8

OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA)

8.1 INTRODUCTION

Op-amp discussed earlier is a very versatile linear *IC* as can be seen from its applications in the design of various active circuits. A major limitation of op-amp based circuits is that it is not possible to electronically tune the various circuit parameters such as gain, bandwidth, cut-off frequency etc. In this chapter, we introduce operational transconduction amplifier (OTA) which has external bias control terminal through which electronic tunability is possible. Thus, using OTAs, active circuits with greater flexibility can be designed.

The internal schematic circuit, biasing techniques and important applications of OTA have been discussed.

8.2 OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA)

The characteristics of an ideal OTA are very similar to those of an ideal op-amp except that an OTA has *very high output impedance*. The output signal, therefore is current which is proportional to the differential input voltage. Thus, an OTA is described in terms of transconductance (g_m) is place of voltage gain. The other important feature of OTA is that it has an extra control terminal through which its transconductance, g_m can be linearly controlled over several decades by varying the bias current (I_B).

OTAs are available in both bipolar and MOS versions. We shall here discuss the bipolar OTAs only. The bipolar OTAs are commercially available as single device LM3080, CA3080, or dual OTA on a chip (LM13600, CA3280) and triple OTA on a chip (CA3060).

The symbolic representation of OTA is shown in Fig. 8.1. The pin diagram of dual-in-line package and metal can package ICs are shown in Figs. 8.1 (b) and 8.1 (c) respectively. The ideal small signal model of OTA is shown in Fig. 8.1 (*d*).

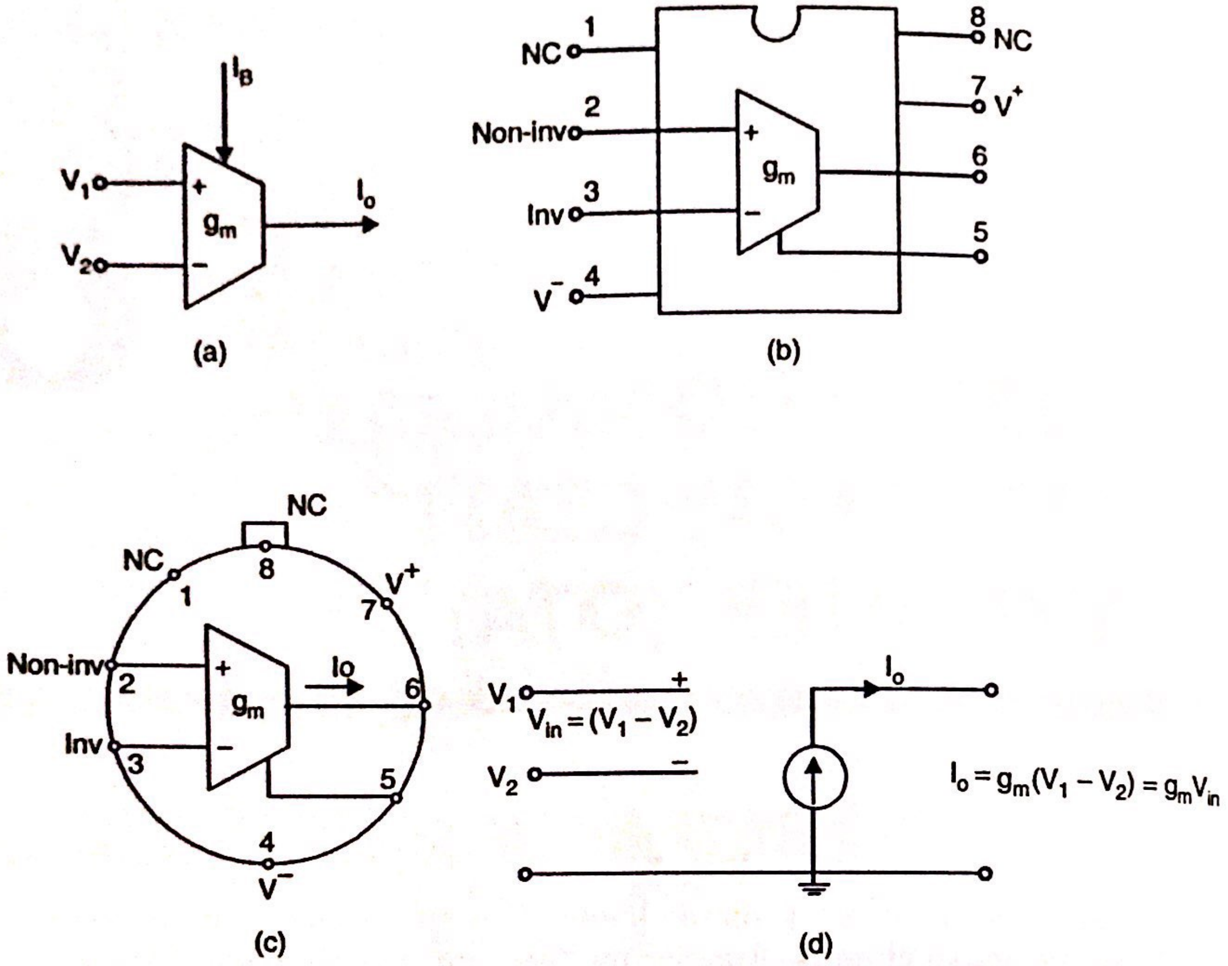


Fig. 8.1 (a) OTA symbol, Pin diagram of (b) dual-in line package (c) metal-can package and (d) ideal small signal model

The important features of an ideal OTA are as listed below :

(*i*) Input resistance $R_{in} = \infty$

(*ii*) Output resistance $R_o = \infty$

(*iii*) Bandwidth $= \infty$

(*iv*) Perfect balance, i.e. $I_o = 0$, when $V_1 = V_2$

(*v*) Trans conductance, g_m is finite and controllable with amplifier bias current I_B

The output current in Fig. 8.1 (*d*) is given by

$$I_o = g_m (V_1 - V_2) = g_m V_{in} \tag{8.1}$$

where, $I_o \rightarrow$ Output current

$g_m \rightarrow$ Transconductance

$V_1 \rightarrow$ Non-inverting input voltage

$V_2 \rightarrow$ Inverting input voltage

A commercially available OTA is, however non-ideal with finite values of input and output resistance, offset voltage and finite bandwidth. Typical parameters of OTA (CA3280A) are shown in Table 8.1.

8.2.2 OTA Applications

Basic Inverting Amplifier

A basic inverting amplifier using OTA is shown in Fig 8.6. The input voltage V_{in} is applied to the inverting terminal of OTA and non-inverting terminal is grounded. Assuming OTA to be ideal, the output current I_o is given by

$$I_o = -g_m V_{in} \tag{8.18}$$

Also,

$$V_o = I_o R_L \tag{8.19}$$

From Eqs. (8.18) & (8.19), we get

$$\frac{V_o}{V_{in}} = -g_m R_L \tag{8.20}$$

and output impedance, $$Z_o = \frac{V_o}{I_o} = R_L$$

Thus, the voltage gain is directly proportional to g_m and therefore, it can be varied by changing the bias control current I_B.

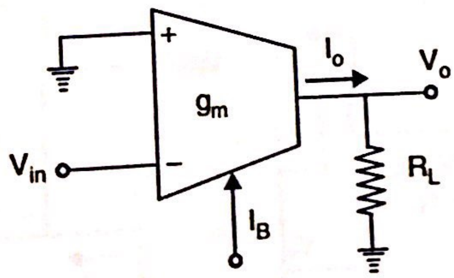


Fig 8.6 Basic inverting amplifier

Basic Non-Innerting Amplifier

Figure 8.7 shows the circuit of a basic non-inverting amplifier. The input voltage V_{in} is now applied to the non-inverting terminal and inverting terminal of OTA is grounded. The output current (I_o) of OTA is given as

$$I_o = g_m V_{in} \tag{8.21}$$

Also,

$$I_o = \frac{V_o}{R_L} \tag{8.22}$$

Scanned with CamScanner

From equation, (8.21) and (8.22), we get

$$\frac{V_o}{V_{in}} = g_m R_L \tag{8.23}$$

The voltage gain is positive and directly proportional to g_m and thus can be controlled by the bias control current. The output impedance, Z_o is R_L as can be seen.

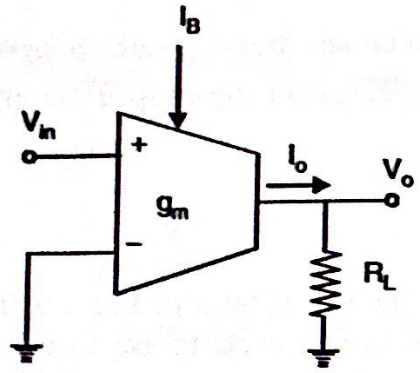


Fig. 8.7 Basic non-inverting amplifier

One basic limitation of two circuits discussed is that the output impedance is equal to load resistance R_L which is usually high to obtain high gain. Since OTA is a current source, its output impedance is very high in contrast to the op-amps having very low output impedance. We know that low output impedance is a desirable feature in amplifiers used to drive resistive loads. To overcome this, OTA feedback amplifiers called Buffered Amplifiers are used.

Buffered Amplifiers

The commercial OTA's such as National semiconductors LM13600, therefore, provide on-chip controlled impedance buffers. Figure 8.8 shows an inverting amplifier using OTA which can provide not only controllable gain, but also uses negative feedback to reduce the output resistance.

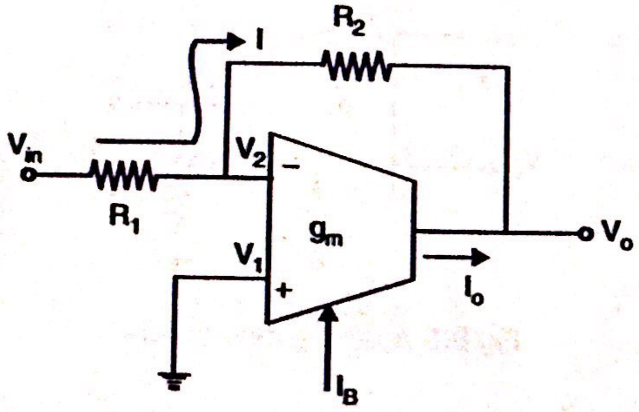


Fig. 8.8 Buffered inverting amplifier

The expression of the voltage gain may be obtained as follows:
From the basic behaviour of the OTA, we may write

$$\begin{aligned} I_o &= g_m(V_1 - V_2) \\ &= -g_m V_2 \quad [\text{as } V_1 = 0] \end{aligned} \tag{8.24}$$

$$\frac{V_{in}-V_2}{R_1} = \frac{V_2-V_o}{R_2} = I \tag{8.25}$$

and

$$I_o = -I$$

therefore,

$$g_m V_2 = \frac{V_2-V_o}{R_2}$$

$$V_2 = \frac{V_o}{1-g_m R_2} \tag{8.26}$$

Putting the value of V_2 from Eq. (8.26) in Eq. (8.25) and simplifying, we get

$$\frac{V_o}{V_{in}} = \frac{1-g_m R_2}{1+g_m R_1} \tag{8.27}$$

and the output impedance is found to be

$$R_o = \frac{R_1+R_2}{1+g_m R_1} \tag{8.28}$$

It is evident from Eq. (8.27) that voltage gain can be continuously adjusted between positive and negative values with the parameter g_m.

Similarly, one can make a buffered non-inverting amplifier by applying input voltage V_{in} to non-inverting terminal and grounding the resistance R_1. It can be shown that its voltage gain and output impedance and given by

$$\frac{V_o}{V_{in}} = \frac{g_m(R_1+R_2)}{1+g_m R_1} \tag{8.29}$$

and

$$R_o = \frac{R_1+R_2}{1+g_m R_1} \tag{8.30}$$

OTA Amplifier without Feedback

An all-OTA inverting amplifier without negative feedback is shown in Fig. 8.9.

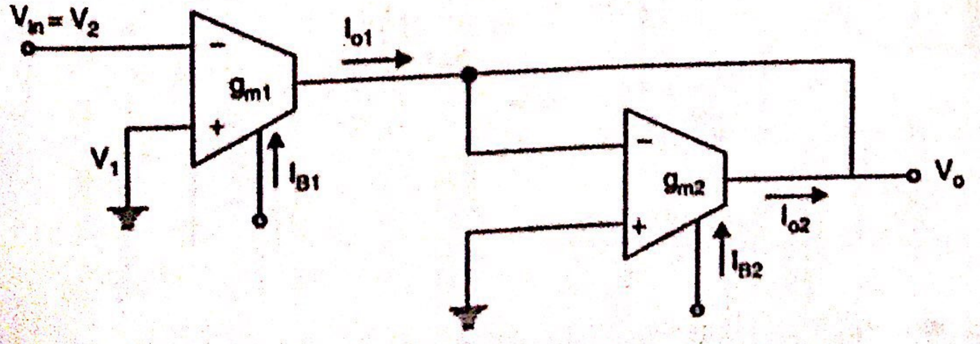


Fig. 8.9 An all-OTA amplifier without feedback

The amplifier circuit uses two OTA's with transconductance g_{m1} and g_{m2} respectively. We may write

$$I_{o1} = g_{m1}(V_1 - V_2) \quad (8.31)$$

$$= g_{m1}(0 - V_{in}) = -g_{m1} V_{in} \quad (8.32)$$

Similarly,

$$I_{o2} = g_{m2}(0 - V_o) \quad (8.33)$$

$$= -g_{m2} V_o \quad (8.34)$$

Since

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

Therefore,

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

or

$$\frac{V_o}{V_{in}} = \frac{-g_{m1}}{g_{m2}} \quad (8.35)$$

The voltage gain, therefore, is completely controllable by the external control currents I_{B1} and I_{B2} which control g_{m1} and g_{m2} respectively.

OTA based Grounded Resistor

An OTA based positive and negative grounded resistor is shown in Fig. 8.10 (a) and (b) respectively.

In Fig. 8.10 (*a*), the output current (I_o) of the OTA is

$$I_o = -g_m V_{in} \quad (8.36)$$

Also

$$I_o = -I_{in} \quad (8.37)$$

From Eqs. (8.36) & (8.37), we get

$$\frac{V_{in}}{I_{in}} = \frac{1}{g_m}$$

Thus, the circuit simulates ideal ground resistor of value

$$Z_{in} = R = \frac{1}{g_m}$$

or

$$R = \frac{2V_T}{I_B} \quad (8.38)$$

The value of the resistor R can be tuned over several dacades with bias current I_B.

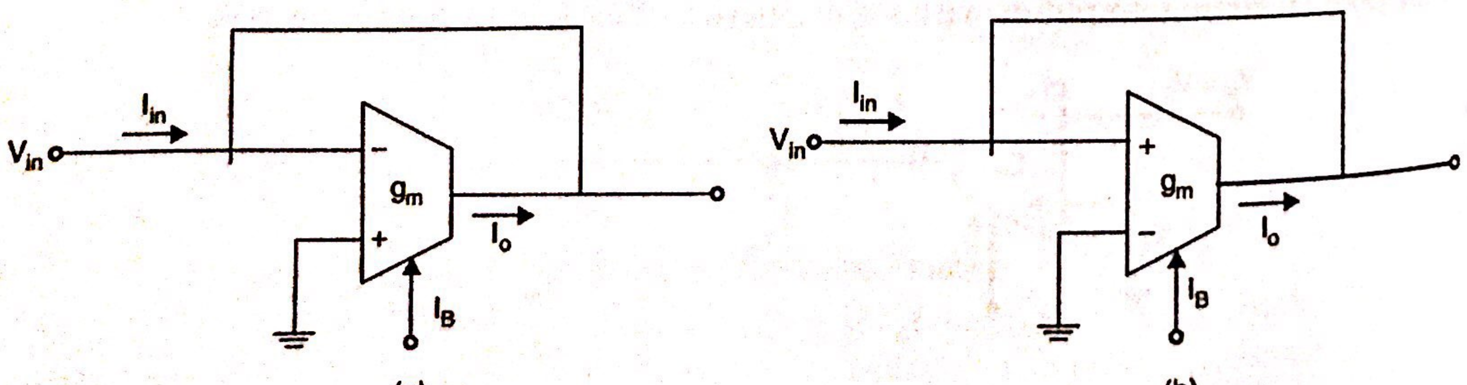


Fig. 8.10 (a) Ideal positive grounded resistor, (b) Ideal negative grounded resistor

Summing Amplifier

The circuit of an all OTA summing amplifier is shown in Fig. 8.11. It consists of three OTAs with transconduces g_{m1}, g_{m2} & g_{m3}. The outputs of OTA_1 and OTA_2 are connected to produce a summing current I and the OTA_3 is connected as a positive resistor. The output current of OTA_1 is

$$I_{o1} = g_{m1} V_1 \tag{8.39}$$

The output current of OTA_2 is

$$I_{o2} = g_{m2} V_2 \tag{8.40}$$

Further,

$$I = I_{o1} + I_{o2} \tag{8.41}$$

and output of OTA_3 is

$$I_{o3} = -g_{m3} V_3 \tag{8.42}$$

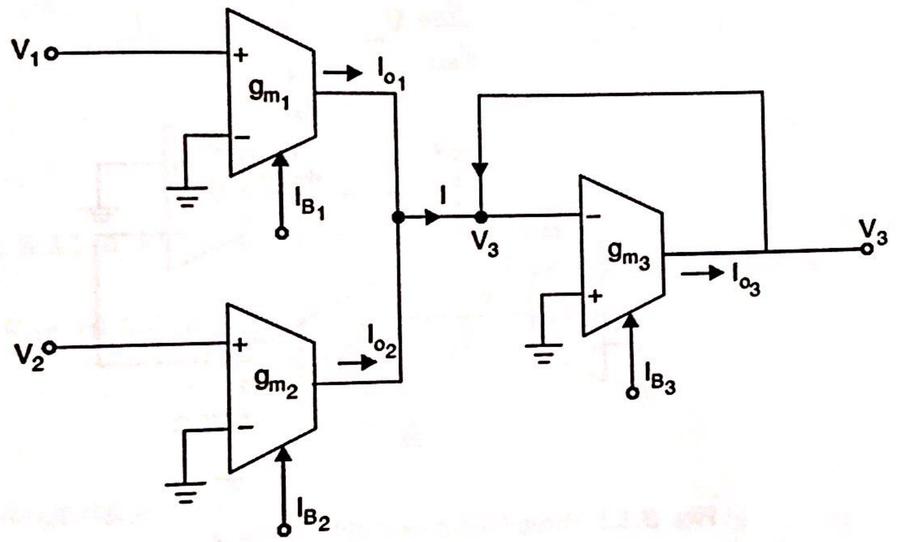


Fig. 8.11 Summing amplifier

KCL at V_3 gives

$$I = -I_{o3} = -(-g_{m3} V_3)$$

or

$$I = g_{m3} V_3 \tag{8.43}$$

Using equations (8.39), (8.40) and (8.41), we get

$$I = g_{m1} V_1 + g_{m2} V_2 \tag{8.44}$$

or

$$g_{m3} V_3 = g_{m_1} V_1 + g_{m_2} V_2$$

i.e.

$$V_3 = \frac{g_{m_1} V_1}{g_{m3}} + \frac{g_{m_2} V_2}{g_{m3}} \tag{8.45}$$

Thus, it is evident that the output voltage V_3 is the sum of two scaled voltages, and it can be controlled by g_{m_1}, g_{m_2} or g_{m_3}. This circuit can also be extended to more than two signals. Further, by changing the input terminal of any feed in OTA one can change the sign of the corresponding summing coefficient.

A Three OTA-based Differentiator

The circuit for a differentiator using three OTAs is shown in Fig. 8.12. It consists of three OTAs and a capacitor.

We may write the following equations:

$$I_{o1} = -g_{m1} V_{in} \tag{8.46}$$

$$I_{o1} = -I_{o3} \tag{8.47}$$

Also

$$I_{o2} = g_{m2} V_o \tag{8.48}$$

$$= sV_2 C \tag{8.49}$$

$$I_{o3} = -g_{m_3} V_2 \tag{8.50}$$

Using Eqs. (8.46) to (8.50) and simplifying, we get

$$V_2 = \frac{g_{m2} V_o}{sC} = \frac{-g_{m1}}{g_{m3}} V_{in} \tag{8.51}$$

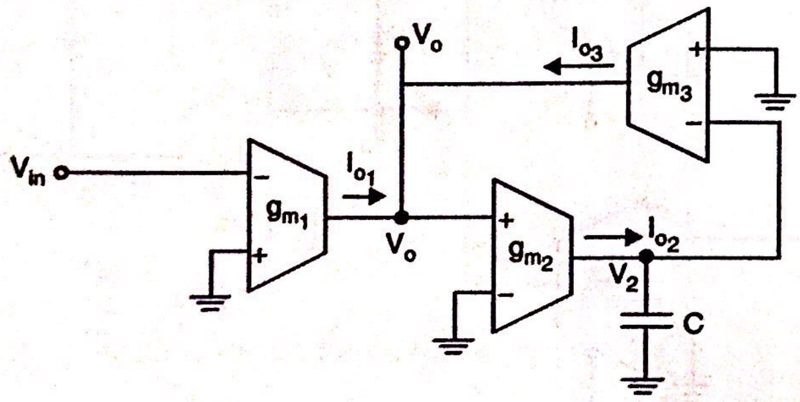


Fig. 8.12 Three OTA-based differentiator

Thus, the voltage gain of the differentiator is obtained as,

or

$$\frac{V_o}{V_{in}} = \frac{-sCg_{m1}}{g_{m2} g_{m3}} \tag{8.52}$$

From equation (8.52), it is seen that an ideal inverting differentiator has been realized. The voltage gain of the realized differentiator can be conveniently controlled with the bias current control of the OTAs, i.e., either by g_{m1}, g_{m2} or g_{m3}. Inverting and non-inverting differentiators can be obtained by connecting the inverting or non-inverting terminal of OTA_1 to ground respectively.

A Programmable Integrator

The circuit of a voltage variable integrator with a differential input in shown in Fig. 8.13 (*a*). It is also known as programmable integrator (PI). Its symbolic representation is shown in Fig. 8.13 (*b*). In the circuit of Fig. 8.13 (*a*), the OTA is loaded by a capacitor. Since the output impedance of OTA is ideally infinite, a very high input impedance buffer is used to avoid undesirable loading.

The output current (I_o) of OTA is given as

$$I_o = g_m V_{in} \tag{8.53}$$

Also

$$I_o = sV_oC \tag{8.54}$$

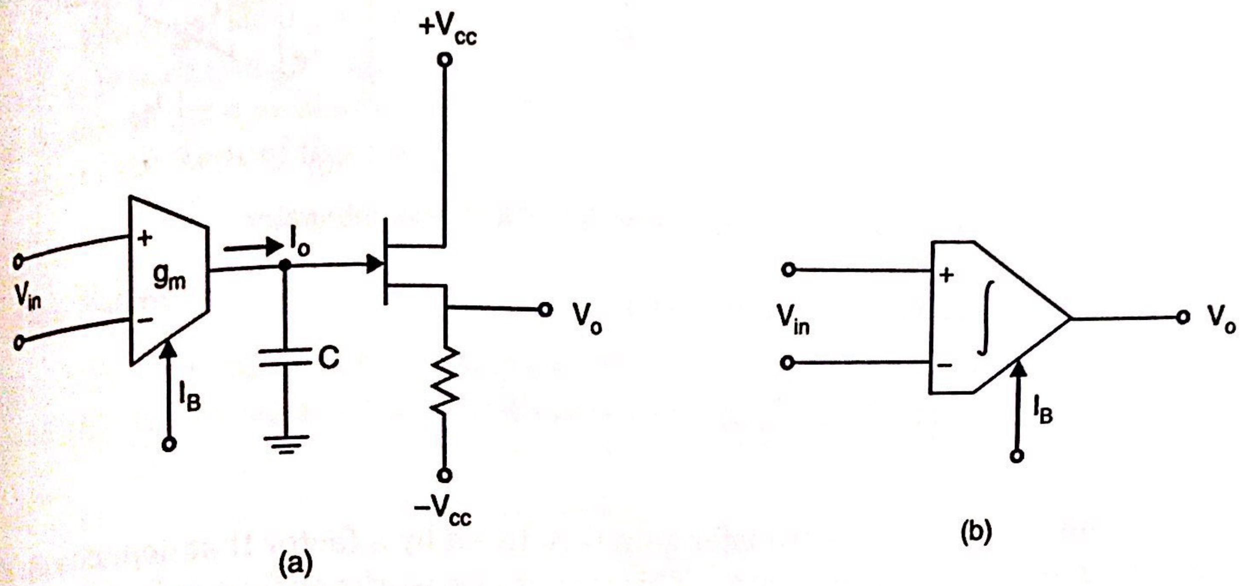


Fig. 8.13 (a) Programmable integrator, (b) Symbol of a programmable integrator

Simplifying, Eqs. (8.53) and (8.54), we get the tranfer gain as,

$$\frac{V_o}{V_{in}} = \frac{g_m}{sC} = \frac{K}{s} \tag{8.55}$$

where, the integration constant K is,

$$K = \frac{g_m}{C} = \frac{I_B}{2V_T C} \tag{8.56}$$

From Eq. (8.56), it is clear that the circuit realizes an ideal integrator and its gain is directly proportional to OTA's bias current I_B. Hence, gain can be controlled by varying the bias current I_B. Inverting and non-inverting integrator can be obtained by connecting inverting or non-inverting terminal of OTA to ground respectively.

Lossy or Practical Integrators

The ideal integrator discussed above has a limitation. At low frequencies or for dc, the gain becomes infinite and leads to saturation of transistors. Therefore, a resister R is connected across C to make it a practical or lossy integrator.

The circuit for a lossy integrator is shown in Fig. 8.14(a). The input is applied to the non-inverting terminal of OTA. The output current, I_o is given as

$$I_o = g_m V_{in} \tag{8.57(a)}$$

Also

$$I_o = V_o\,(I/R + sC) \tag{8.57(b)}$$

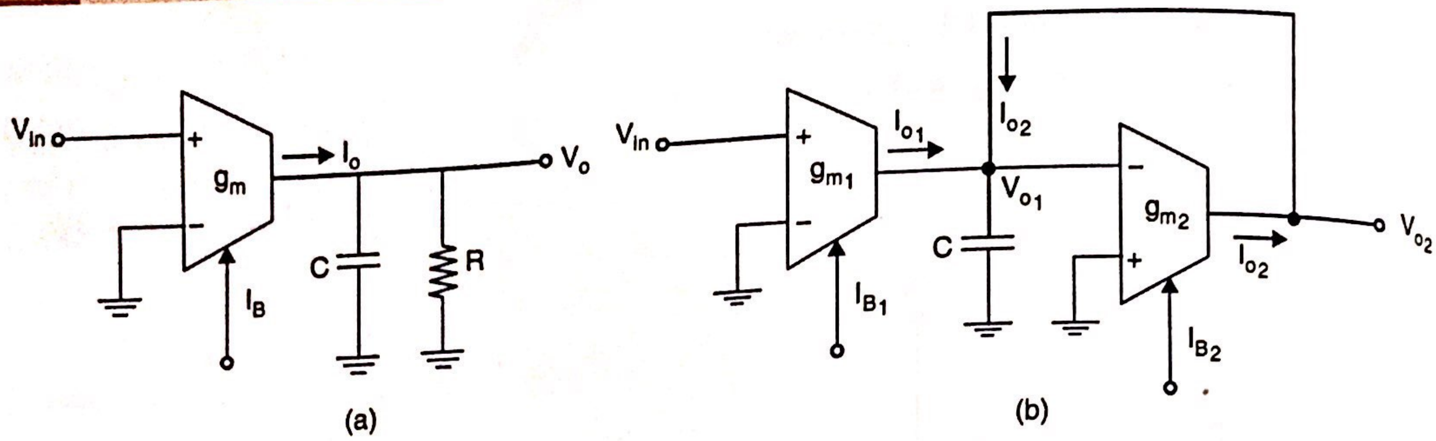


Fig. 8.14 (a) Lossy integrator, (b) OTA-C lossy integrator

Thus,
$$g_m V_{in} = V_o\ (1/R + sC) \tag{8.58}$$

or
$$\frac{V_o}{V_{in}} = \frac{g_m R}{1+sCR} \tag{8.59}$$

Equation (8.59) shows that the transfer gain is reduced by a factor that depends on the RC product and the gain can be adjusted by g_m. This circuit also works as first order low-pass filter. [Also, refer to section 4.11, Ch-4]

The resistance R in Fig. 8.14(a) can be easly replaced by an OTA-based simulated resistor as shown in Fig. 8.14 (*b*). The output current I_{o1} of first OTA is

$$I_{o1} = g_{m1}\ V_{in} \tag{8.60}$$

KCL at V_{o1} gives

$$I_{o1} = -\ I_{o2} \tag{8.61}$$

The current flowing through the capacitor is $I_{o1} + I_{o2}$ and is given as

$$I_{o1} + I_{o2} = sCV_{o1} \tag{8.62}$$

so
$$V_{o1} = \frac{I_{o1} + I_{o2}}{sC} \tag{8.63}$$

for OTA-2
$$I_{o2} = -\ g_{m2}\ V_{o1} \tag{8.64}$$

Also, since, $V_{o1} = V_{o2}$, we may write

$$V_{o2} = \frac{g_{m1} V_{in} - g_{m2} V_{o2}}{sC}$$

After simplification, the voltage gain of the integrator is given by

$$\frac{V_{o2}}{V_{in}} = \frac{g_{m1}}{sC + g_{m2}} \tag{8.65}$$

Equation (8.65) shows that the cut-off frequency can be controlled by g_{m2} and dc gain by g_{m1}.

8.3 ACTIVE FILTERS USING OTAs

OTA-based Tunable Filters

The use of OTA for making active filters has been found to be very attractive as it is possible to control the performance parameters of the filter by an external control voltage, V_c (or control current I_B). The controlled parameters are the midband gain of the circuit, or the 3-dB frequency of a filter. It is also possible to control the critical frequency of the filter without altering the passband gain or even to change the type of response from low pass to all pass to high pass by continuous adjustment of the transconductance, g_m. A few of the basic structures are developed here.

First-order Filters

The first order low pass (*LP*), high-pass (*HP*) and an all-pass (*AP*) filter section play an important role in the realization of odd-order and higher-order filters.

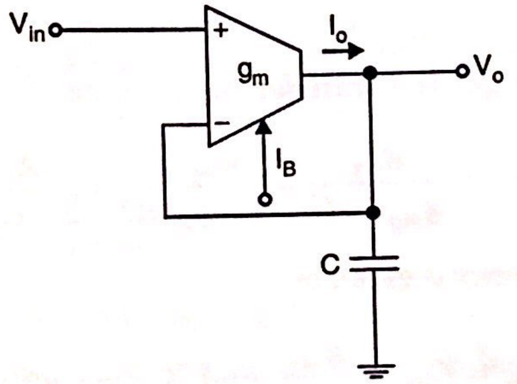


Fig. 8.15 OTA-C low-pass filter

Fig. 8.13 shows the circuit of a first order OTA-C low pass filter. We may write,

$$I_o = g_m (V_{in} - V_o) \tag{8.66}$$

Also,

$$I_o = sC\ V_o \tag{8.67}$$

Simplifying, we get

$$\frac{V_o}{V_{in}} = \frac{g_m}{g_m + sC} = \frac{\omega_h}{s + \omega_h} \tag{8.68}$$

where, the upper cut off frequency, $\omega_h = g_m/C$.

Another improved first order (with roll-off rate of 20 dB/decade) low pass filter using two OTAs is shown in Fig. 8.16. This filter provides independent tuning of gain and uper-cut off frequency.

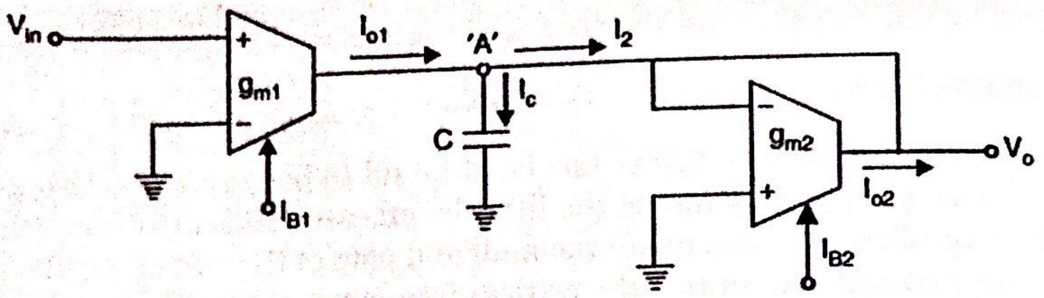


Fig. 8.16 First order low pass filter using two OTA

At node 'A',

$$I_{o1} = I_c + I_2$$

$$= I_c - I_{o2} \tag{8.69}$$

Also,

$$I_{o1} = g_{m1} V_{in} \tag{8.70}$$

$$I_c = sCV_o \tag{8.71}$$

and

$$I_{o2} = -g_{m2} V_o \tag{8.72}$$

Simplifying and solving, we get the transfer function as

$$\frac{V_o}{V_{in}} = \frac{g_{m_1}}{g_{m_2} + sC} = \frac{g_{m_1}/C}{s + g_{m_2}/C} = \frac{A}{s + \omega_h} \tag{8.73}$$

The upper cut-off 3-dB frequency is given by

$$f_h = f_{3dB} = \frac{g_{m2}}{2\pi C} \text{ and } A = g_{m1}/C \tag{8.74}$$

In this circuit, the second OTA has been configured as a voltage variable resistor. It is this variable resistor which provides the variable cut-off frequency in Eq. (8.74). This circuit thus realizes a first order filter having independent electronic tunibility of gain with I_{B1} and upper cut-off frequency with I_{B2}.

OTA-C High Pass Filter

The circuit for a first order high-pass filter is shown in Fig. 8.17. We may write,

$$I_o = -g_m V_o \tag{8.75}$$

$$I_i = (V_{in} - V_o)\, sC$$

and

$$I_i = -I_o$$

Solving and simplifying, we get the transfer function as,

$$T_{HP}(s) = \frac{s}{s + g_m/C} \tag{8.76}$$

where, the lower cut off frequency, $\omega_l = g_m/C$

Equation (8.77) shows that the circuit in Fig. (8.17) realizes a first order high pass filter with a constant gain of 1 and electronically tunable lower cut-off frequency by varying I_B.

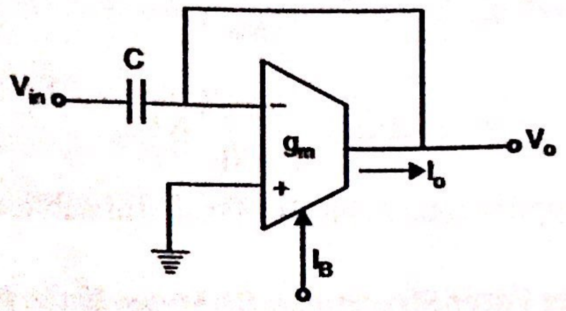


Fig. 8.17 First order high pass filter

OTA-C All Pass Filter

The circuit in Fig. 8.18(a) uses one OTA and two equal valued capacitors. The transfer function is given by

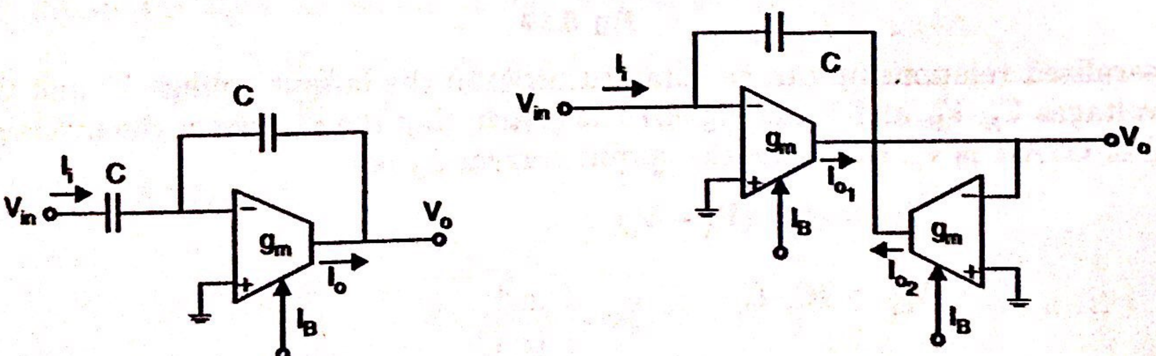
$$T_{AP}(s) = \frac{s - g_m/C}{s + g_m/C} \tag{8.77}$$

i.e.

$$\frac{V_o}{V_i} = \frac{s - \omega_o}{s + \omega_o} \tag{8.78}$$

where $A = 1$ and $\omega_o = g_m/C$

Thus, the circuit provides unity gain and phase shift can be varied by I_B.



(b)

(a)

Fig. 8.18 All pass filter (a) using two capacitor (b) one capacitor

Another circuit for all-pass filter is shown in Fig. 8.18(b) which uses two matched OTAs and one capacitor. The transfer function is found as

$$T_{AP}(s) = \frac{s - g_m/C}{s + g_m/C} \tag{8.79}$$

where $A = 1$

and $\omega_o = g_m/C$

Thus both the circuits provide unity gain with phase relationship given by :

$$\phi = \pi - 2\tan^{-1}\left(\frac{\omega C}{g_m}\right) \tag{8.80}$$

$$= \pi - 2\tan^{-1}\left(\frac{2\omega C V_T}{I_B}\right)$$

These circuits find wide applications in electronically turnable two quadrant phase shifters.

A Generalised Second-order Filter Structure with Independent Turing of ω_0 and Q

Bi-quadratic filter structures find wide applications in the design of higher-order filters. Here, we discuss a generalised second order filter which can realize low pass, high pass, band pass and a notch filter. The circuit shown in Fig. 8.19 uses three OTA's, two capacitors and three voltage control terminals V_A, V_B and V_C.

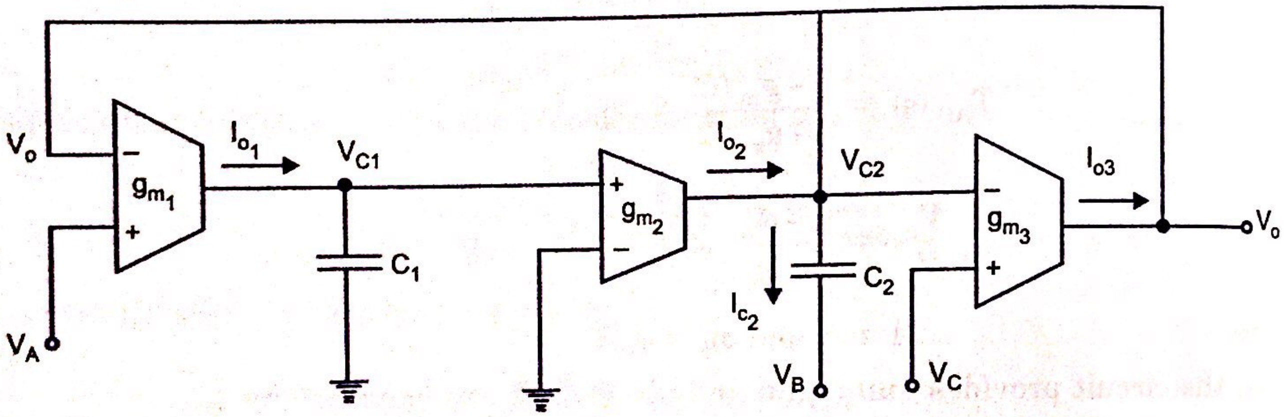


Fig. 8.19

A generalised relationship can be obtained between the output voltage V_o and the three control voltages V_A, V_B, and V_C, in Fig. 8.19. It is seen that the voltage at the inverting input terminal of OTA-1 is V_o, therefore, the output current I_{o1} is

$$I_{o1} = g_{m1}\ (V_A - V_o) \tag{8.81}$$

and,

$$V_{C_1} = sC_1\ I_{o_1} \tag{8.82}$$

$$= g_{m1}\left(\frac{V_A - V_o}{sC_1}\right) \tag{8.83}$$

The output current I_{o2} of OTA-2 is

$$I_{o2} = g_{m_2} V_{C_1} = g_{m2} g_{m1} \frac{(V_A - V_o)}{sC_1} \tag{8.84}$$

and

$$V_{C_2} = V_o = sC_2\ I_{o2} \tag{8.85}$$

For OTA-3 $$I_{o3} = g_{m_3}(V_C - V_o) \quad \left[\text{as } V_{C_2} = V_o\right] \tag{8.86}$$

Further, at the output of OTA-2, using kirchhoff's current law, we may write,

$$\begin{aligned} I_{o2} + I_{o3} &= I_{c_2} \\ &= (V_o - V_B)sC_2 \end{aligned} \tag{8.87}$$

Collecting all terms in V_o together, we get,

$$V_o\left[\frac{s^2C_1C_2 + sC_1g_{m_3} + g_{m_1}g_{m_2}}{sC_1}\right] = sC_2V_B + g_{m_3}V_C + \frac{g_{m_1}g_{m_2}V_A}{sC_1} \tag{8.88}$$

Solving for V_o gives a generalised expression,

$$V_o = \frac{s^2C_1C_2V_B + sC_1g_{m3}V_C + g_{m_1}g_{m_2}V_A}{s^2C_1C_2 + sC_1g_{m_3} + g_{m_1}g_{m_2}} \tag{8.89}$$

Low Pass Filter

A second order low pass filter can be obtained by applying input signal to V_A and grounding V_B and V_C. i.e. making $V_B = V_C = 0$ in Eq. (8.89), we obtain,

$$\frac{V_o}{V_A} = \frac{g_{m1}\,g_{m2}}{s^2C_1C_2 + sC_1g_{m3} + g_{m1}g_{m2}} \tag{8.90}$$

Thus, we obtain, the standard second order LP filter as

$$\frac{V_{LP}}{V_A} = \frac{g_{m1}g_{m2}/C_1C_2}{s^2 + \dfrac{sg_{m3}}{C_2} + \dfrac{g_{m1}\,g_{m2}}{C_1C_2}} \tag{8.91(a)}$$

or $$\frac{V_{LP}}{V_A} = \frac{\omega_h^2}{s^2 + s\left(\dfrac{\omega_h}{Q}\right) + \omega_h^2} \tag{8.91(b)}$$

where, $$\omega_h = \sqrt{\frac{g_{m1}\,g_{m2}}{C_1C_2}} \tag{8.92 (a)}$$

and $$Q = \frac{1}{g_{m3}}\sqrt{\frac{g_{m1}\,g_{m2}\,C_2}{C_1}} \tag{8.92 (b)}$$

High Pass Filter

A high pass filter can be obtained by setting V_A and V_C to ground and applying input signal to V_B. The expression for high pass filter is obtained as

$$\frac{V_o}{V_B} = \frac{V_{HP}}{V_B} = \frac{s^2}{s^2 + s\frac{g_m}{C_2} + \frac{g_m^2}{C_1C_2}} \quad \text{[Assuming } g_{m1} = g_{m2} = g_{m3} = g_m\text{]} \tag{8.93}$$

One can make a band pass filter by setting V_A and V_B to ground and applying input voltage to V_C. The transfer function then becomes,

$$\frac{V_o}{V_C} = \frac{V_{BP}}{V_C} = \frac{sC_1g_{m3}}{s^2C_1C_2 + sC_1g_{m3} + g_{m1}g_{m2}} \tag{8.94}$$

Dividing both numerator and denominator by C_1C_2, gives

$$\frac{V_{BP}}{V_C} = \frac{s(g_{m3}/C_2)}{s^2 + s\left(\frac{g_{m3}}{C_2}\right) + \frac{g_{m1}g_{m2}}{C_1C_2}} \tag{8.95}$$

This transfer function is of the form

$$\frac{V_o}{V_{in}} = \frac{A_1s}{s^2 + s\left(\frac{\omega_o}{Q}\right) + \omega_o^2} \tag{8.96}$$

which is the standard second-order band pass filter.

The center frequency of the filter is

$$\omega_o = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} \tag{8.97}$$

and 3-dB bandwidth is found directly from the transfer function to be

$$BW = \frac{\omega_o}{Q} = \frac{g_{m3}}{C_2} \text{ (radians/sec)} \tag{8.98}$$

Notch Filter

For notch filter, make $V_A = V_B = V_{in}$, the input signal and $V_C = 0$, then the transfer function of notch filter is given by:

$$\frac{V_o}{V_{in}} = \frac{V_{NF}}{V_{in}} = \frac{s^2C_1C_2 + g_{m1}g_{m2}}{s^2C_1C_2 + sC_1g_{m2} + g_{m1}g_{m2}} \tag{8.99}$$

The expressions for notch frequency ω_o and Q is given by

$$\omega_o = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} \tag{8.100(a)}$$

and

$$Q = \frac{1}{g_{m3}}\sqrt{\frac{g_{m1}g_{m2}C_2}{C_1}} \tag{8.100(b)}$$