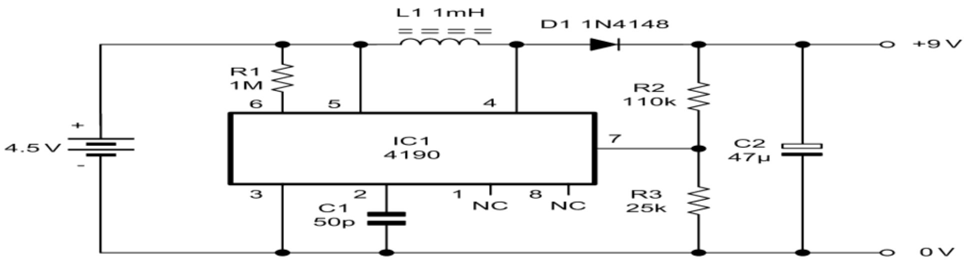
Determine the output resistance of the power supply and estimate the output voltage at a load current of 400 mA.

9. The following data were obtained during a regulation test on a d.c. power supply:

Output voltage (a.c. input: 230 V) = 15 V Output voltage (a.c. input: 190 V) = 14.6 V

Determine the regulation of the power supply and estimate the output voltage when the input voltage is 245 V.

10. Fig. 1.22 shows a switching regulator circuit that produces an output of 9 V for an input of 4.5 V. What type of regulator is this? Between which pins of IC1 is the switching transistor connected? Which pin on IC1 is used to feed back a proportion of the output voltage to the internal comparator stage?



AMPLIFIER

1.9 Types of amplifier

a.c. coupled amplifiers

In a.c. coupled amplifiers, stages are coupled together in such a way that d.c. levels are isolated and only the a.c. components of a signal are transferred from stage to stage.

d.c. coupled amplifiers

In d.c. (or direct) coupled amplifiers, stages are coupled together in such a way that stages are not isolated to d.c. potentials. Both a.c. and d.c. signal components are transferred from stage to stage.

Large-signal amplifiers

Large-signal amplifiers are designed to cater for appreciable voltage and/or current levels (typically from 1 V to 100 V or more).

Small-signal amplifiers

Small-signal amplifiers are designed to cater for low-level signals (normally less than 1 V and often much smaller). Small-signal amplifiers have to be specially designed to combat the effects of noise.

Audio frequency amplifiers

Audio frequency amplifiers operate in the band of frequencies that is normally associated with audio signals (e.g. 20 Hz to 20 kHz).

Wideband amplifiers

Wideband amplifiers are capable of amplifying a very wide range of frequencies, typically from a few tens of hertz to several megahertz.

Radio frequency amplifiers

Radio frequency amplifiers operate in the band of frequencies that is normally associated with radio

Low-noise amplifiers

Low-noise amplifiers are designed so that they contribute negligible noise (signal disturbance) to the signal being amplified. These amplifiers are usually designed for use with very small signal levels (usually less than 10 mV or so).

Gain

One of the most important parameters of an amplifier is the amount of amplification or gain that it provides. Gain is simply the ratio of output voltage to input voltage, output current to input current, or output power to input power (see Fig. 1.23). These three ratios give, respectively, the voltage gain, current gain and power gain.

Thus:

Voltage gain, $A_V = V_{out} / V_{in}$ (4)

Current gain, $A_i = I_{out} / I_{in}$ (5)

Power gain, $A_P = P_{out} / P_{in}$

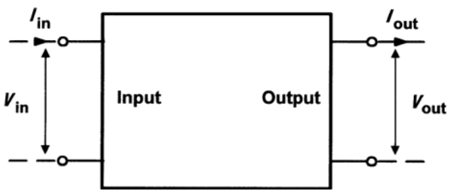


Fig. 1.23 Block diagram for an amplifier showing input and output voltages and currents

Note that, since power is the product of current and voltage ($P = I\,V$), we can infer that:

$A_P = P_{out}/P_{in} = I_{out} / I_{in} \times V_{out} / V_{in}$ (6)

Example 1.3

An amplifier produces an output voltage of 2 V for an input of 50 mV. If the input and output currents in this condition are, respectively, 4 mA and
200 mA, determine:

(a) the voltage gain;

(b) the current gain;

(c) the power gain.

Solution

(a) The voltage gain is calculated from:

Voltage gain, $A_V = V_{out}/V_{in} = 2v/50mv = 40$

(b) Current gain, $A_i = I_{out}/I_{in} = 200mA/4Ma = 50$

(c) Power gain, $A_P = P_{out}/P_{in} = 2$

1.10 Input and output resistance

Input resistance is the ratio of input voltage to input current and it is expressed in ohms. The input of an amplifier is normally purely resistive (i.e. any reactive component is negligible) in the middle of its working frequency range (i.e. the **mid-band**). In some cases, the reactance of the input may become appreciable (e.g. if a large value of stray capacitance appears in parallel with the input resistance). In such cases we would refer to **input impedance** rather than input resistance.

Output resistance is the ratio of open-circuit output voltage to short-circuit output current and is measured in ohms. Note that this resistance is internal to the amplifier and should not be confused with the resistance of a load connected externally.

As with input resistance, the output of an amplifier is normally purely resistive and we can safely ignore any reactive component. If this is not the case, we would once again need to refer to **output impedance** rather than output resistance.

Fig. 1.24 shows how the input and output resistances are 'seen' looking into the input and output terminals, respectively. We shall be returning to this equivalent circuit a little later in this chapter. Finally, it's important to note that, although these resistances are meaningful in terms of the signals present, they cannot be measured using a conventional meter!

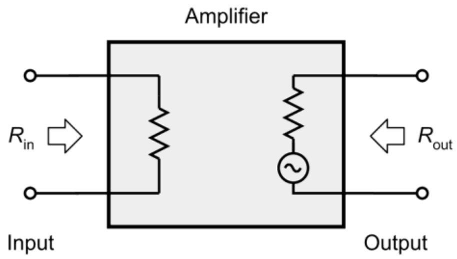


Fig. 1.24 Input and output resistances 'seen' looking into the input and output terminals, respectively

1.11 Frequency response

The frequency response characteristics for various types of amplifier are shown in Fig. 1.25. Note that, for response curves of this type, frequency is almost invariably plotted on a **logarithmic scale**.

The frequency response of an amplifier is usually specified in terms of the upper and lower **cut-off frequencies** of the amplifier. These frequencies are those at which the output power has dropped to 50% (otherwise known as the **-3 dB points**) or where the voltage gain has dropped to 70.7% of its mid-band value.

Figs 1.26 and 1.27, respectively, show how the bandwidth can be expressed in terms of either power or voltage (the cut-off frequencies, f_1 and f_2, and bandwidth are identical).

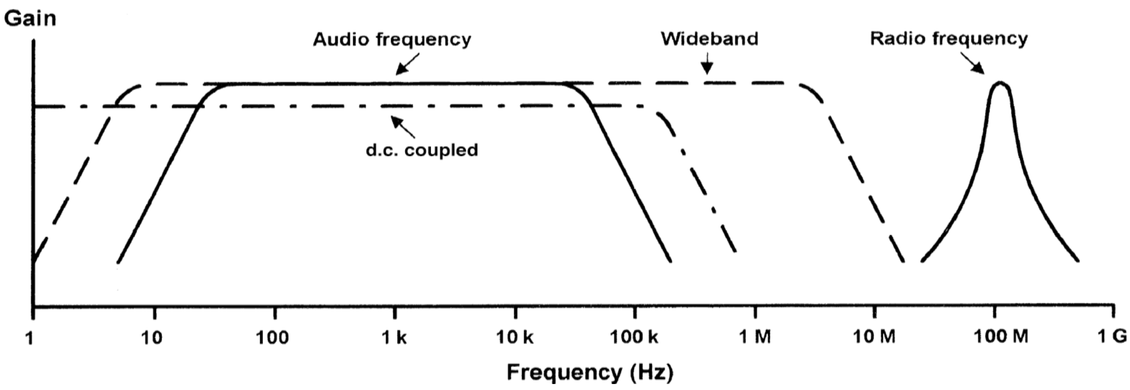


Fig. 1.26 Frequency response and bandwidth (output power plotted against frequency)

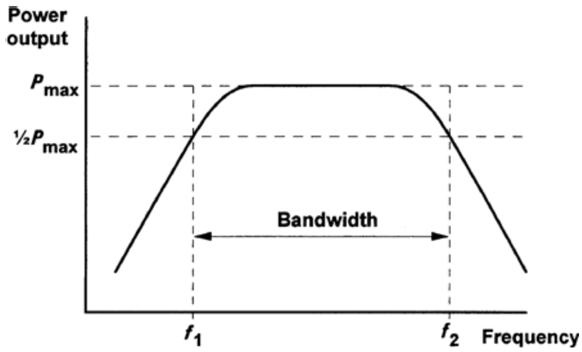


Fig. 1.27 Frequency response and bandwidth (output power plotted against frequency)

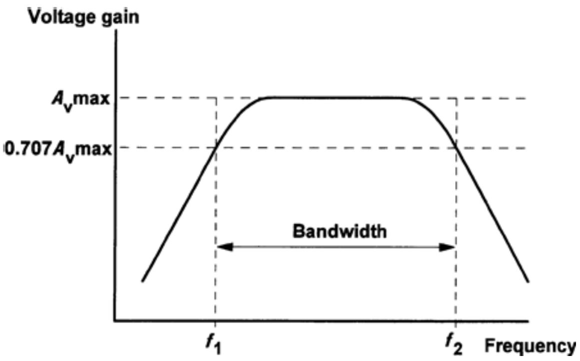


Fig. 1.28 Frequency response and bandwidth (output voltage plotted against frequency)

Example 1.4

Determine the mid-band voltage gain and upper and lower cut-off frequencies for the amplifier whose frequency response is shown

Solution

The mid-band voltage gain corresponds with the flat part of the frequency response characteristic. At that point the voltage gain reaches a maximum of 35 (see Fig. 1.29).

The voltage gain at the two cut-off frequencies can be calculated from:

A_V cut-off = 0.707 X A_V max = 0.707 X 35 = 24.7, this value of gain intercepts the frequency response graph at f_1 = 57 Hz and f_2 = 590 kHz (see Fig. 1.29).

1.12 Bandwidth

The bandwidth of an amplifier is usually taken as the difference between the upper and lower cut-off frequencies (i.e. $f_2 - f_1$ in Figs 1.27 and 1.28). The bandwidth of an amplifier must be sufficient to accommodate the range of frequencies present within the signals that it is to be presented with. Many signals contain **harmonic** components (i.e. signals at 2*f*, 3*f*, 4*f*, etc. where *f* is the frequency of the **fundamental** signal). To reproduce a square wave, for example, requires an amplifier with a very wide bandwidth (note that a square wave comprises an infinite series of harmonics). Clearly it is not possible to *perfectly* reproduce such a wave, but it does explain why it can be desirable for an amplifier's bandwidth to greatly exceed the highest signal frequency that it is required to handle

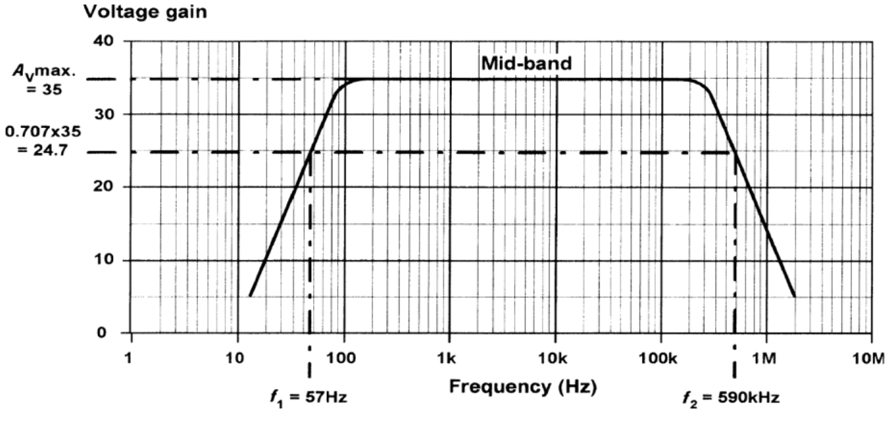


Fig. 1.29 An Example

1.13 Phase shift

Phase shift is the phase angle between the input and output signal voltages measured in degrees. The measurement is usually carried out in the mid-band where, for most amplifiers, the phase shift remains relatively constant. Note also that conventional single-stage transistor amplifiers provide phase shifts of either 180° or 360°.

1.14 Negative feedback

Many practical amplifiers use negative feedback in order to precisely control the gain, reduce Distortion and improve bandwidth. The gain can be reduced to a manageable value by feeding back a small proportion of the output. The amount of feedback determines the overall (or **closed-loop**) gain. Because this form of feedback has the effect of reducing the overall gain of the circuit, this form of feedback is known as **negative feedback**. 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**. This form of feedback is used in oscillator circuits

Fig. 1.30 shows the block diagram of an amplifier stage with negative feedback applied. In this circuit, the proportion of the output voltage fed back to the input is given by the overall voltage gain will be given by:

Overall gain $= V_{out} / V_{in}$

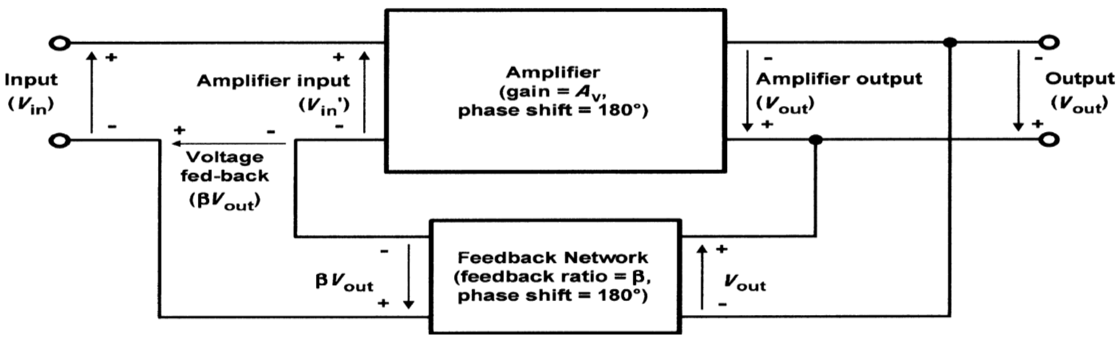


Fig. 1.30 Amplifier with negative feedback applied

Now $V_{in}' = V_{in} - \beta V_{out}$ (by applying Kirchhoff's Voltage Law) (note that the amplifier's input Voltage has been *reduced* by applying negative feedback) thus

$$V_{in} = V_{in}' + \beta V_{out} \qquad (7)$$

And

$V_{out} = A_V \times V_{in}$ (note that A_V is the **internal gain** of the amplifier)

Hence:

$$\text{Overall gain, } G = A_v V_{in}' / (V_{in}' + \beta V_{out}) \qquad (8)$$

$$G = A_v V_{in}' / ((V_{in}' + \beta (A_V \times V_{in})) \qquad (9)$$

$$G = A_v / (1 + \beta A_v) \qquad (10)$$

Hence, the overall gain with negative feedback applied will be less than the gain without feedback. Furthermore, if A_V is very large. The overall gain with negative feedback applied will be given by:

$$G = 1/\beta \text{ (when } A_V \text{ is very large)} \qquad (11)$$

Note, also, that the **loop gain** of a feedback

Amplifier is defined as the product of β and A_V.

Example 1.5

An amplifier with negative feedback applied has an open-loop voltage gain of 50, and one-tenth of its output is fed back to the input (i.e. β= 0.1). Determine the overall voltage gain with negative feedback applied.

Solution

With negative feedback applied the overall voltage gain will be given by

$G = A_v / (1 + \beta A_v)$

50/6=8.33

1.15 Multi-stage amplifiers

In order to provide sufficiently large values of gain, it is frequently necessary to use a number of interconnected stages within an amplifier. The overall gain of an amplifier with several stages (i.e. a multi-stage amplifier) is simply the product of the individual voltage gains. Hence:

$$A_V = A_{V1} \times A_{V2} \times A_{V3}, \text{ etc.} \qquad (12)$$

Note, however, that the bandwidth of a multi- stage amplifier will be less than the bandwidth of each individual stage. In other words, an increase in gain can only be achieved at the expense of a reduction in bandwidth.

Signals can be coupled between the individual stages of a multi-stage amplifier using one of a number of different methods shown in Fig. 1.31

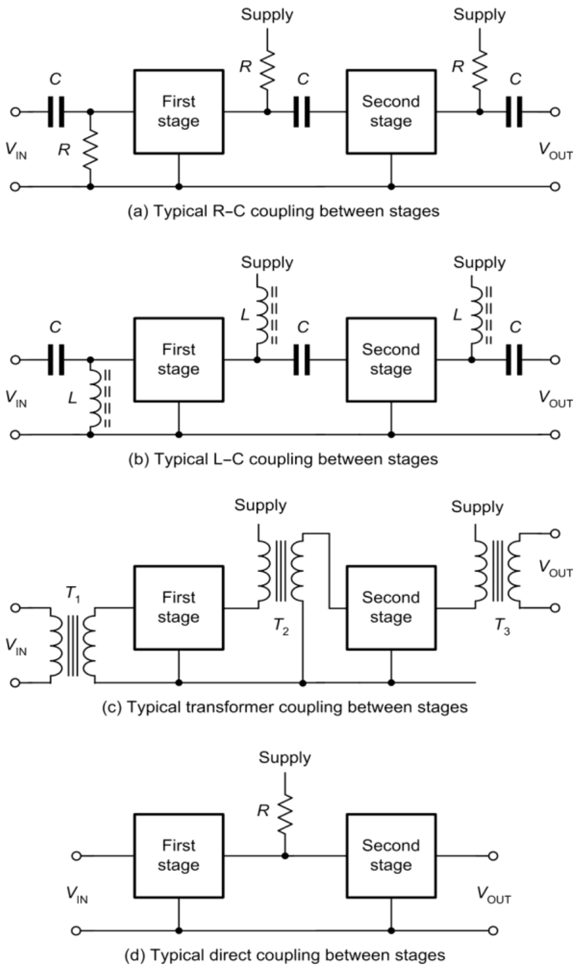


Fig. 1.31 Different methods used for inter stage coupling

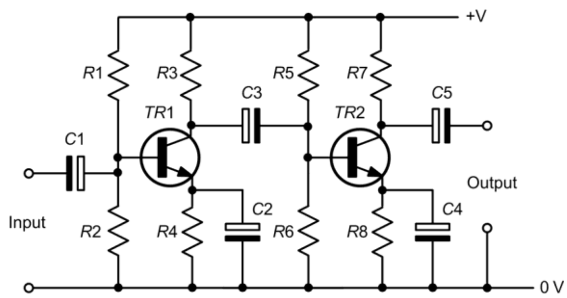


Fig. 1.32 A typical two-stage high-gain *R–C* coupled common-emitter amplifier

The most commonly used method is that of ***R–C* coupling** as shown in In Fig. 1.31(a). In this coupling method, the stages are coupled together using capacitors having a low reactance at the signal frequency and resistors (which also provide a means of connecting the supply). Fig. 1.32 shows a practical example of this coupling method.

A similar coupling method, known as ***L–C*** **coupling**, is shown in Fig. 1.31 (b). In this method, the inductors have a high reactance at the signal frequency. This type of coupling is generally only used in RF and high-frequency amplifiers.

Two further methods, **transformer coupling** and **direct coupling,** are shown in Figs 1.31 (c) and 1.31 (d), respectively. The latter method is used where d.c. levels present on signals must be preserved.

OPERATIONAL AMPLIFIER

1.16 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.

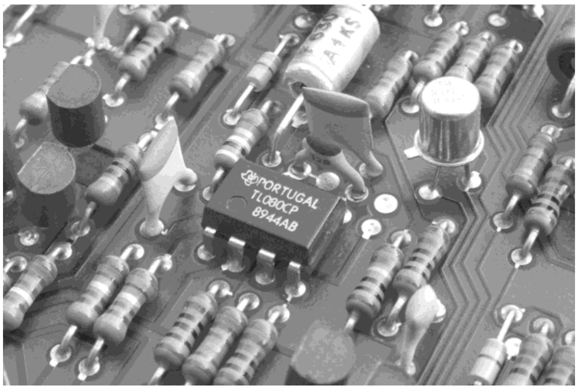


Fig. 1.33 A typical operational amplifier. This device is supplied in an eight-pin dual-in-line (DIL) package. It has a JFET input stage and produces a typical open-loop voltage gain of 200,000

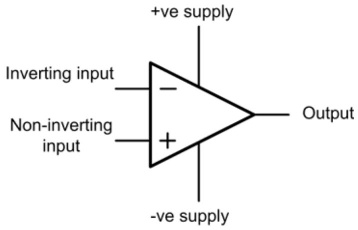


Fig. 1.34 Symbol for an operational amplifier

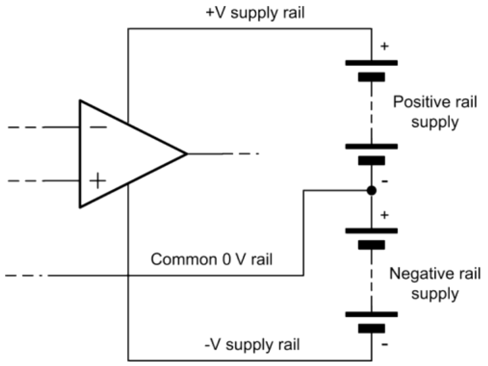


Fig. 1.35 Supply connections for an operational amplifier

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

$$A_{V(OL)} = V_{out} / V_{in} \quad (13)$$

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:

$$A_{V(OL)} = 20 \log_{10} V_{out} / V_{in} \quad (14)$$

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:

$$A_{V(CL)} = V_{out} / V_{in} \tag{15}$$

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 1.6

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.

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

$$A_{V(CL)} = 20 \log_{10} (5,000) = 20 \text{ X } 3.7 = 74 \text{ dB}$$

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

$$R_{IN} = V_{IN} / I_{IN} \tag{16}$$

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.

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:

$$R_{out} = V_{OUT(OC)} / I_{OUT(SC)} \qquad (17)$$

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. 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:

$$\text{Slew rate} = \Delta V_{OUT} / \Delta t \qquad (18)$$

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.

1.17 Operational amplifier characteristics

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

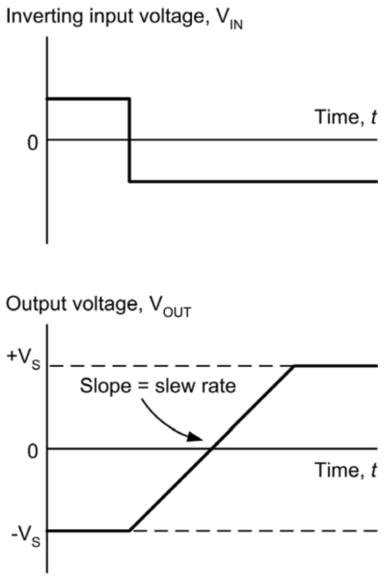


Fig. 1.36 Slew rate for an operational amplifier

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 1.1.

Table 1.1 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 1.7

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:

Slew rate = $\Delta V_{OUT} / \Delta t$

= 10V/4 µs = 2.5 V / µs

1.18 Operational amplifier configurations

The three basic configurations for operational voltage amplifiers, together with the expressions for their voltage gain, are shown in Fig. 1.37. 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, RF
- The value of the input and feedback capacitors, CIN and CF respectively, are chosen so as to rolloff 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
- 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:

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

  $= 0.159 / C_{IN}R_{IN}$

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

$$f_2 = 1 / 2\pi C_F R_F \qquad (19)$$

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

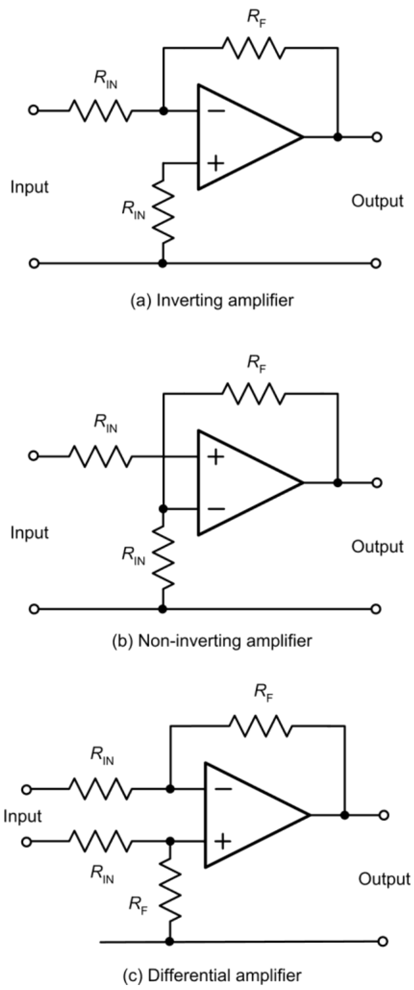


Fig. 1.37 The three basic configurations for operational voltage amplifiers

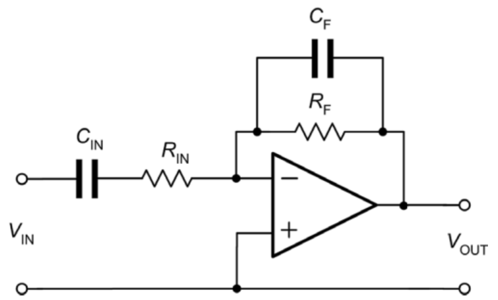


Fig. 1.38 Adding capacitors to modify the frequency response of an inverting operational amplifier

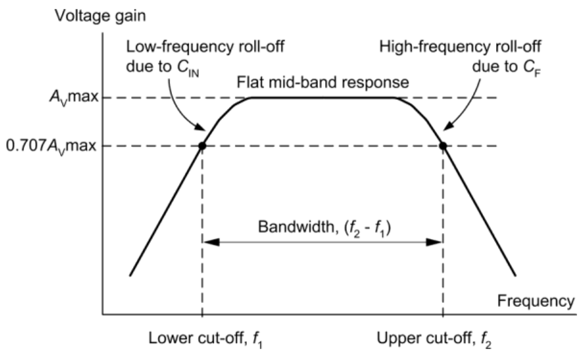


Fig. 1.39 Effect of adding capacitors, C_{IN} and C_F, to modify the frequency response of an operational amplifier

1.19 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.

Voltage followers

A voltage follower using an operational amplifier is shown in Fig. 1.40 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. 1.41 Notice how the input and output waveforms are both in-phase (they rise and fall together) and that they are identical in amplitude.

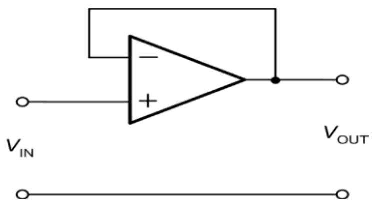


Fig. 1.40 A voltage follower

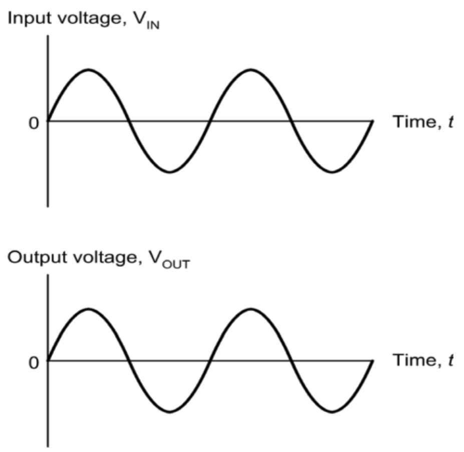


Fig. 1.41 Typical input and output waveforms for a voltage follower

Differentiators

A differentiator using an operational amplifier is shown in Fig. 1.42. 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. 1.43. 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.

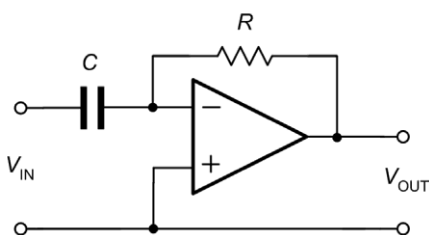


Fig. 1.42 A differentiator

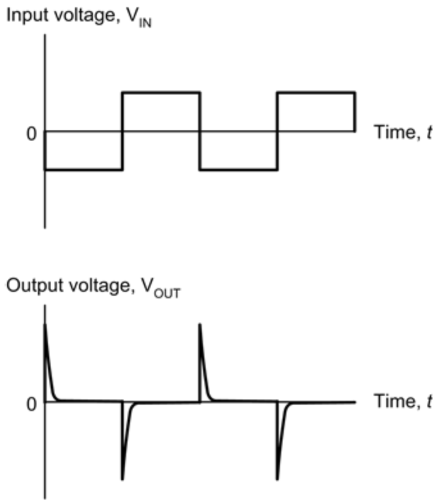


Fig. 1.43 Typical input and output waveforms for a differentiator

Integrators

An integrator using an operational amplifier is shown in Fig. 1.44. This circuit provides the

Opposite function to that of a differentiator 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. 1.45. 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.

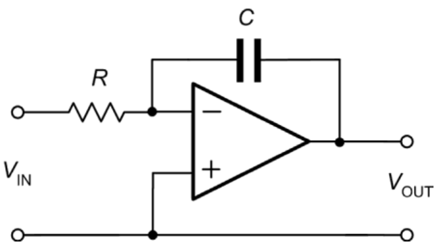


Fig. 1.44 An integrator

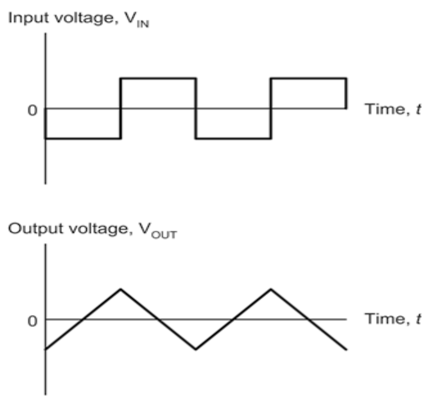


Fig. 1.45 Typical input and output waveforms for an integrator

Comparators

A comparator using an operational amplifier is shown in Fig. 1.46. 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 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. 1.47. 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.

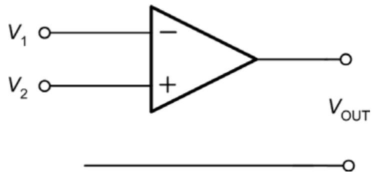


Fig. 1.46 A comparator

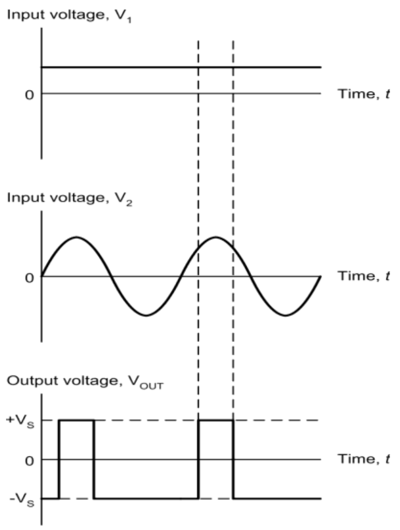


Fig. 1.47 Typical input and output waveforms for a comparator

Summing amplifiers

A summing amplifier using an operational amplifier is shown in Fig. 1.48. 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:

$$V\mathrm{OUT} = -(V1 + V2) \qquad (20)$$

where V_1 and V_2 are the input voltages (note that all of the resistors used in the circuit have the same value).

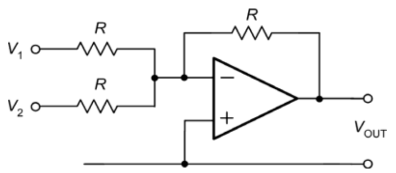


Fig. 1.48 Typical input and output waveforms for a summing amplifier

OSCILLATOR

1.20 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 positive so that the output signal arrives back at the input in such a sense as to reinforce the original signal.

Ladder network oscillator

A simple phase-shift oscillator based on a three- stage *C–R* ladder network is shown in Fig. 1.49. 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. 1.49 is given by:

$$f=1/2\pi\sqrt{6}\,xCR \qquad (21)$$

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.

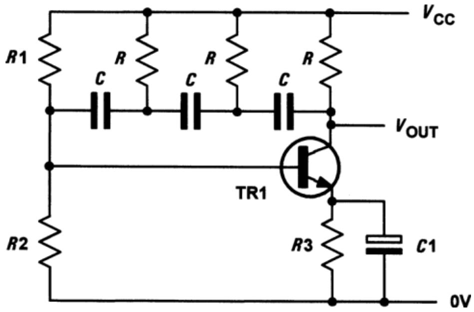


Fig. 1.49 Sine wave oscillator based on a three- stage *C–R* ladder network

Wien bridge oscillator

An alternative approach to providing the phase shift required is the use of a Wien bridge network. 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:

$$f_r = 1/2\pi\sqrt{(R_1C_1R_2C_2)}$$

$$\text{if } R_1 = R_2 = R \text{ and } C_1 = C_2 = C$$

Then,

$$f_r = 1/2\pi RC \qquad (22)$$

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

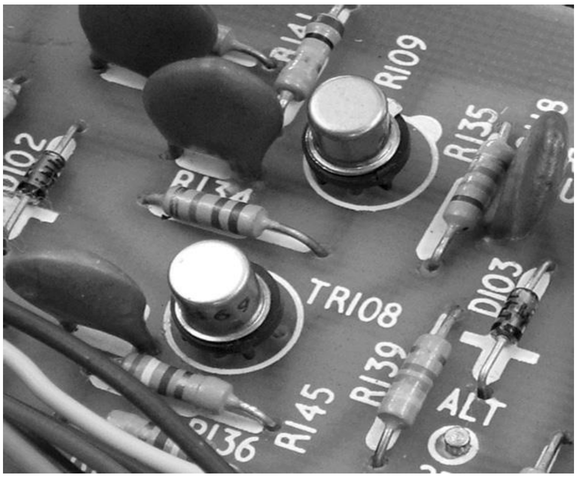


Fig. 1.50 This high-speed bistable multivibrator uses two general-purpose silicon transistors and works at frequencies of up to 1 MHz triggered from an external signal

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. 1.51. The circuit employs positive feedback with the output fed back to the non-inverting input via the potential divider formed by R1 and R2. 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. 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 +VCC 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 −VCC. 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 +VCC 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

$V_{UT} = V_{CC} (R_2 / (R_1+R_2)$

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

$V_{LT} = V_{CC} (R_2 / (R_1+R_2)$

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

$T=2CR \ln(1+(R_2 +R_1)$ (23)

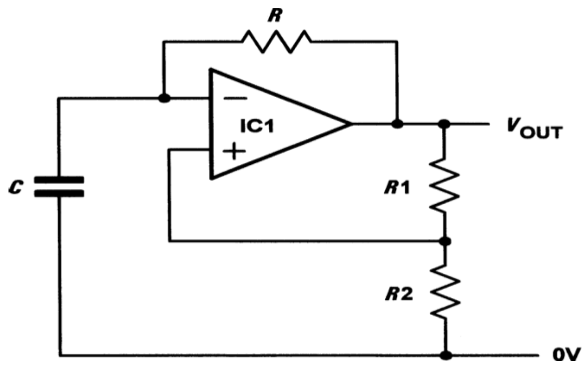


Fig. 1.51 Single-stage astable oscillator using an operational amplifier

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 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.