(ii) Step input: If input is a step voltage $v_i = 1$ V for $0 \le t \le 0.3$ ms, then the output voltage at $t = 0.3$ ms is

$$v_o = -\frac{1}{R_1 C_F}\int_0^{0.3\text{ ms}} 1 \, . \, dt = -\frac{1}{10\text{ k}\Omega \times 10\text{ nF}} \times t\Big|_{t=0}^{t=0.3\text{ ms}}$$

$$= -10^4 \times 0.3 \times 10^{-3} = -3\text{ V}$$

The output voltage is a ramp function with a slope of 10 V/ms and is shown in Fig. 4.25 (b).

(iii) Square wave input: The output waveform for an input of 5 KHz, 1 V peak square wave is shown in Fig. 4.25 (c).

It can be seen that input is of constant amplitude of 1 V from 0 to 0.1 ms and – 1 V from 0.1 ms to 0.2 ms. The output for each of these half periods will be ramps as seen above for step inputs. Thus, the expected output wave form will be a triangular wave. The peak value of the output for first half cycle is

$$v_o = -\frac{1}{R_1 C_F}\int_0^{0.1\text{ ms}} 1 . \, dt = -10^4 \times 0.1 \times 10^{-3} = -1\text{V}$$

This represents the total change in the output voltage over the first half cycle from 0 to 0.1 ms. Similarly, integration over the next half cycle produces a positive change of 1V.

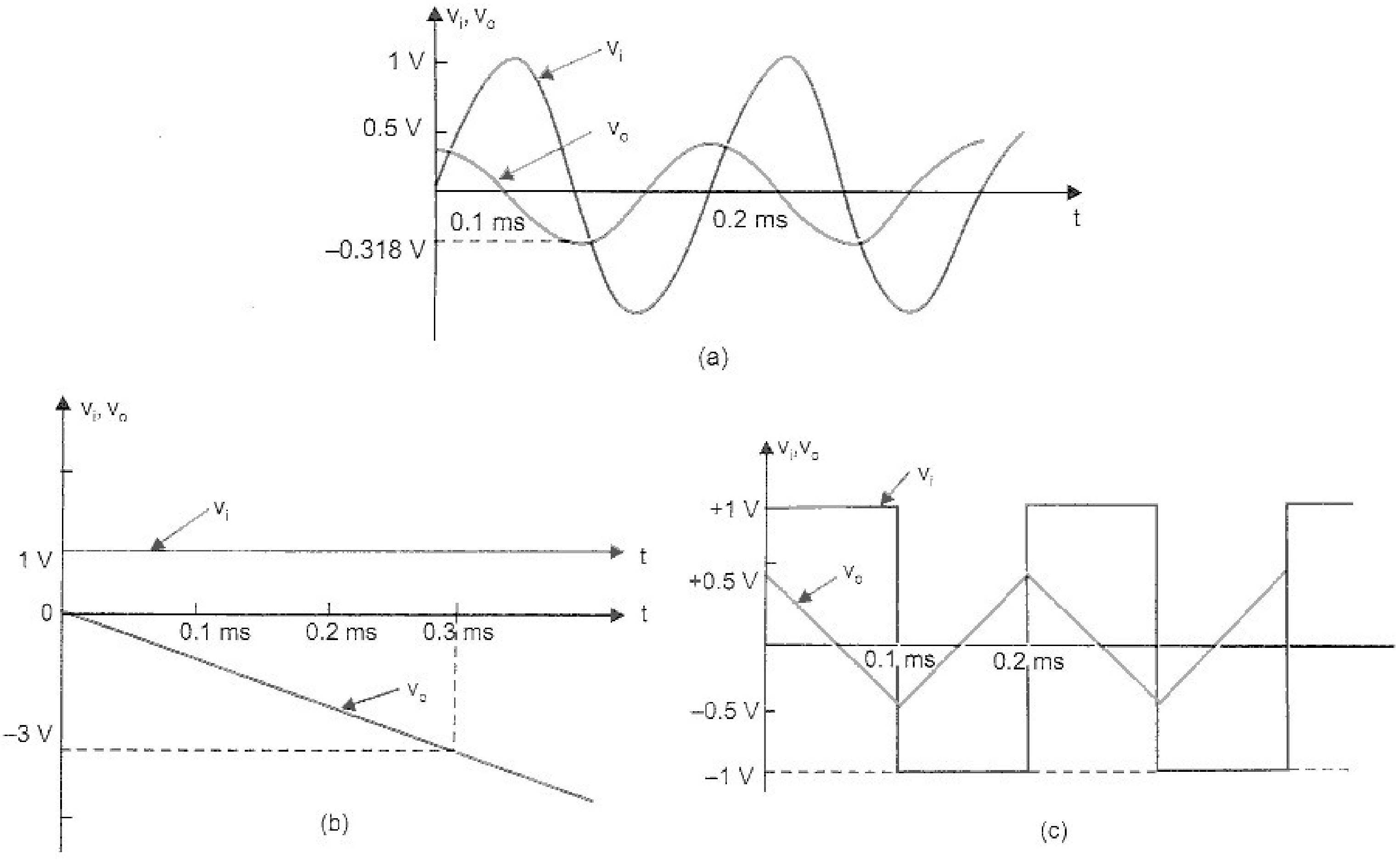


Fig. 4.25 Input and output waveforms for the integrator in Example 4.4
(a) Sine wave input, (b) Step input, (c) Square wave input

Example 4.5

Find R_1 and R_F in the lossy integrator so that the peak gain is 20 dB and the gain is 3 dB down from its peak when ω = 10,000 rad/s. Use a capacitance of 0.01 μF.

Solution

From Eq. (4.84), we see that gain is at its peak when ω = 0. The peak value in dB is therefore,

$$A\,(\text{dB}) = 20\ \log_{10} \frac{R_F/R_1}{\sqrt{1+0}} = 20 \qquad \text{(given)}$$

or,
$$\log_{10} \frac{R_F}{R_1} = 1$$

Thus we have,
$$\frac{R_F}{R_1} = 10$$

or,
$$R_1 = \frac{R_F}{10}$$

At ω = 10^4 rad/s, gain in dB is down by 3 dB from its peak of 20 dB, and thus is 17 dB. Therefore, converting gain to dB in Eq. (4.84) and substituting for ω, C and R_F/R, we have

$$20\ \log_{10} \frac{10}{\sqrt{1+[(10^4)10^{-8}\ R_F]^2}} = 17\ \text{dB}$$

or, $$20\ \log_{10} 10 - 20\ \log_{10} \sqrt{1+(10^{-4}\ R_F)^2} = 17\ \text{dB}$$

This simplifies to

$$20\ \log_{10} \left[1+(10^{-4}\ R_F)^2\right] = 3\ \text{dB}$$

or,
$$1 + (10^{-4}\ R_F)^2 = 10^{3/10} = 2$$

Thus we have
$$(10^{-4}\ R_F)^2 = 1$$

or,
$$R_F = 10^4\ \Omega = 10\ \text{k}\Omega$$

and
$$R_1 = 10\ \text{k}\Omega/10 = 1\ \text{k}\Omega$$

Example 4.6

Show that the output of an op-amp integrator to a step input of magnitude V volts is given by

$$v_o = A_v V(1 - e^{-t/R_1 C_F(1-A_v)})$$

Compare this result with the output obtained from a low pass RC circuit.

Solution

Figure 4.26 is a simple op-amp integrator where Miller's theorem is applied across the feedback capacitor C_F. The input time constant $T = R_1C_F\,(1 - A_v)$. Therefore,

$$v_i = V(1 - e^{-t/T}) \tag{4.86}$$

and
$$v_o = A_v\, v_i = A_v\, V(1 - e^{-t/R_1 C_F (1 - A_v)}) \tag{4.87}$$

or,
$$v_o = A_v\, V\left[1 - \left(1 - \frac{t}{R_1C_F\,(1 - A_v)} - \frac{t^2}{2\left(R_1C_F\,(1 - A_v)^2\right)} - \ldots\right)\right]$$

$$= \frac{A_v\, V t}{R_1\, C_F\,(1 - A_v)}\left[1 - \frac{t}{2R_1C_F\,(1 - A_v)} - \ldots.\right]$$

or,
$$v_o \approx -\frac{Vt}{R_1C_F}\left[1 - \frac{t}{2R_1C_F(1 - A_v)}\right];\ \text{if } A_v >> 1 \tag{4.88}$$

Also we know that for a simple low pass RC integrating network (without op-amp) the output v_o for a step input of V becomes

$$v_o = V(1 - e^{-t/RC}) \tag{4.89}$$

For large RC,
$$v_o \approx \frac{Vt}{RC}\left(1 - \frac{t}{2RC}\right) \tag{4.90}$$

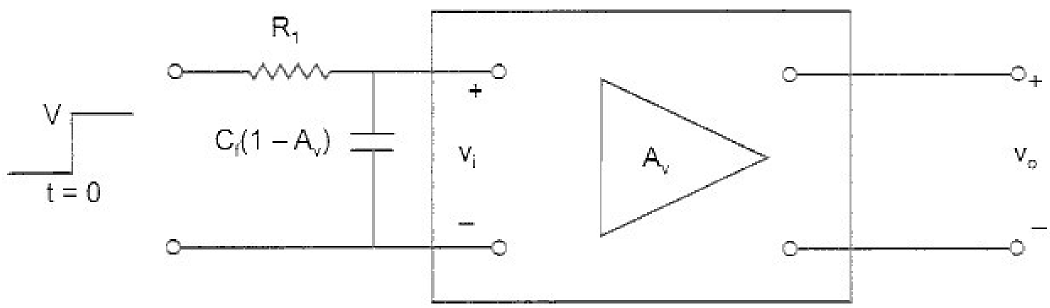


Fig. 4.26 Circuit for Example 4.6

It can be seen that the output voltage of both circuits varies approximately linearly with time (for a large RC) and for either case, $\frac{dv_o}{dt} = \frac{V}{RC}$. However, the second term in both the expressions represent deviation from the linearity. We see that op-amp integrator is more linear than the simple RC circuit by a factor of $1/(1 - A_v)$.

Example 4.7

For the circuit shown in Fig. 4.27 if the input is a constant V, show that the output $v_o\,(t)$ is given by a differential equation.

Solution

The transfer gain of the circuit is,

$$\frac{V_o(s)}{V_i(s)} = -\frac{Z_F}{R_1} = -\frac{R_2 + \dfrac{R_3/sC}{R_3 + 1/sC}}{R_1}$$

$$= \frac{(R_2+R_3)+sC\,R_2R_3}{R_1(1+sC\,R_3)} \quad (4.91)$$

or, $$R_1(1+s\,C\,R_3)\,V_o(s) + [(R_2+R_3) + s\,C\,R_2R_3]\,V_i(s) = 0 \quad (4.92)$$

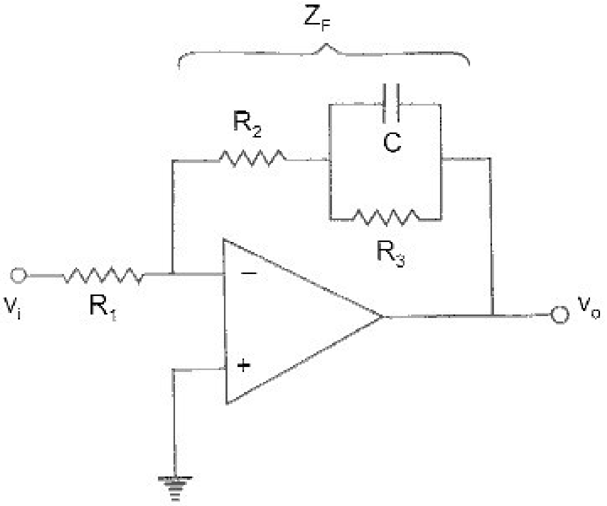


Fig. 4.27 Circuit for Example 4.7

Writing Eq. (4.92) in time domain $\left(s \to \frac{d}{dt}\right)$, we get

$$R_1\left(1 + CR_3\frac{d}{dt}\right)v_o(t) + \left[(R_2+R_3) + CR_2R_3\frac{d}{dt}\right]v_i(t) = 0 \quad (4.93)$$

Since $$v_i(t) = V$$

Therefore, $$\frac{dv_i(t)}{dt} = 0.$$

Hence $$CR_1R_3\frac{dv_o}{dt} + R_1v_o + (R_2+R_3)V = 0$$

or, $$C\frac{dv_o}{dt} + \frac{v_o}{R_3} + \frac{V}{R_1} + \frac{R_2}{R_1R_3}V = 0 \quad (4.94)$$

Example 4.8

Figure 4.28 shows a non-inverting integrator circuit. Show that $v_o = \frac{1}{RC}\int v_i\,dt$.

Solution

The voltage at the (+) input terminal of the op-amp due to the potential divider is,

$$V(+) = \frac{1/sC}{R + 1/sC}V_i(s) \quad (4.95)$$

The output voltage $V_o(s)$ for the non-inverting amplifier is

$$V_o(s) = \left(1 + \frac{1/sC}{R}\right)V(+) = \frac{1}{sRC}V_i(s) \quad (4.96)$$

Hence in time-domain, we get,

$$v_o = \frac{1}{RC}\int v_i\,dt.$$

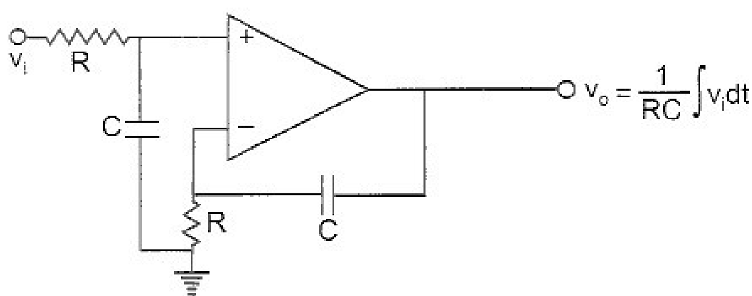


Fig. 4.28 Non-inverting integrator circuit

Note that there is no phase inversion in a non-inverting integrator.

4.12 ELECTRONIC ANALOG COMPUTATION

An analog computer performs linear operations such as multiplication by a constant, addition, subtraction and integration. Since these operations are sufficient for the solution of linear differential equation, it is possible to connect the various modules of an analog computer for obtaining the solution of differential equation.

We have already discussed the building blocks of the analog computer, that is, op-amp used as inverter, scale changer, summer, integrator, summing integrator etc. Potentiometer is widely used in analog computer to multiply voltages by a constant less than unity. The symbolic representation of a summer, potentiometer and summing integrator is shown in Fig. 4.29 (a, b, c).

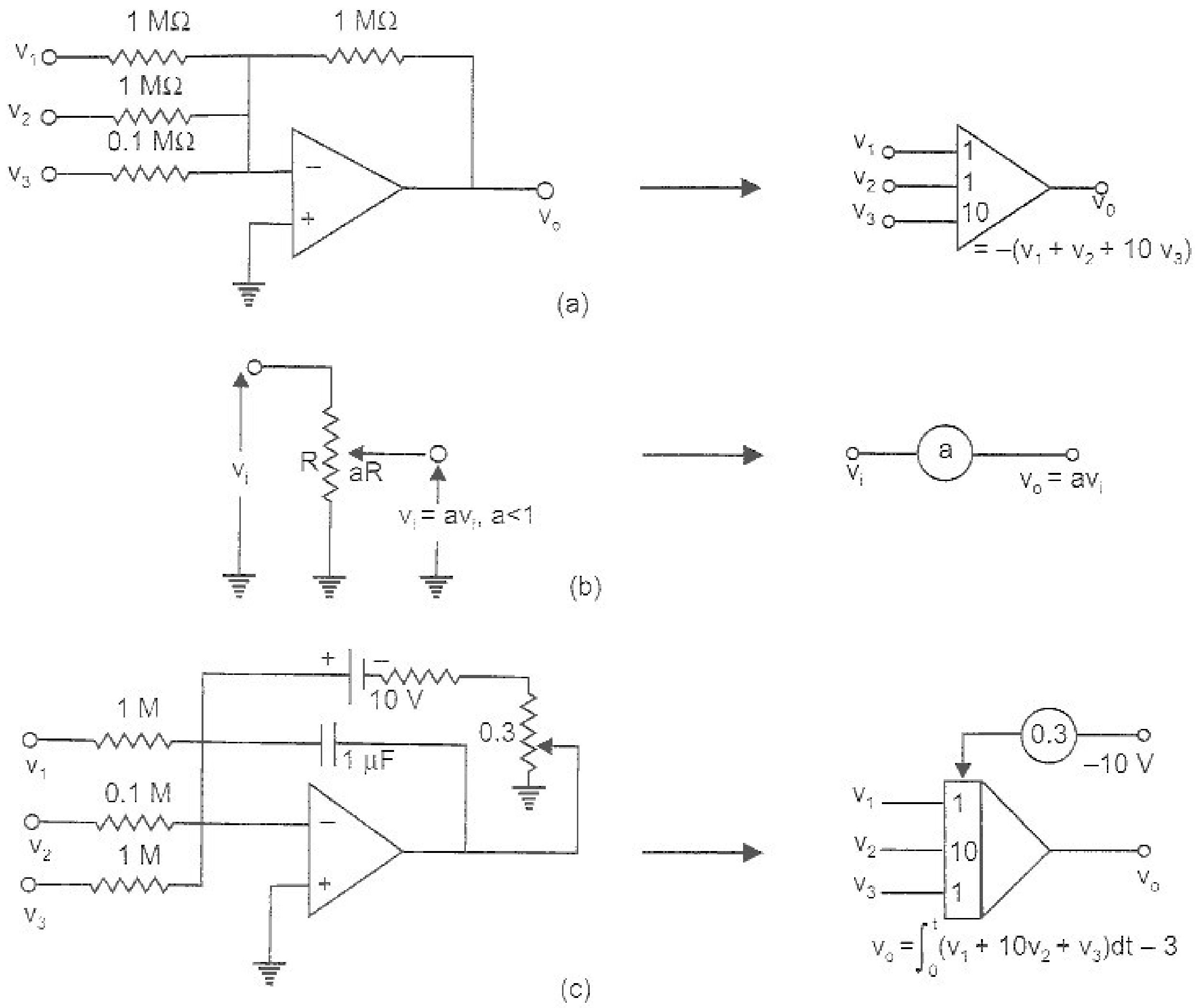


Fig. 4.29 (a) Summer and its symbolic representation, (b) Potentiometer and its symbol, (c) Summing integrator and its symbol

Let us now see how an analog computer can be used to solve a second order differential equation given as,

$$\frac{d^2y}{dt^2} + 5.4\frac{dy}{dt} + 0.58\,y = u(t) \tag{4.97}$$

with initial conditions,

$$y(0) = -4.8 \text{ and } \left.\frac{dy}{dt}\right|_{t=0} = \dot{y}(0) = 2.3$$

Rewrite Eq. (4.97) by keeping the highest order derivative on the left hand side and taking all other terms to the right side as

$$\ddot{y} = -5.4\,\dot{y} - 0.58\,y + u(t) \tag{4.98}$$

Assuming $\ddot{y}$ is available, it may be successively integrated to obtain $\dot{y}$ and y as shown in Fig. 4.30 (a). At the output of amplifier 4, i.e. point B, we obtain the sum

$$-5.4\,\dot{y} - 0.58\,y + u(t)$$

which is precisely equal to $\ddot{y}$ with which we started in Eq. (4.98). Thus points A and B may be connected together to get the computer set-up for solving the given differential equation. The initial conditions $y(0) = -4.8$ and $\dot{y}(0) = 2.3$ have to be placed in the computer set-up with the help of the reference voltage (either $+V_{ref}$ or $-V_{ref}$ as required) and potentiometer. One has to be careful about the polarity of the reference voltage for setting up the initial condition. As in the computer set-up, the output of integrator 1 is $-\dot{y}$ which is initially set to -2.3 V to achieve $\dot{y}(0) = 2.3$ V. Similarly the output of integrator 2 is y which is initially set to -4.8 V so that $y(0) = -4.8$ V.

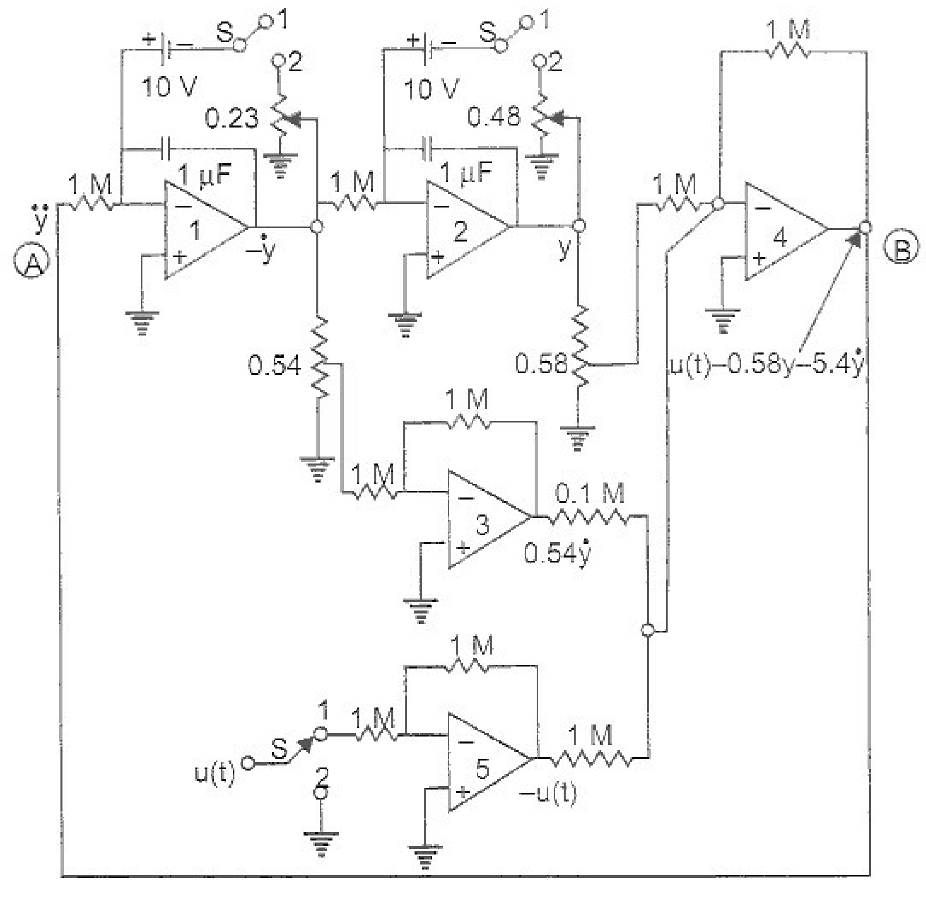


Fig. 4.30 (a) Simulation of 2nd order differential equation

The initial conditions are first established by putting the switch S in position 2. With S in position 1, the solution is obtained at the output terminal to which a CRO or plotter is connected.

Using the analog computer symbols, the set-up of Fig. 4.30 (a) is redrawn in Fig. 4.30 (b). Minimization of the components can be achieved using a summer integrator as shown in Fig. 4.30 (c).

The solution of Eq. (4.97) can also be obtained by using differentiators instead of integrators. However, the gain of a differentiator increases linearly with frequency and it tends to amplify noise, drift which may result in spurious oscillations. Therefore, integrators are invariably preferred over differentiators.

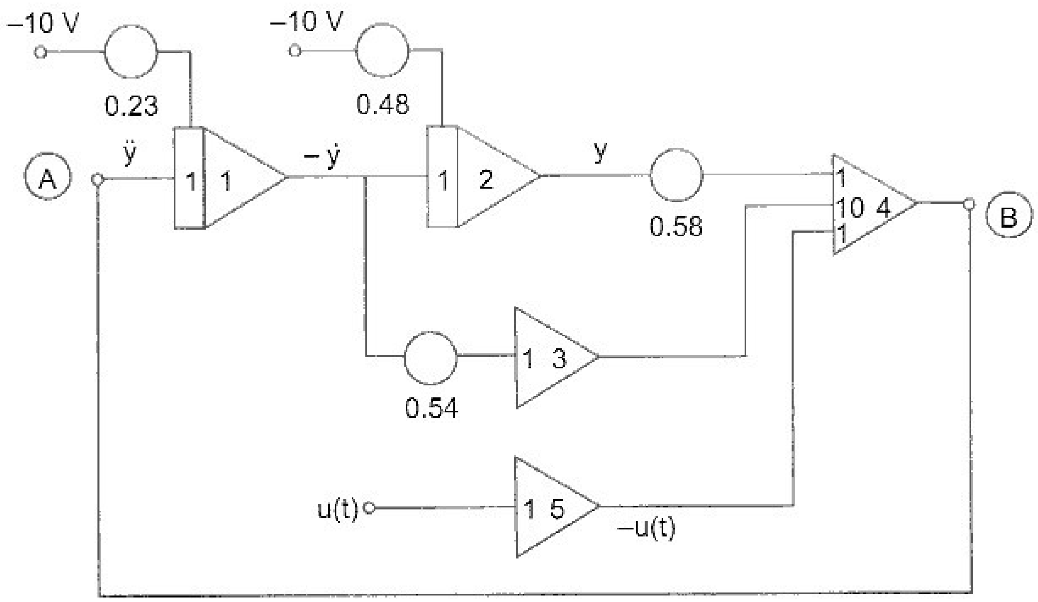


Fig. 4.30 (b) Symbolic diagram

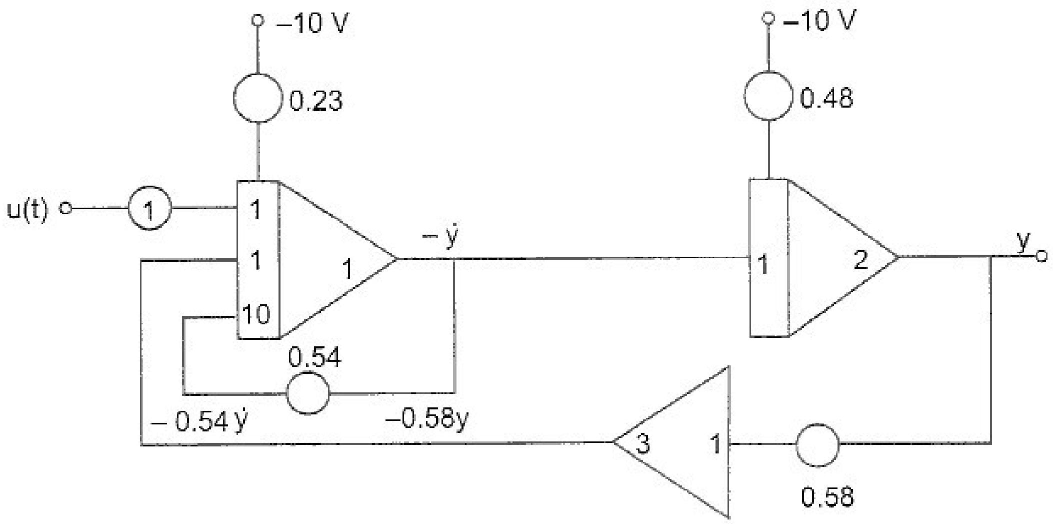


Fig. 4.30 (c) Minimum component simulation

Example 4.9

Set up an analog computer simulation to generate a sinusoidal signal 10 sin $3t$.

Solution

Let us first obtain a differential equation whose solution is 10 sin $3t$.

Let

$$x(t) = 10 \sin 3t$$

$$\dot{x}(t) = 30 \cos 3t \tag{4.99}$$

$$\ddot{x}(t) = -90 \sin 3t = -9x$$

The required differential equation is,

$$\ddot{x} + 9x = 0$$

and the initial conditions are obtained from Eq. (4.99) putting $t = 0$ as,

$$x(0) = 0,\ \dot{x}(0) = 30$$

Assuming that $\ddot{x}$ is available, $x(t)$ can be obtained by integrating x twice. The computer set up is shown in Fig. 4.31.

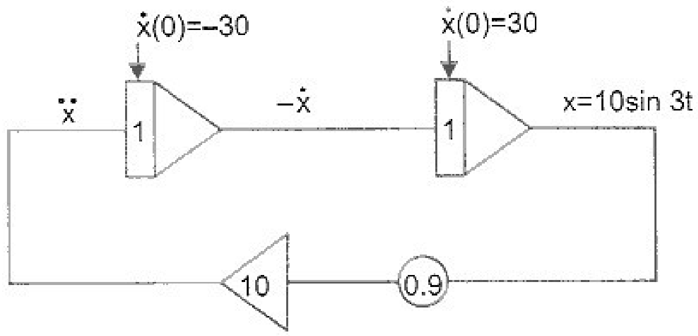


Fig. 4.31 Simulation of 10 sin 3t

Simultaneous Equations

A set of simultaneous equations in two unknowns can also be solved using analog simulation. Consider two first order differential equations:

$$\frac{dx}{dt} = -a_1x - b_1y + c_1f \tag{4.100}$$

and

$$\frac{dy}{dt} = -a_2x - b_2y + c_2f \tag{4.101}$$

where x and y are unknown variables, f is the input and all coefficients are known constants. Equations (4.100) and (4.101) may be simulated separately as shown in Figs. 4.32 (a) and (b). Now interconnect the two systems to get the unknown x and y as shown in Figs. 4.32 (c).

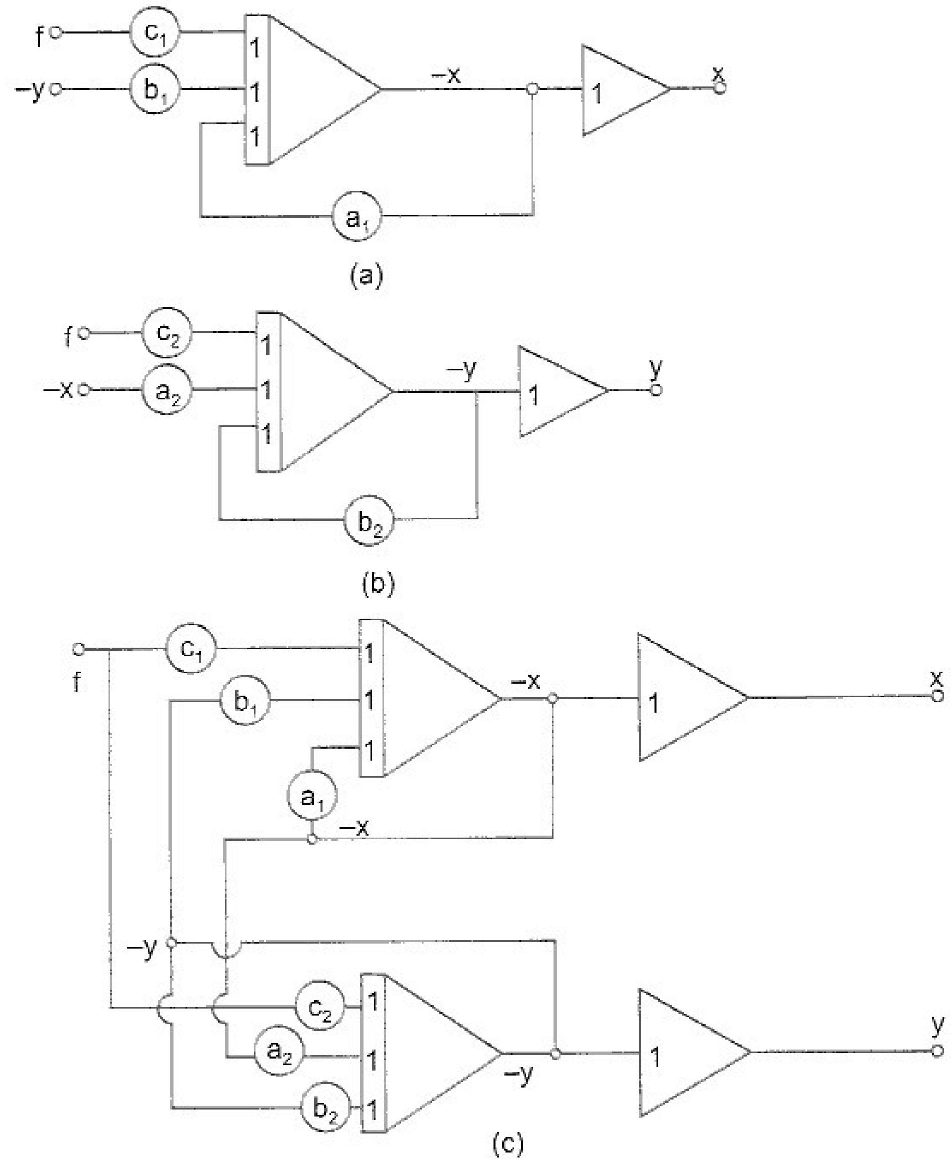


Fig. 4.32 (a) Simulation of Eq. (4.100), (b) Simulation of Eq. (4.101), (c) Final circuit

The simulation procedure can be extended to any number of simultaneous equations.

Simulation of Transfer Functions

Another important problem that one come across is to develop a circuit that has a given transfer function. As an example, in the design of an electric filter consider a first order transfer function,

$$H(s) = \frac{V_o}{V_i} = \frac{-K}{s+a} \tag{4.102}$$

so, $$V_o(s + a) = -KV_i$$

or, $$-s\,V_o = a\,V_o + K\,V_i \tag{4.103}$$

which may be written in time-domain as,

$$v_o = -\int (a v_o + K v_i)\,dt \tag{4.104}$$

Thus we need a summing integrator and its simulation is shown in Fig. 4.33 (a). The corresponding electric circuit is shown in Fig. 4.33 (b).

In another example, let us simulate the transfer function

$$\frac{V_o}{V_i} = \frac{K(s+a)}{s^2 + sb + c} \tag{4.105}$$

which may be written in s-domain as,

$$s^2 V_o = -sbV_o - cV_o + Ks\,V_i + Ka\,V_i$$

or $$-sV_o = bV_o - KV_i - \frac{1}{s}(Ka\,V_i - cV_o) \tag{4.106}$$

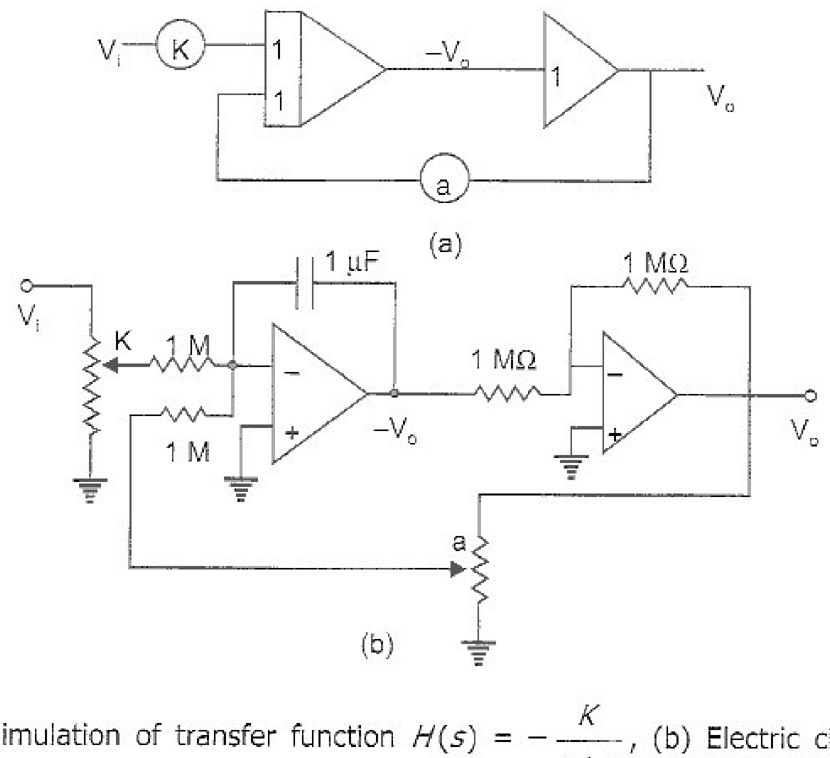


Fig. 4.33 (a) Simulation of transfer function $H(s) = -\frac{K}{s+a}$, (b) Electric circuit for (a)

Thus V_o is the output of an integrator whose input is $-sV_o$ which is the sum of the terms on the right side of Eq. (4.106). The term $-\frac{1}{s}$ $(Ka\,V_i - c\,V_o)$ is the output of an integrator whose inputs are KaV_i and $-c\,V_o$ as shown in Fig. 4.34 (a). The complete circuit is shown in Fig. 4.34 (b).

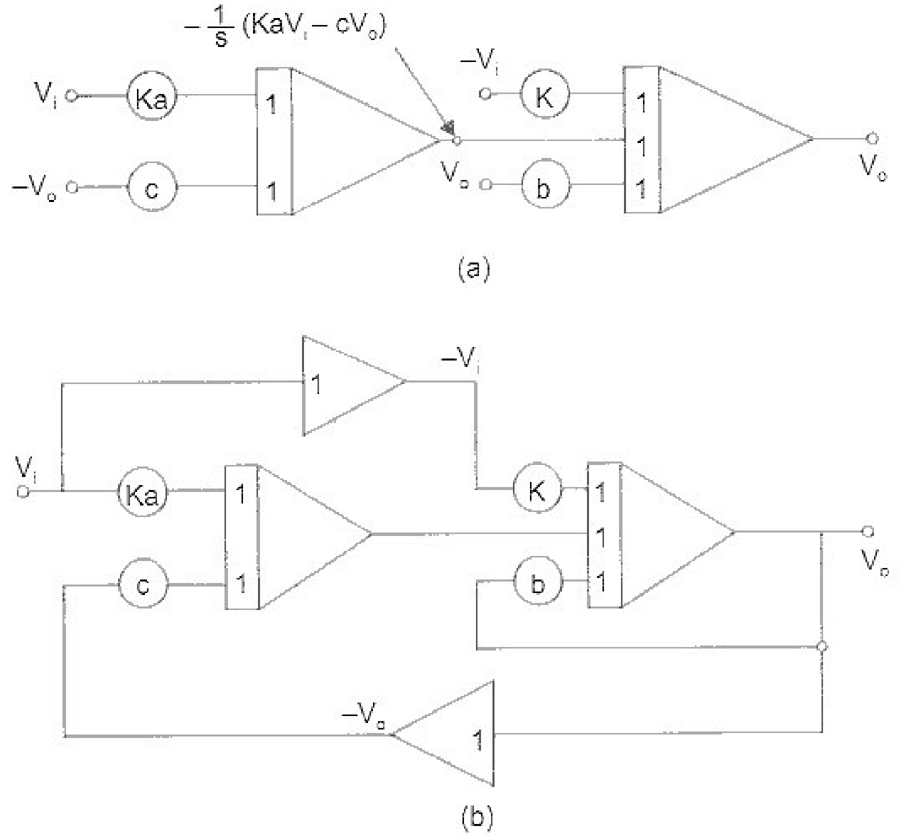


Fig. 4.34 (a) Preliminary simulation, (b) Final simulation

4.13 OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA)

In Sec. 4.5, we have discussed the use of 741 op-amp as a voltage to current converter. A voltage to current converter is an amplifier which produces an output current proportional to an input voltage. The constant of proportionality is the transconductance of the amplifier and therefore such amplifiers are also known as transconductance amplifier. Due to wide applications, specially designed single chip transconductance amplifiers are available, called, operational transconductance amplifier (OTA). The symbolic representation of an OTA is shown in Fig. 4.35. An OTA is a voltage-input, current-output device such that

$$I_o = g_m V_{in} = g_m (V_1 - V_2) \tag{4.107}$$

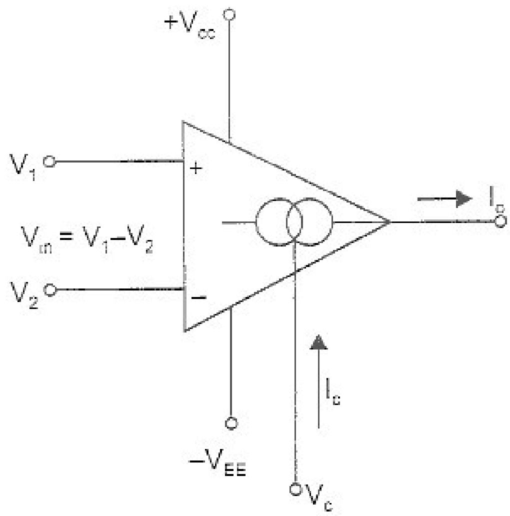


Fig. 4.35 Symbol for OTA

where g_m is the transconductance, or gain of the OTA. The unique feature of an OTA is that it is possible to vary g_m over a wide range by means of an external control current. OTAs are used to implement programmable amplifiers and integrators in audio processing and electronic music synthesis. They are also used as current switches in sample-and-hold applications. Another important application of OTA using VLSI technique is in Neural networks. Popular OTAs are the CA3080 (RCA), the LMI 3600/700 (National Semiconductor) and the NE 5517 (Signetics).

The simplified internal circuit diagram of an OTA is shown in Fig. 4.36. Transistors Q_1 and Q_2 form a differential pair. Current mirror $Q_3 - Q_4$ accepts the control current I_c which can be adjusted by an external resistance R_{ext} and control voltage V_c. Due to current mirror $Q_3 - Q_4$, we get $I_4 = I_c$. The current I_4 is divided at the emitters of Q_1 and Q_2. Thus

$$I_1 + I_2 = I_4$$

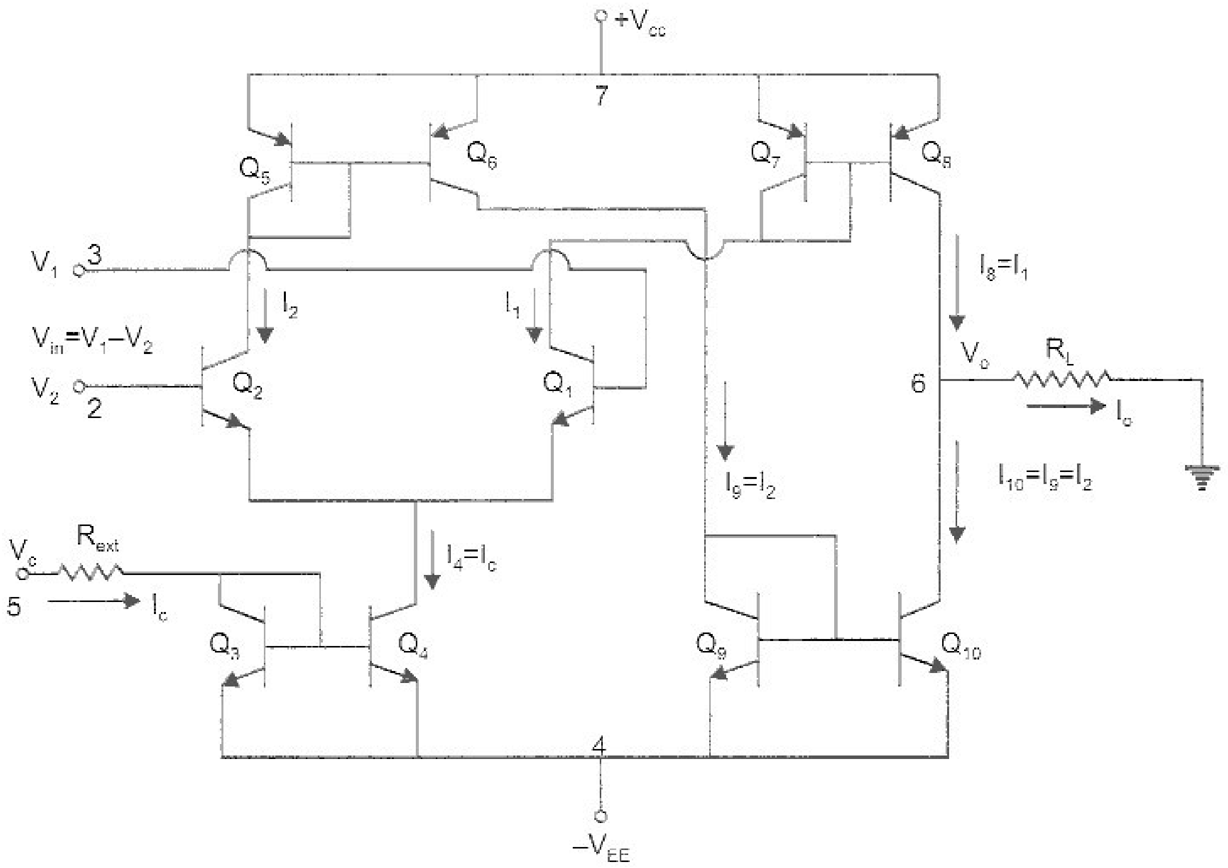


Fig. 4.36 Simplified internal diagram of OTA

Current mirror $Q_5 - Q_6$ duplicates I_2 to yield $I_9 = I_2$. The current I_2 is in turn duplicated by current mirror $Q_9 - Q_{10}$ to produce $I_{10} = I_9 = I_2$. Similarly, current mirror $Q_7 - Q_8$ duplicates I_1 to yield $I_8 = I_1$. By KCL, we have,

$$I_o = I_8 - I_{10} = I_1 - I_2 \tag{4.108}$$

Hence, the voltage gain A_v can be written as

$$A_v = \frac{V_o}{V_{in}} = \frac{I_o R_L}{V_{in}} = g_m R_L \tag{4.109}$$

The transconductance g_m can be calculated as

$$I_1 \cong I_s \; e^{V_1/V_T} \tag{4.110}$$

and

$$I_2 \cong I_s \; e^{V_2/V_T} \tag{4.111}$$

where, I_s = reverse saturation current of transistor Q_1, Q_2 assumed to be equal and, V_T = volts equivalent of temperature

Now $$I_c = I_1 + I_2 = I_s\left[e^{V_1/V_T} + e^{V_2/V_T}\right] \tag{4.112}$$

or, $$I_s = \frac{I_c}{e^{V_1/V_T} + e^{V_2/V_T}} \tag{4.113}$$

So, $$I_1 = \frac{I_c\, e^{V_1/V_T}}{e^{V_1/V_T} + e^{V_2/V_T}} \tag{4.114}$$

and $$I_2 = \frac{I_c\, e^{V_2/V_T}}{e^{V_1/V_T} + e^{V_2/V_T}} \tag{4.115}$$

Hence $$I_1 - I_2 = I_c\, \frac{e^{V_1/V_T} - e^{V_2/V_T}}{e^{V_1/V_T} + e^{V_2/V_T}} \tag{4.116}$$

Multiplying both the numerator and denominator by $e^{-\frac{V_1 + V_2}{2}}$; Eq. (4.116) can be expressed in terms of voltage difference $(V_1 - V_2)$ as,

$$I_o = I_1 - I_2 = I_c\, \frac{e^{\frac{V_1 - V_2}{2V_T}} - e^{-\frac{V_1 - V_2}{2V_T}}}{e^{\frac{V_1 - V_2}{2V_T}} + e^{-\frac{V_1 + V_2}{2V_T}}} \tag{4.117}$$

$$= I_c \tanh\left(\frac{V_1 - V_2}{2V_T}\right) \tag{4.118}$$

A plot of output current I_o as a function of $(V_1 - V_2)$ is shown in Fig. 4.37. A transconductance amplifier basically computes a tan-hyperbolic. It operates linearly for a very small range of inputs and smoothly transits to saturation. The transconductance is given by

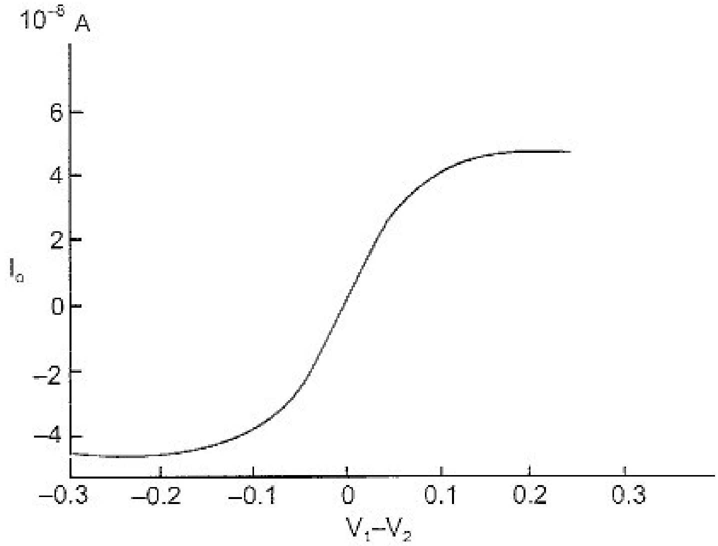


Fig. 4.37 Transfer characteristics of OTA

$$g_m = \left|\frac{\partial I_o}{\partial V_{in}}\right| = \frac{I_c}{2V_T} \tag{4.119}$$

$$\left[\text{with the assumption that for small } V_1 - V_2\,;\ \tanh\left(\frac{V_1 - V_2}{2V_T}\right) \approx \frac{V_1 - V_2}{2V_T}\right]$$

Then the voltage gain A_v is,

$$A_v = g_m R_L = \frac{I_c R_L}{2V_T} \tag{4.120}$$

Thus the voltage gain of the OTA circuit can be externally controlled by the control current I_c.

There are two main differences between the OTA and the conventional op-amp. An OTA is a voltage controlled current source (VCCS) whereas the conventional op-amp is a voltage controlled voltage source (VCVS). Now, since the OTA is a current source, its output impedance is high, in contrast to the op-amp's very low output impedance. We know that the low output impedance is a desirable feature in general amplifiers used to drive resistive loads, therefore commercial OTA's such as National Semiconductors LM13600, have on-chip controlled impedance buffers. Another difference between OTA's and op-amps is that it is possible to design circuits using the OTA without employing negative feedback. Thus, instead of employing feedback to reduce the sensitivity of a circuit's performance to device parameters, the transconductance is treated as a design parameter in place of resistors and capacitors in op-amps based circuits.

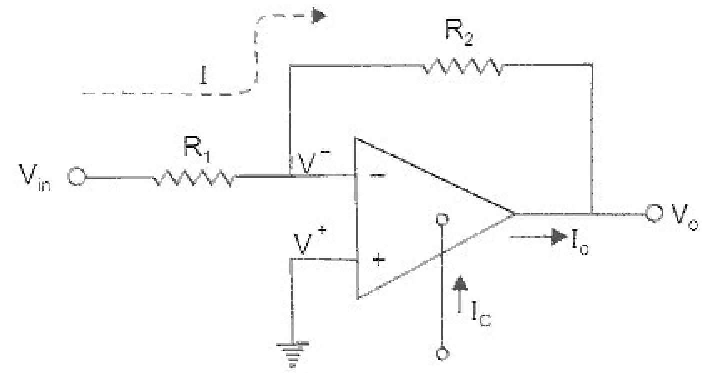


Fig. 4.38 An inverting amplifier using OTA

Figure 4.38 shows an inverting amplifier using OTA which can provide not only controllable gain, but also uses negative feedback to reduce the output resistance.

The expression of the voltage gain may be obtained as follows:

From the basic behaviour of the OTA, we may write

$$\begin{aligned} I_o &= g_m(V^+ - V^-) \\ &= -g_m V^- \qquad [\text{As } V^+ = 0] \end{aligned} \tag{4.121}$$

KVL around the OTA gives

$$\frac{V_{in} - V^-}{R_1} = \frac{V^- - V_o}{R_2} = I \tag{4.122}$$

Also, since

$$I_o = -I$$

$$g_m V^- = \frac{V^- - V_o}{R_2}$$

or $$V^-(1 - g_m R_2) = V_o$$

or $$V^- = \frac{V_o}{1 - g_m R_2} \tag{4.123}$$

Putting the value of V^- from Eq. (4.123) to Eq. (4.122) and simplifying, we get

$$\frac{V_o}{V_{in}} = \frac{1 - g_m R_2}{1 + g_m R_1} \tag{4.124}$$

For $g_m R_1 >> 1$, Eq. (4.124) reduces to

$$\frac{V_o}{V_{in}} \cong \frac{-R_2}{R_1} \quad (4.125)$$

An all-OTA amplifier without negative feedback is shown in Fig. 4.39.

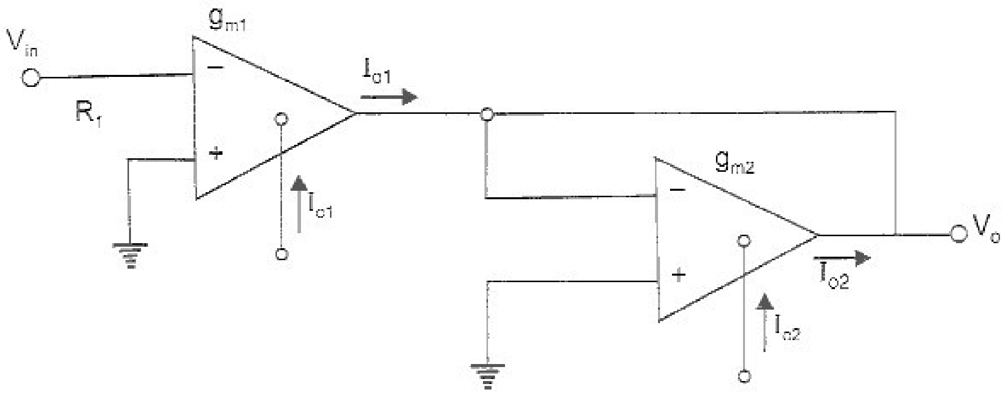


Fig. 4.39 OTA amplifier without feedback

The amplifier circuit uses two OTA's with transconductance g_{m1} and g_{m2}.
In Fig. 4.39

$$I_{o1} = g_{m1}(V^+ - V^-) \quad (4.126)$$

$$= g_{m1}(0 - V_{in}) = -g_{m1} V_{in} \quad (4.127)$$

and

$$I_{o2} = g_{m2}(V^+ - V^-) \quad (4.128)$$

$$= g_{m2}(0 - V_o)$$

$$= -g_{m2} V_o \quad (4.129)$$

Since

$$I_{o2} = -I_{o1}$$

$$-g_{m2} V_o = + g_{m1} V_{in}$$

or

$$\frac{V_o}{V_{in}} = \frac{-g_{m1}}{g_{m2}} \quad (4.130)$$

The voltage gain, therefore is completely controllable by the external control currents I_{c1} and I_{c2} which control g_{m1} and g_{m2} respectively.

Figure 4.40 shows the schematic of widely used CA3080 OTA. Its transconductance or gain (g_m) is controlled by the current I_c driven into pin 5. At room temperature

$$g_m = \frac{I_c}{2\,V_T} = \frac{I_c}{2 \times 26 \text{ mV}} \quad (4.131)$$

where V_T = 26 mV at room temperature

or,

$$g_m = \frac{19.2}{V} I_c \quad (4.132)$$

where V stands for volt as unit.

(The unit of transconductance g_m is Siemens)

This relationship holds linearly for 0.1 μA < I_c < 400 μA.
From Eq. (4.107), we get

$$I_o = \left(\frac{19.2}{V} I_c\right) V_{in} \tag{4.133}$$

For CA3080, Eq. (4.133) is linear for I_o < 400 μA and V_{in} < 20 mV. If input voltage V_{in} becomes greater than 20 mV, Eq. (4.133) will no longer remain linear. So input voltage should be restricted to less than 20 mV. Another limitation is that since input signals are fed directly into the bases of the first stage, input impedance is of the order of 10 kΩ to 30 kΩ. To avoid loading, op-amp voltage followers should be used to buffer input signals.

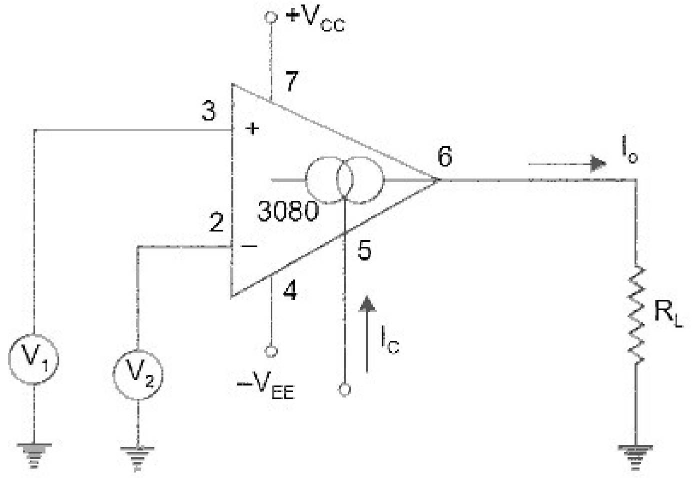


Fig. 4.40 Schematic of CA3080 OTA

Figure 4.41 shows the various ways to set I_c. A fixed current I_c is obtained by connecting a resistance R_{ext} between pin 5 and ground as in Fig. 4.41 (a). The control current I_c can be written as,

$$I_c = \frac{|-V_{EE}| - 0.6\text{ V}}{R_{ext}} \tag{4.134}$$

Here, 0.6 V is the forward voltage drop of the diode connected transistor Q_3 in Fig. 4.36. In Fig. 4.41 (b), I_c is controlled by control voltage V_c and I_c is given by

$$I_c = \frac{V_c + |-V_{EE}| - 0.6\text{ V}}{R_{ext}} \tag{4.135}$$

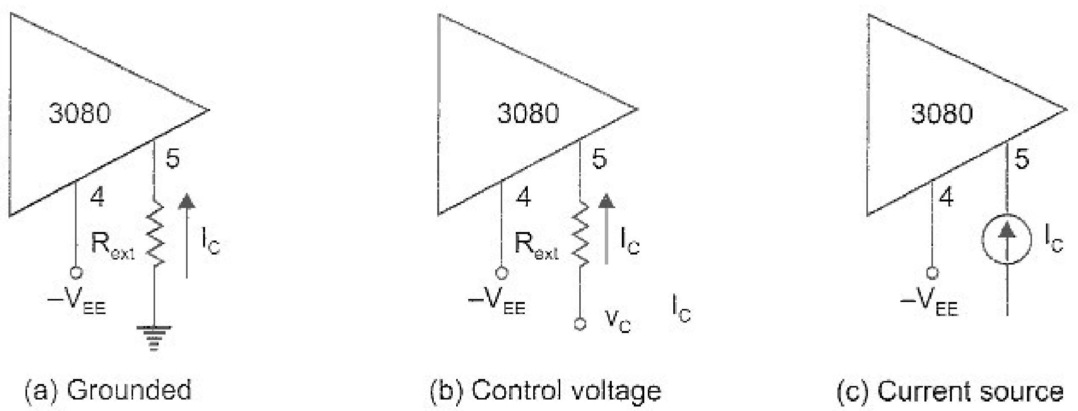


Fig. 4.41 Circuits for setting control current I_c

The control current I_c can be set directly with a current source as in Fig. 4.41 (c). This current source could be made with an FET, a BJT, an op-amp, an IC current source chip or another 3080.

An OTA can be viewed as a programmable resistor whose resistance is set by control current I_c. Since

$$I_o = \frac{19.2}{V} I_c V_{in}$$

The resistor $$R = \frac{V_{in}}{I_o} = \frac{V}{19.2\, I_c} = \frac{10.3\ \text{mV}}{I_c} \tag{4.136}$$

Figure 4.42 shows an electronically tuned resistor using two OTAs. The 'resistance' expression for the circuit can be derived as follows:

For U_2,

$$I_{c2} = \frac{|-V_{EE}| - 0.6\ \text{V}}{R_{ext}} \approx \frac{|-V_{EE}|}{R_{ext}} \tag{4.137}$$

$$I_{o2} = \left(\frac{19.2}{V}\right) I_{c2} V_c = \frac{19.2}{V} \frac{|-V_{EE}|}{R_{ext}} V_c \tag{4.138}$$

For U_1,

$$I_{c1} = I_{o2}$$

and $$I_o = \frac{19.2}{V} I_{c1} V_{in} = \left(\frac{19.2}{V}\right)\left[\left(\frac{19.2}{V}\right)\frac{|-V_{EE}|}{R_{ext}} V_c\right] V_{in}$$

$$= \left[\left(\frac{19.2}{V}\right)^2 \frac{|-V_{EE}|}{R_{ext}}\right] V_c V_{in} \tag{4.139}$$

$$R = \frac{V_{in}}{I_o} = \frac{R_{ext}}{(369/V^2)|-V_{EE}|V_c} \tag{4.140}$$

Thus the circuit of Fig. 4.42 can be considered as a voltage controlled resistance whose value is set in accordance with Eq. (4.140). However, V_{in} must be restricted to be less than 20 mV and I_o should be less than 400 μA.

The application of operational transconductance amplifier on sample-and-hold circuit is shown in Fig. 4.43. The control terminal is biased ***on*** (= $+V_{CC}$) and ***off*** (= $-V_{EE}$) to sample and hold an input signal across a holding capacitor. In the sample mode, the control voltage V_c is high, that is at $+V_{CC}$, and the output of the transconductance amplifier charges the holding capacitor C_H to a voltage equal to V_{in}. In the hold mode, I_c is reduced to zero, and the output of the OTA is at virtually open circuit so that the sampled input voltage is held on C_H. The decay of the voltage across C_H during the hold mode depends upon the output impedance of the OTA and the input resistance of the buffer stage.

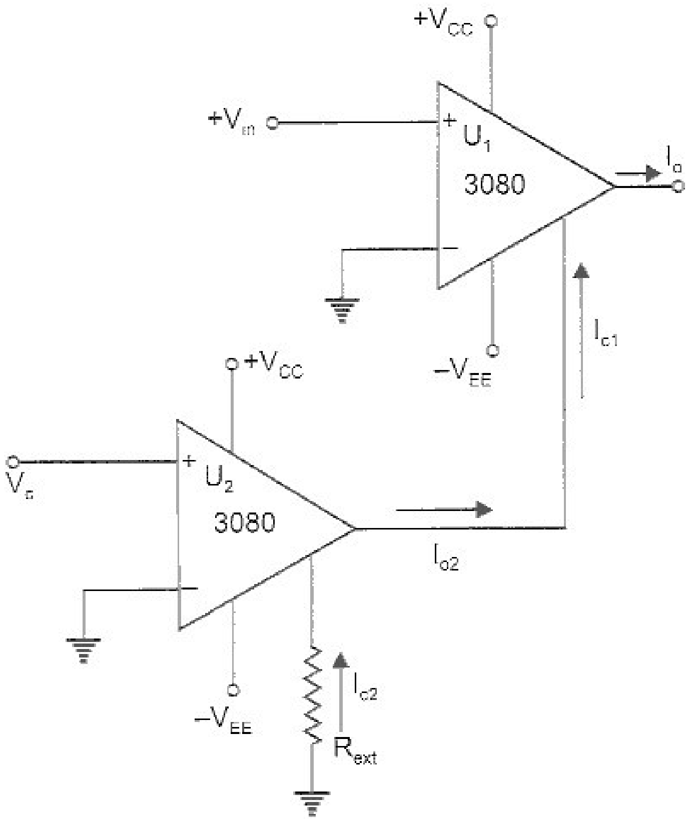


Fig. 4.42 Electronically tunable 'resistor' using two OTAs

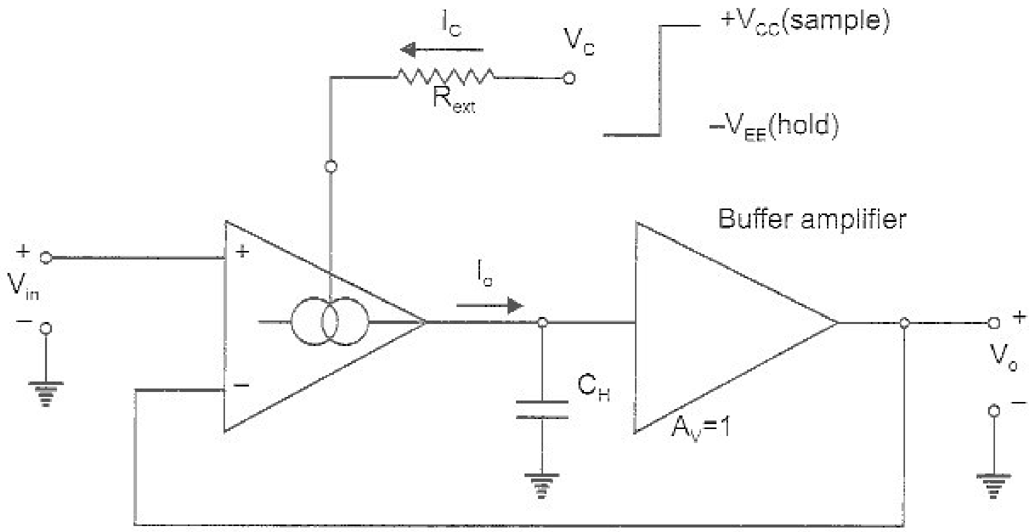


Fig. 4.43 Sample and hold circuit using OTA

There are a few basic limitations of OTAs. One is the severe restrictions on voltage and current magnitudes as already discussed. Another limitation is that as the control current I_c is varied to adjust gain or resistance, several other parameters of the amplifier are also affected. These include offset voltage, input bias current and slew-rate etc. To minimize these effects, input, output buffers and frequency compensating techniques may have to be used. Temperature also alters the performance of the OTA. The proportionality constant 19.2/V is valid only for room temperature. It may be noticed that the OTA performance is inversely proportional to temperature.

4.14 POWER AMPLIFIERS

Most of the practical electronic systems consist of a number of amplifying stages in cacade. The function of the input and intermediate stages is to amplify the small signals received from an input source, such as from a cassette, CD, microphone or an antenna. The signal is amplified to a large value sufficient to drive the final device which may be a speaker in a public address system or a music system or a CRT or a servomotor etc. The input amplifying stages basically operate as voltage amplifiers. Since signal voltages and currents are small, the power handling capacity and power efficiency are of little concern for the small signal amplifiers.

The last stage or the output stage of any electronic system, however has to supply large current and sufficient power to the load which may be a loudspeaker or a servomotor. Such amplifiers are called large signal amplifiers or power amplifiers. The important requirements of a power amplifier are:

1. It should have low output impedance so as to provide impedance matching for maximum power transfer.
2. High power conversion efficiency.
3. As large amount of power gets dissipated at the junction of power transistor, heat sinks have to be used.

Classification of Output Stages

One way of classifying amplifiers is based upon the position of quiescent (Q) point on the load line and the shape of the output collector current waveform when a sinusoidal input signal is applied. The Q-point of the BJT need not necessarily be in the middle of the load line. Depending upon the position of the Q-point, amplifiers may be classified as class-A, class-B, class-AB or class-C as per the following definitions.

Class A

The operating point is chosen somewhere in the center of the load line as shown by point A in Fig. 4.44. The amplitude of the input signal is such so that the transistor conducts for the entire cycle of the input signal that is for 360°. The output collector current waveform for class-A amplifier is therefore as shown in Fig. 4.45 (a).

The input stage or intermediate stages of any electronic system are operated in class-A operation.

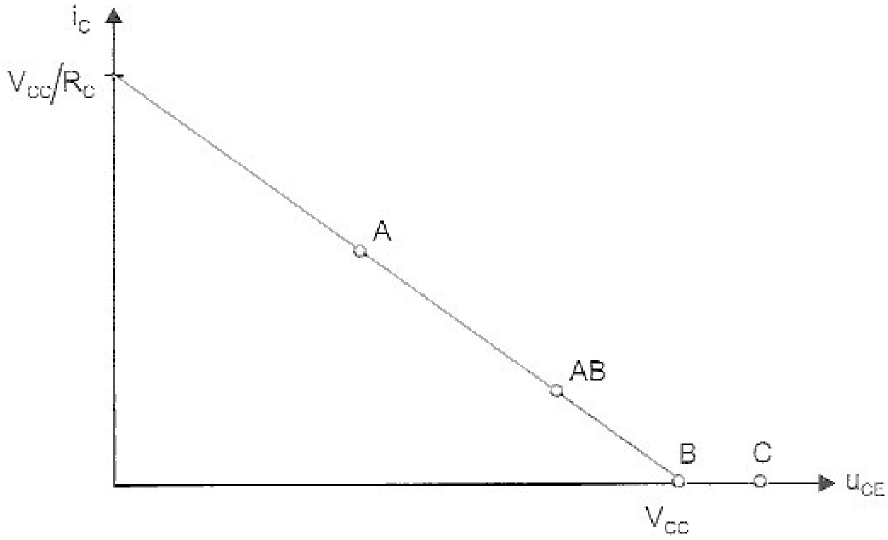


Fig. 4.44 Classification of amplifiers based on the position of Q-point

Class B

The Q-point is selected at the extreme end of the load line as shown by point B in Fig. 4.44. The quiescent current is, therefore, zero and the transistor conducts only for half the cycle of the input waveform. The output collector current consists of half waves only as shown in Fig. 4.45 (b). Obviously, the output is not a faithful reproduction of the input as only half cycle is present in the output. In a practical circuit, two transistors are operated in class-B in such a manner so that one transistor conducts during positive half cycle of the input waveform and the other conducts during the negative half cycle so that the output current through the load is of full cycle. This type of connection is referred to as push-pull operation and is most commonly used in power amplifiers.

Class AB

An intermediate operation between class A and class B is class AB in which the operating point is selected between the points A and B in Fig. 4.44 as shown by point AB. As a result, the transistor conducts for an internal greater than the half cycle but less than 360°. The output collector current waveform is as shown in Fig. 4.45 (c).

Class C

The operating point is selected beyond the load line on the horizontal axis as shown by point C in Fig. 4.44. The transistor in this case conducts for an interval lesser than the half cycle resulting in pulsating current waveforms as shown in Fig. 4.45 (d). Class C amplifiers are usually employed in radio frequency (RF) power amplification (such as in mobile phones, radio and TV transmitters). These type of amplifiers have not been included in the book.

Power Conversion Efficiency (η)

An amplifier is basically converting dc power/energy supplied by the dc power supply into ac power or amplified signal at the load. The ability of an active device to connect dc power

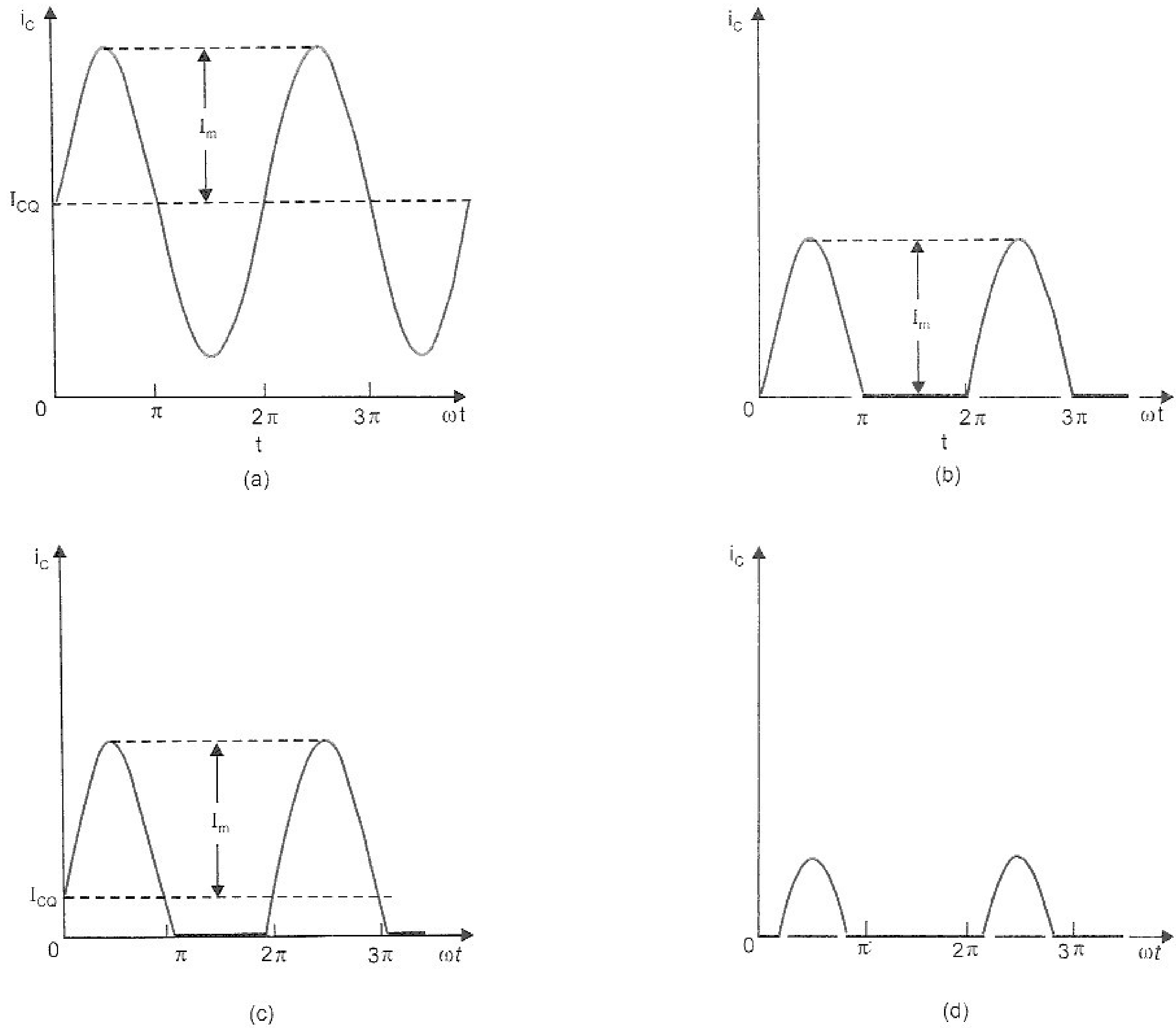


Fig. 4.45 Collector current waveforms for transistors operating in (a) Class A, (b) Class B, (c) Class AB, and (d) Class C amplifier stages

into ac power is measured by η called the power conversion efficiency. It is usually given as a percentage, so

$$\eta = \frac{\text{ac signal power delivered to load}}{\text{dc power supplied to the circuit}} \times 100\%$$

The maximum power conversion efficiency of a class-A single stage CE amplifier is only 25%. In many practical applications, such as a radio or public address system, the load is a loudspeaker and is coupled to the amplifier through a transformer. The maximum efficiency for a transformer coupled load increases to 50%. Efficiencies up to 78.5% can be achieved by using two transistors in push-pull configuration and biasing each transistor in class-B operation. In discrete circuits, a center-tapped input transformer is used for providing signals to the two transistors. Similarly the output is obtained using a center-tapped output transformer.

In IC amplifer circuits, use of transformers is not possible as it is difficult to fabricate these. The output stage of a commercially available 741 op-amp uses a complementary symmetry emitter follower which is discussed in detail in Ch-2, Sec. 2.4.8.

Monolithic Power Amplifiers

The general purpose op-amp 741 can deliver about 100 mW of power which is not sufficient for most of the applications. A wide range of IC power amplifiers are now commercially available. National Semiconductors produces two popular IC power amplifiers LM380 and LM384. The pin configuration, circuit diagram and typical applications are discussed.

LM380 Power Audio Amplifiers

The LM380 power audio amplifier is designed to deliver 2.5W (r.m.s) to a capacitively coupled 8-Ω load. Figures 4.46 (a) and (b) shows the pin configuration and block diagram of LM380 respectively. It is available in a 14-pin DIP package. A copper lead frame used with the center three pins on either side (3, 4, 5 on the left and 10, 11, 12 on the right) forms a heat sink. Thus there is no need to use a separate heat sink for the audio amplifier. The internal schematic diagram of LM380 is shown in Fig. 4.46 (c).

It can be seen that it has four stages: (i) PNP emitter follower; (ii) differential amplifier (iii) common emitter and (iv) quasi-complementary emitter follower.

The input is coupled through inverting input terminal (pin-6) and non-inverting input terminal (pin-2) to emitter follower stages composed of PNP input transistors Q_1 and Q_2. The output from Q_1 and Q_2 drives the PNP Q_3–Q_4 differential pair. Transistors Q_5 and Q_6 constitute the collector loads for the PNP differential pair. The current in the PNP differential pair Q_3–Q_4 is established by Q_7, R_3 and V_{CC}. The transistors Q_7 and Q_8 form the current mirror and establish current in Q_9 which forms the common emitter voltage gain stage. The output of the differential pair Q_3–Q_4 is taken at the junction of Q_4 and Q_6 and is applied as input to CE voltage gain stage. The capacitor C (10 pF) between base and collector of Q_9 provides internal compensation and establishes the upper cut-off frequency of 100 kHz at 2W for 8 Ω loads. As Q_7-Q_8 form a current mirror, the current through diodes D_1 and D_2 is same as that through R_3. The output stage is a quasi (false)–complementary pair emitter follower formed by transistors Q_{10} and Q_{12}. In fact, the combination of PNP transistor Q_{11} and NPN transistor Q_{12} has the power capability of an NPN transistor but the characteristics of a PNP transistor. The diodes D_1 and D_2 are used to minimize cross-over distortion. The resistors R_6 and R_7 are used for current limiting. The quiescent output voltage is established at $V_{CC}/2$ by resistors R_3 and R_5. To decouple the input stages from the supply voltage V_{CC}, a by-pass capacitor of the order of micro farads should be connected between the by-pass terminal (pin 1) and ground (pin 7).

Some of the important features of LM380 are listed below:

(i) Internally fixed gain of 50 (34 dB)
(ii) Wide supply voltage range (5 V to 22 V)
(iii) High peak current capability (1.3 A max.)
(iv) Low total harmonic distortion (0.2%)
(v) Output automatically self-centering to one-half of the supply voltage
(vi) Output short circuit proof with internal thermal limiting
(vii) High input impedance (15° kΩ)
(viii) BW of 100 kHz at an output power of 2W and a load of 8 Ω.

Another commonly used audio power amplifier is LM384. The internal diagram is similar to that of LM380, except that it is designed to deliver 5W power output.

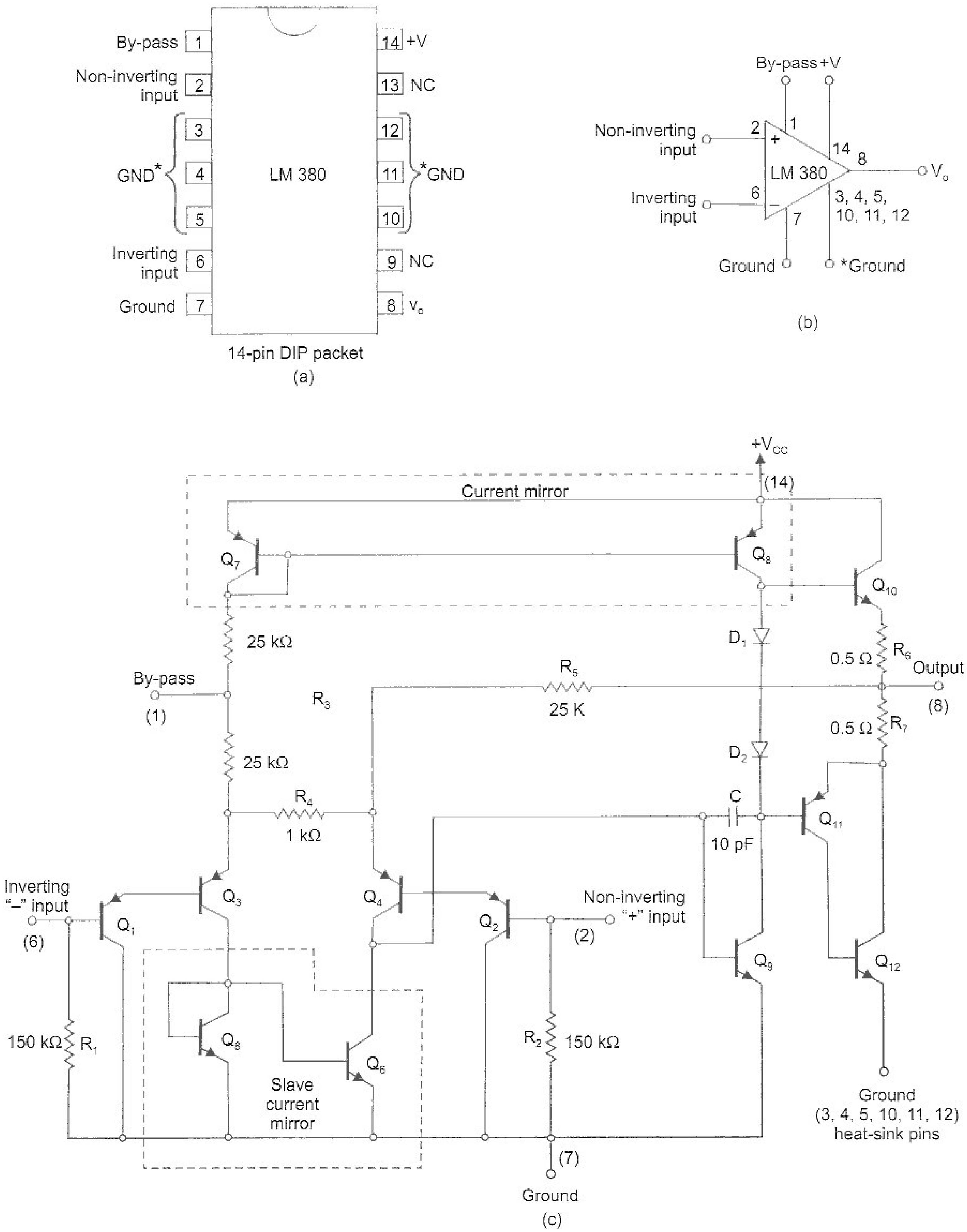
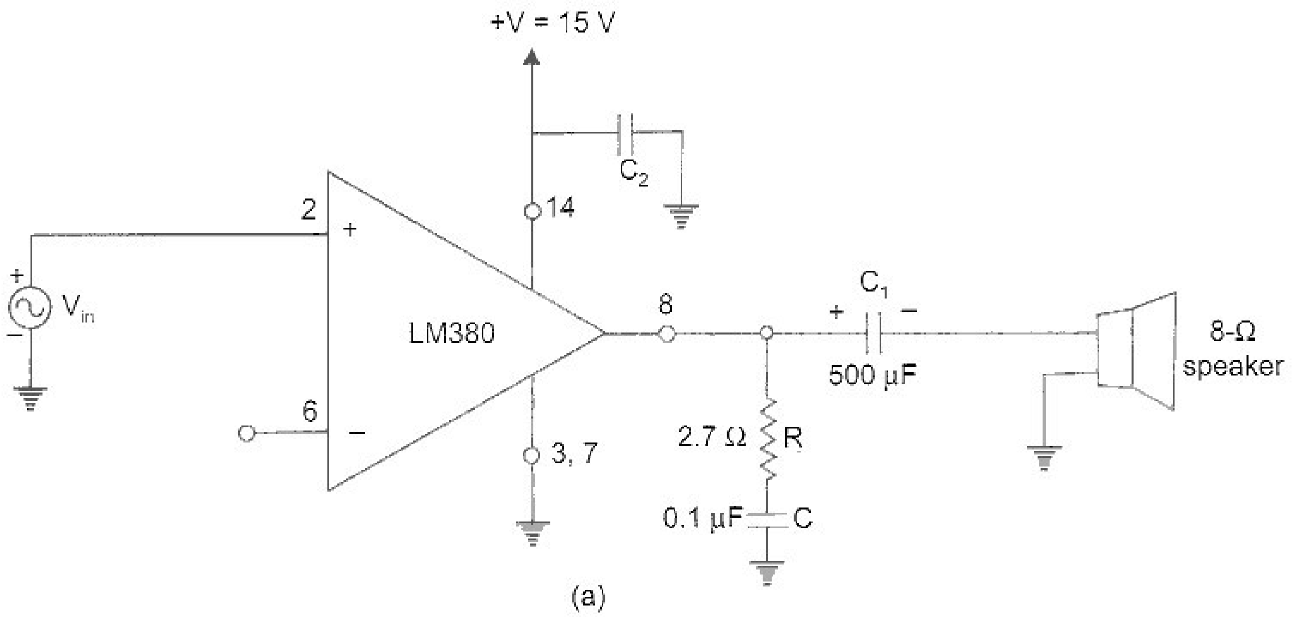


Fig. 4.46 (a) Pin diagram for LM380, (b) Block diagram, (c) Schematic diagram

Applications

The most basic application of LM380 is its use as an audio power amplifier. The IC can be used both in inverting and non-inverting configuration. Figure 4.47 (a) shows its use in non-inverting mode. The inverting terminal (pin 6) can be either started to ground, left open or returned to ground through a resistor or a capacitor. The capacitor C_2 is used to cancel the effects of inductance in the power supply leads. A lag compensating RC network is usually used at the output terminal (pin 8) to eliminate 5 to 10 MHz oscillations especially in RF sensitive environment.

The gain of LM380 is internally fixed at 50 but it is possible to get the gain increased up to 300 using positive feedback. Figure 4.47 (b) shows LM380 used in the inverting configuration using positive feedback to increase the gain.



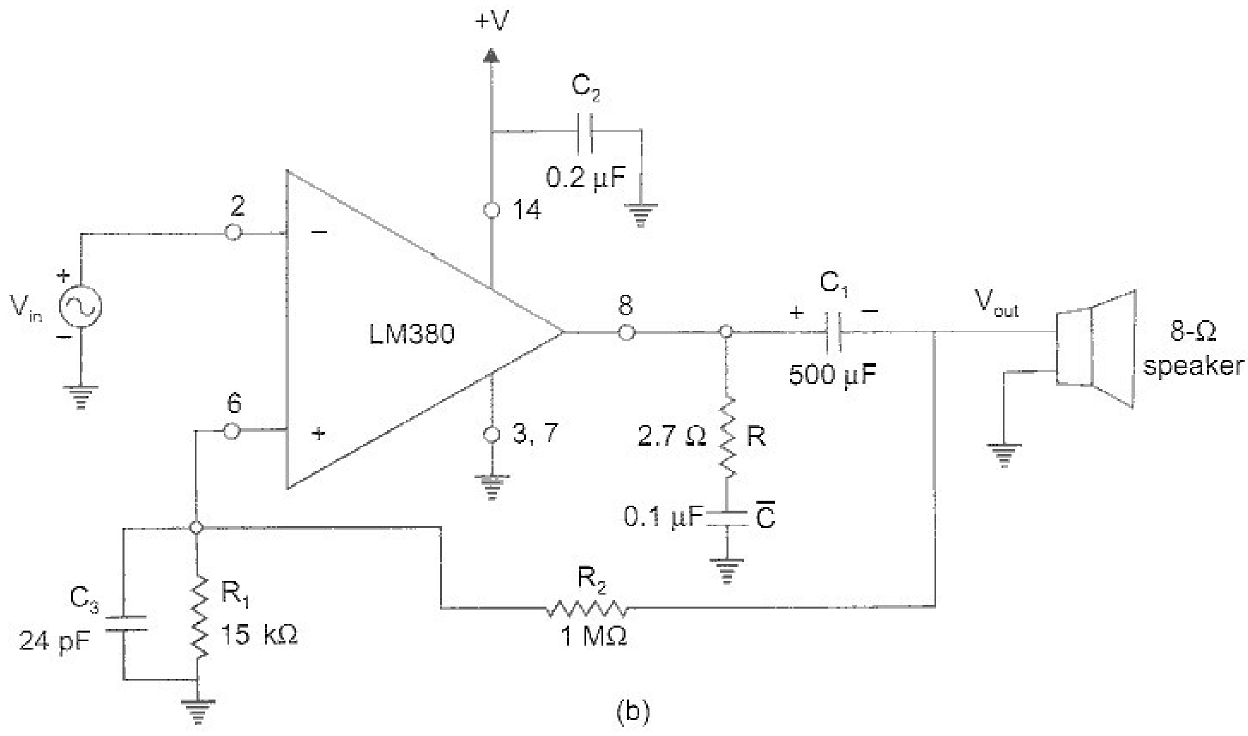


Fig. 4.47 (a) Audio power amplifier using LM380 (b) Use of positive feedback to increase gain