CHAPTER –II

OPERATIONAL TRANSCONDUCTANCE AMPLIFIERS AND THEIR USES IN SIGNAL PROCESSING

2.1 INTRODUCTION

Operational transconductance amplifiers (OTA) have been attracting considerable attention in the literature in the context of various signal processing and signal generation applications. The main reason is the programmability of the transconductance gain by varying the bias current. Another factor responsible for its increasing popularity is the availability of current output instead of voltage output. Current output makes this active building block very attractive from the point of view of current mode circuits. The circuit of OTA being very simple and symmetrical is apt for integration. The major disadvantage being very limited linear range of the input differential voltage for commercially available bipolar OTA (CA 3080 type).

Many of the basic OTA based structures use only OTAs and capacitors and hence, are attractive for integration. Component count of these structures is often very low (e.g., second-order bi-quadratic filters can be constructed with two OTAs and two capacitors) when compared to VCVS designs. Convenient internal or external voltage or current control of filter characteristics is attainable with these designs. They are attractive for frequency referenced (e.g., master/slave) applications.

From a practical viewpoint, the high-frequency performance of discrete bipolar OTAs, such as the CA 3080, is quite good. The transconductance gain, g_m, can be varied over several decades by adjusting an external dc bias current, I_{ABC}. The major limitation of existing OTAs is the restricted differential input voltage swing required to maintain linearity. For the CA 3080, it is limited to about 30 mV p-p to maintain a reasonable degree of linearity. Although feedback structures in which the sensitivity of the filter parameters are reduced (as is the goal in op amp based filter design) will be discussed, major emphasis will be placed upon those structures in which the standard filter parameters of interest are directly proportional to g_m of the OTA. Thus, the g_m will be a design parameter much as are resistors and capacitors. Since the transconductance gain of the OTA is assumed proportional to an external dc bias current, external control of the filter parameters via the bias current can be obtained.

The present chapter is based on an excellent tutorial review[1]. In this review, starting from basic OTA model, the fundamental signal processing blocks have been presented. Some first and second order filter structures have also been reviewed.

2.2 OTA MODEL

The symbol used for the OTA is shown in Fig.2.1, along with the ideal small signal equivalent circuit. The transconductance gain, g_m, is assumed proportional to I_{ABC}. The proportionality constant h is dependent upon temperature, device geometry, and the process [2].

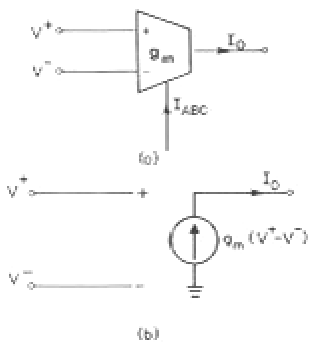


Fig. 2.1 OTA. (a) Symbol. (b) Equivalent circuit of ideal OTA.

A linear dependence on bias current is typically obtained for bipolar OTAs and MOS configurations operating in weak inversion. MOS structures operating in the saturation region typically exhibit a quadratic dependence on I_{ABC}.

$$g_m = h\, I_{ABC} \tag{2.2.1}$$

The output current is given by

$$I_o = g_m(V_+ - V_-) \tag{2.2.2}$$

[1] R.L.Geiger and E.Sánchez-Sinencio, "Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial," IEEE Circuits and Devices Magazine, Vol. 1, pp.20-32, March 1985.

As shown in the model, the input and output impedances in the model assume ideal values of infinity. Current control of the transconductance gain can be directly obtained with control of I_{ABC}. Since techniques abound for creating a current proportional to a given voltage, voltage control of the OTA gain can also be attained through the I_{ABC} input. Throughout this paper, when reference is made to either the current or voltage controllability of OTA based circuits it is assumed to be attained via control of g_m by I_{ABC}.

2.3 BASIC OTA BUILDING BLOCKS

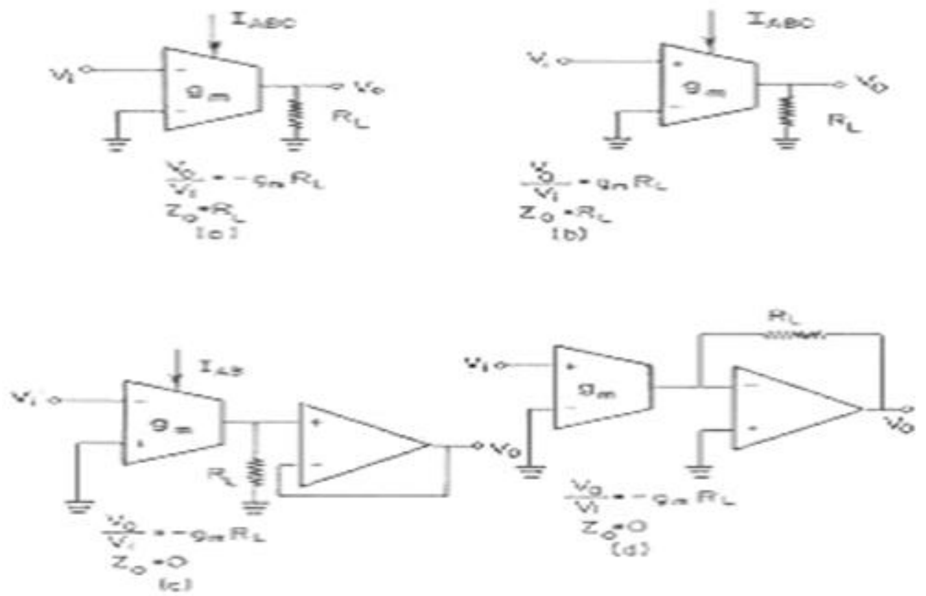
Some of the basic OTA building blocks [6] are introduced in this section. A brief discussion about these circuits follows.

Voltage amplifiers using OTAs are shown in Fig. 2.2, along with voltage gain and output impedance expressions. The basic inverting and non-inverting configurations of Figs. 2.2(a) and 2.2(b) have a voltage gain directly proportional to g_m, which makes current (voltage) control of the gain via I_{ABC} straightforward. Furthermore, observe that a differential amplifier can be easily obtained by using both input terminals of the OTA in Figs. 2.2(a) or 2.2(b). The major limitation of these circuits is the relatively high output impedance.

A voltage buffer, such as used in Figs. 2.2(c) and 2.2(d), is often useful for reducing output impedance. Although the gain characteristics of these circuits are ideally identical, the performance of the two circuits is not the same. The performance differences are due to differences in the effects of parasitic in the circuits. Specifically, the parasitic output capacitance of the OTA in Fig. 2.2(c), along with instrumentation parasitic, parallel the resistor *RL* in discrete component structures, thus is causing a roll-off in the frequency response of the circuits. In the circuit of Fig. 2.2(d), the parasitic output capacitance of the OTA is connected across the null port of an op amp and thus has negligible effects when the op amp functions properly. Likewise, instrumentation parasitic will typically appear at the low impedance output of the op amp, and thus not have a major effect on the performance. As with conventional amplifier design using resistors and op amp's, the amplifier bandwidth of these structures warrants consideration. For the circuits of Figs. 2.2(c) and 2.2(d), the major factor limiting the bandwidth is generally the finite gain bandwidth product of the op amps. If the op amps are modelled by the popular single-pole roll-off model, $A(s) = GB/s$, and the OTAs are assumed ideal, it follows that the bandwidth of the circuits of Fig. 2.2(c) and Fig. 2.2(d) is *GB*, independent of the voltage gain of the amplifier. This can be contrasted to the bandwidths of *GB/K* and *GB/*1 + *K* for the basic single op amp non-inverting and inverting amplifiers of gains *K* and - *K*, respectively.

Note that the circuits of Figs. 2.2(a) and 2.2(b) differ only in the labelling of the "+" and "-" terminals. In all circuits presented in this paper, interchanging the "+" and "-" terminals of the OTA will result only in changing the sign of the g_m coefficient in any equation derived for the original circuit. Henceforth, it will be the reader's responsibility to determine when such an interchange provides a useful circuit.

The circuits of Figs. 2.2(e) and 2.2(f) are feedback structures. The circuit of Fig. 2.2(e) offers gains that can be continuously adjusted between positive and negative values with the parameter g_m. By interchanging the + and - terminals of the OTA, very large gains can be obtained as g_m R_1 approaches 1 (as *Zo* approaches infinity). Gain is nonlinearly related to g_m. Control range via *gm* is reduced in these structures when compared to the amplifiers of Figs. 2.2(a) and 2.2(b). If components are sized fitly, the gain of these structures can be made essentially independent of g_m (as in the conventional op amp inverting and non-inverting configurations) and the output impedances can be made reasonably small.



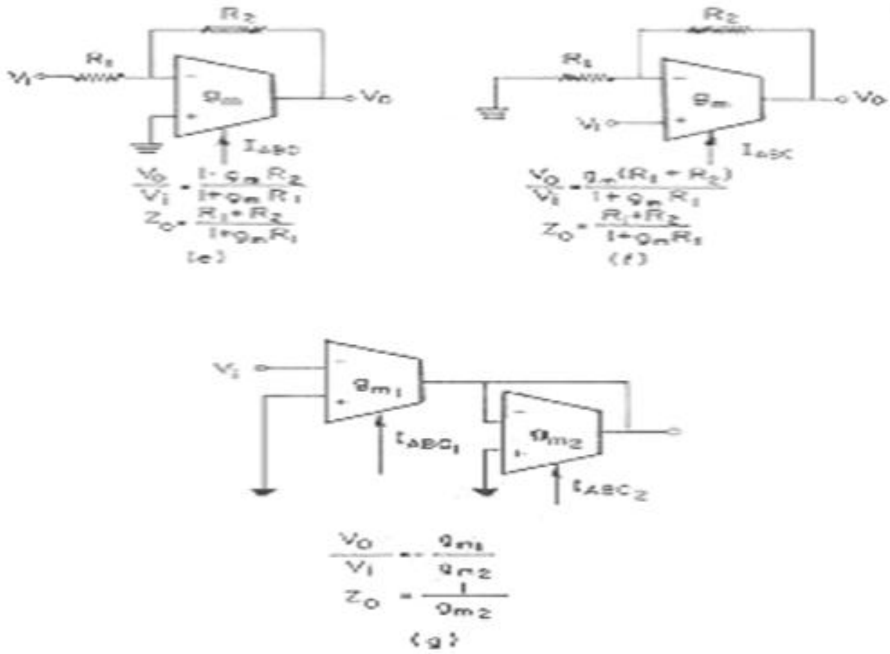


Fig. 2.2 Voltage amplifiers. (a) Basic inverting. (b) Basic non-inverting. (c) Feedback amplifier. (d) Non-inverting feedback amplifier. (e) Buffered amplifier. (f) Buffered VCVC feedback. (g) All OTA amplifiers [6].

The amplifier of Fig. 2.2(g) is attractive since it contains no passive components. Gain adjustment can be attained with either g_{m1}or g_{m2}. The total adjustment range of the gain of this structure is double (in dB) that attainable with the single OTA structures considered in Figs. 2.2(a) and 2.2(b). Furthermore, if both OTAs are in the same chip, the variations with temperature of the g_m's are cancelled.

Several standard controlled impedance elements are shown in Fig. 2.3, along with the input impedance expression. These controlled impedances can be used in place of passive counterparts (when applicable) in active *RC* structures to attain voltage control of the filter characteristics or as building blocks in OTA structures.

The circuit of Fig. 2.3(a) is a grounded Voltage Variable Resistor (VVR). The circuit of Fig. 2.3(b) behaves as a floating VVR, provided $g_{m1}$and g_{m2} are matched. If a mismatch occurs, the structure can be modelled with a floating VVR between terminals 1 and 2 of value g_{m1}, along with a voltage dependent current source of value $(g_{m1} - g_{m2})$ $V1$ driving node 2.

The circuit of Fig. 2.3(c) acts as a scaled VVR. Higher impedances are possible than with the simple structure of Fig. 2.3(a), at the expense of the additional resistors.

A voltage variable impedance inverter is shown in Fig. 2.3(d). Note the doubling of the adjustment range of this circuit, as with the amplifier of Fig. 2.2(g). Of special interest is the case where this circuit is loaded with a capacitor. In this case, a synthetic inductor is obtained. The doubling of the adjustment range is particularly attractive for the synthetic inductor since cut-off frequencies in active filter structures generally involve inductor values raised to the 1/2 power. By making $g_{m1} = g_{m2}$ and adjusting both simultaneously, first-order rather than quadratic control of cut-off frequencies is possible.

A floating impedance inverter is shown in Fig. 2.3(e). Note that it is necessary to match g_{m2} and g_{m3} for proper operation. The circuit of Fig. 2.3(f) serves as an impedance multiplier. That of Fig. 2.3(g) behaves as a super inductor and that of Fig. 2.3(h) as a FDNR.

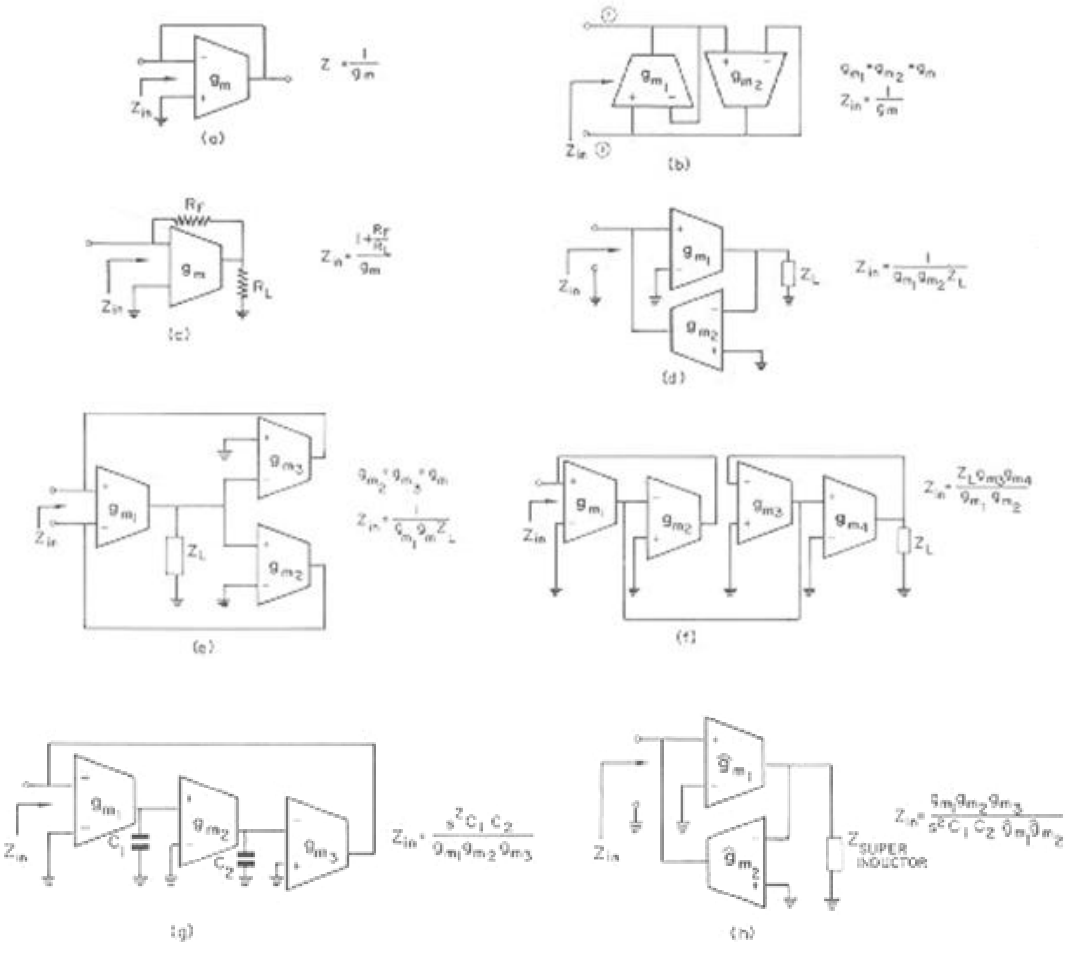


Fig. 2.3 Controlled impedance elements. (a) Single-ended voltage variable resistor (VVR). (b) Floating VVR. (c) Scaled VVR. (d) Voltage variable impedance inverter. (e) Voltage variable floating impedance. (f) Impedance multiplier. (g) Super inductor. (h) FDNR [6].

2.4 FIRST-ORDER FILTER STRUCTURES

A voltage variable integrator structure with a differential input is shown in Fig. 2.4(a). The integrator serves as the basic building block in many filter structures. Two different lossy integrators (first-order lowpass filters) are shown in Figs. 2.4(b) and 2.4(c).

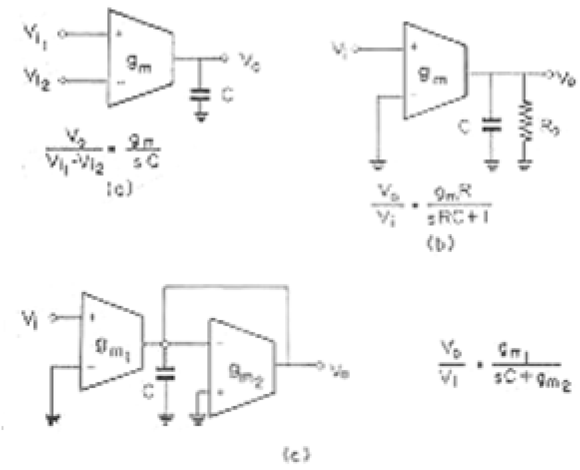


Figure 2.4. Integrator structures. (a) Simple. (b) Lossy. (c) Adjustable.

The circuit of Fig. 2.4(b) has a loss that is fixed by the *RC* product and a gain controllable by g_m. The circuit of Fig. 2.4(c) offers considerably more flexibility. The pole frequency can be adjusted by g_{m2} (interchanging the input terminals of OTA 2 actually allows the pole to enter the right half plane), and the dc gain can be subsequently adjusted by g_{m1}. It should be noted that the structure of Fig. 2.4(c) contains no resistors and can be obtained from the circuit of Fig. 2.4(b) by replacing the resistor R with the controlled impedance of Fig. 2.3(a). Another lossy integrator without adjustable gain but with adjustable pole location and a very simple structure is shown in Fig. 2.5(a).

When designing cascaded integrator-based filter structures, it may be the case that the input impedance to some stages is not infinite. If that be the case, a unity gain buffer would be required for coupling, since the output impedances of all integrators in Fig. 2.4 are nonzero. Note, however, that no buffer is needed for the cascade of any of the integrators of Fig. 2.4, since the input impedance to each circuit is ideally infinite.

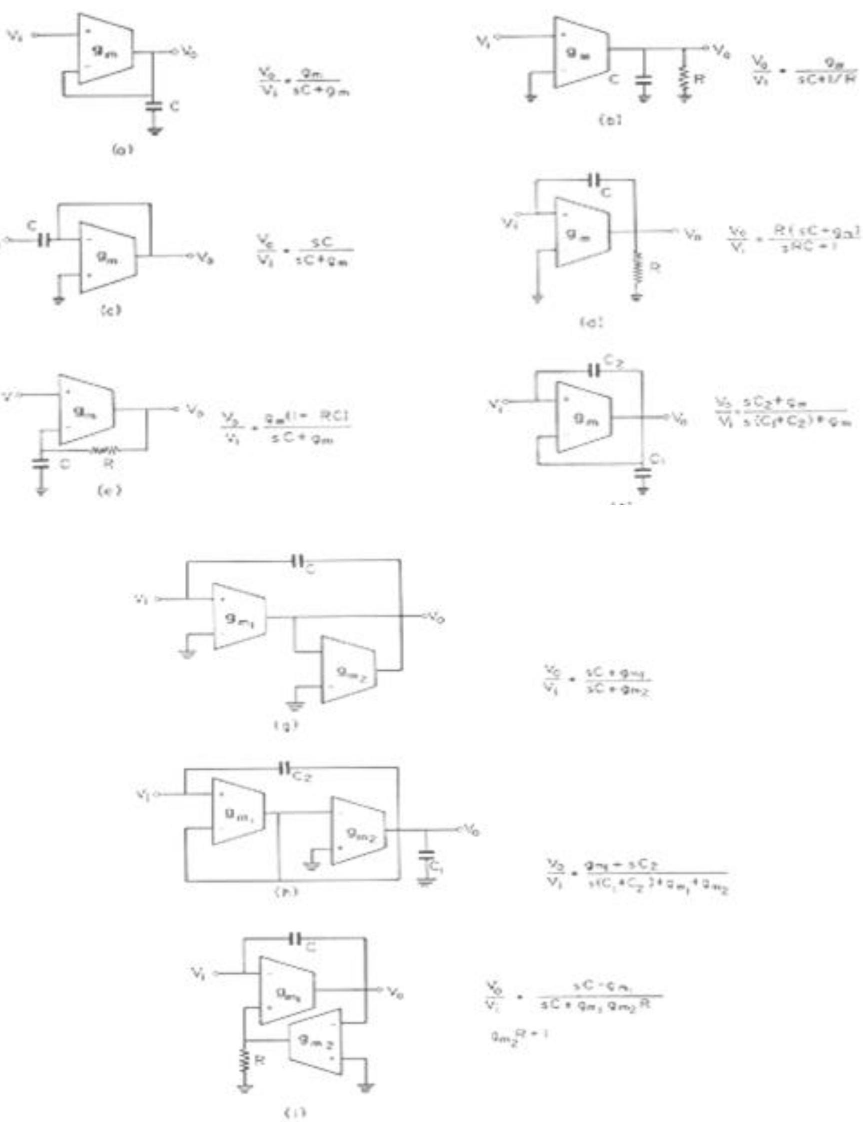


Fig. 2.5 First-order voltage-controlled filters. (a) Low-pass, fixed dc gain pole adjustable. (b) Low-pass fixed pole, adjustable dc gain. (c) High-pass, fixed high-frequency gain, adjustable pole. (d) Shelving equalizer, fixed high-frequency gain, fixed pole, adjustable zero. (e) Shelving equalizer, fixed high-frequency gain, fixed zero, adjustable pole. (f) Low-pass filter adjustable pole and zero, fixed ration. (g) Shelving equalizer, independently adjustable pole and zero. (h) Low-pass or high-pass filter, adjustable zero and pole, fixed ratio or independent adjustment. (i) Phase shifter, adjustable with gm.

First-order filters can be readily built using OTAs. Considerable flexibility in controlling those specific filter characteristics that are usually of interest is possible with these structures. Several first-order voltage-controlled filters are shown in Fig. 2.5, and a functional plot of the transfer characteristics as a function of the transconductance gains is shown in Fig. 2.6.

The 3dB cut-off frequency of the low-pass filter of Fig. 2.5(a) is given by the expression

$$f_{3dB} = \frac{g_m}{2\pi C} \tag{2.4.1}$$

Linear adjustment of f_{3dB} with gm is attainable with this circuit while maintaining a unity dc frequency gain. The structure of Fig. 2.5(b) has a fixed pole location and adjustable dc gain with the transconductance gain g_m. If the resistor in this circuit is replaced with the controlled resistor of Fig. 2.3(a), the circuit would have independently adjustable gain and break frequency. The high-pass structure of Fig. 2.5(c) also has a 3dB cut-off frequency given by

$$f_{3dB} = \frac{g_m}{2\pi C} \tag{2.4.2}$$

It can be observed that the characteristic networks for the low-pass and high-pass structures of Figs. 2.5(a) and 2.5(c) are identical, and thus they have the same pole structures. They differ only in where the excitation is applied.

The circuits of Figs. 2.5(d) and 2.5(e) act as shelving equalizers. The response of both circuits can be continuously changed from low-pass to all-pass to high-pass by adjusting gm as can be seen from Fig. 2.6. The basic difference in the two circuits is that the former has a fixed pole and adjustable zero, whereas the circuit of Fig. 2.5(e) has an adjustable pole and fixed zero. As for the circuit of Fig. 2.5(b), additional flexibility can be obtained if the grounded resistor in the circuit of Fig. 2.5(d) is replaced with the controlled resistor of Fig. 2.3(a).

The circuit of Fig. 2.5(f) acts as a low-pass filter with high frequency gain determined by the $C_1 : C_2$ ratio. Both the pole and zero in this circuit are adjustable through the parameter g_m but the ratio is held constant. This preserves the shape in the transfer characteristics and thus represents only a frequency shift in the response, as shown in Fig. 2.6(f).

The circuit of Fig. 2.5(g) utilizes an additional OTA and offers considerable flexibility. If either g_{m1}or g_{m2} fixed, the circuit behaves much like the shelving equalizers discussed above. If g_{m1} and g_{m2} are adjusted simultaneously, then a fixed pole-zero ratio and, hence, shape preserving response is possible. In this case, the circuit can be low-pass, all-pass, or high-pass, depending upon the g_{m1}: g_{m2} ratio. If the "+" and "-" terminals of g_{m1} are interchanged and the transconductance gains are adjusted so that $g_{m1} = g_{m2}$, the circuit behaves as a phase equalizer.

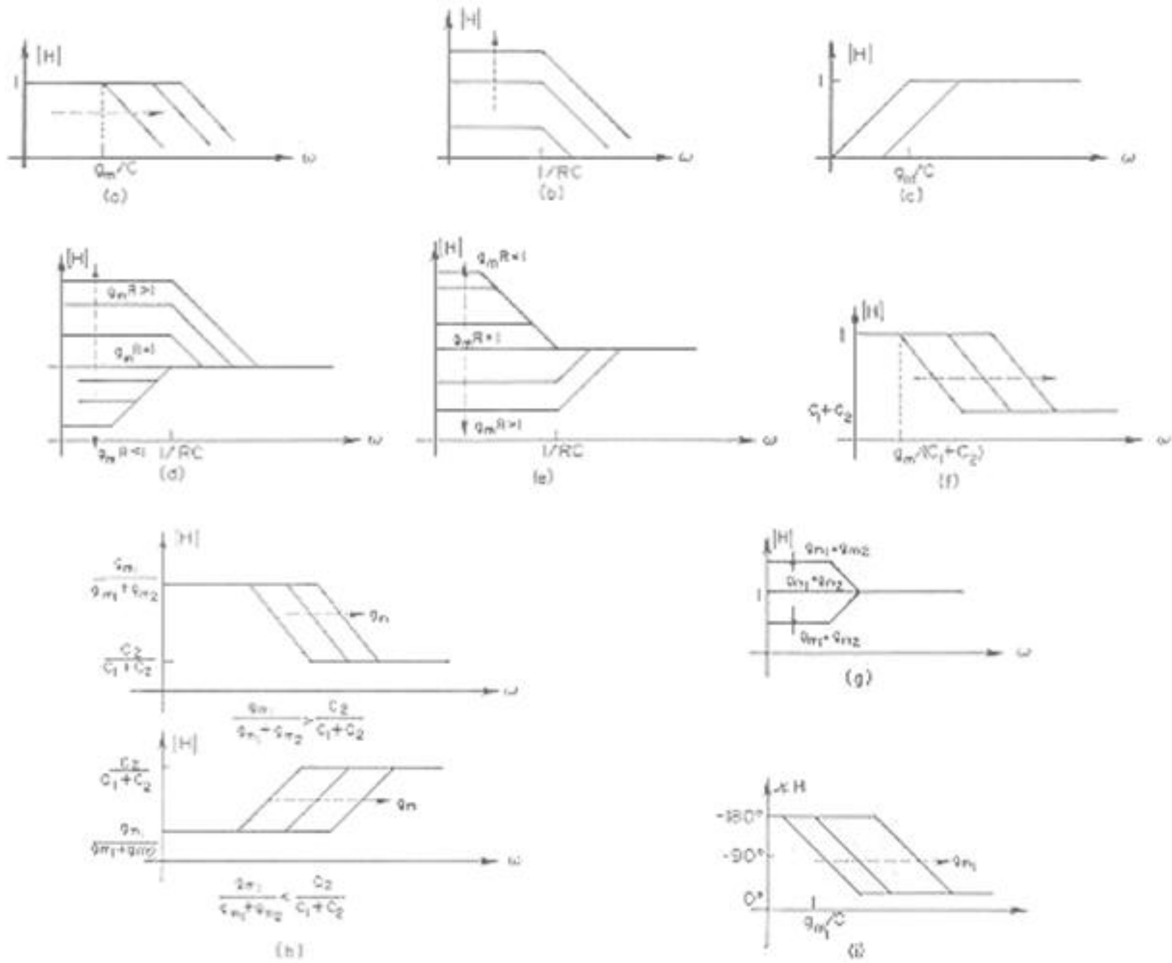


Fig. 2.6 Transfer characteristics for first-order structures of Fig. 5. (a) Circuit of Fig. 5a. (b) Circuit of Fig. 5b. (c) Circuit of Fig. 5c. (d) Circuit of Fig. 5d. (e) Circuit of Fig. 5e. (f) Circuit of Fig. 5f. (g) Circuit of Fig. 5g. (h) Circuit of Fig. 5h. (i) Circuit of Fig. 5i.

The circuit of Fig. 2.5(h) also preserves the shape of the transfer function, provided g_{m1} and g_{m2} are adjusted in such a manner that their ratio remains constant. In this case, the shape of the response is determined by the g_{m1}: g_{m2} and C_1: C_2 ratio. Depending upon these ratios, the response is either low-pass or high-pass in nature, as indicated in Fig. 2.6(h).

If g_{m2}R = 1, the circuit of Fig. 2.5(i) behaves as a phase equalizer, g_{m1} can be used to adjust the phase shift. For monolithic applications, the resistor R can be replaced with a third OTA, using the configuration of Fig. 2.3(a).

2.5 SECOND-ORDER STRUCTURES

Second-order filter structures find widespread applications directly and in the design of higher-order filters. Although the emergence of practical voltage or current-controlled first-order filters and amplifiers has been slow, even fewer techniques exist for the design of controlled second- and higher-order structures. Switched capacitor techniques have been successfully used to build voltage-controlled filter structures by building a voltage-controlled oscillator and using the output as the required clock for the switching of the capacitors. Although useful in some applications, these structures are not continuous time in nature, have limited dynamic range, and are limited to reasonably low-frequency applications. Concentration here will be on continuous-time voltage controlled structures.

One common requirement in the design of voltage controlled filter structures is that the filter characteristics be adjusted in a manner that essentially results in frequency scaling. In all-pole applications, such as the low-pass Butterworth and Chebyschev case, as well as the band-pass and high-pass versions of these approximations, the frequency scaling is tantamount to moving all poles a prescribed distance in a constant-Q manner. Those familiar with active filter structures will recall that pole movement in second-order structures through the adjustment of a single component is always on a circular path (constant wo) or on a straight line (constant bandwidth) parallel to the imaginary axis in the s-plane. The challenges associated with constant-Q pole adjustment through the simultaneous tuning of two or more components should be obvious.

A seemingly more difficult situation exists when considering the design of the popular elliptic filters. To maintain the elliptic characteristics as the cut-off frequencies changed, all poles *and* all zeros of the approximating function must be moved simultaneously and with the appropriate ratio in a constant-Q manner.

A group of second-order voltage-controlled filter structures are discussed in this section. Circuits with constant-Q pole adjustment, circuits with constant bandwidth ω_o, adjustment, and circuits with independent pole and zero adjustment are presented. Some circuits with simultaneous constant-Q adjustment of both the poles and zeros are also presented along with a general bi-quadratic structure. These structures have immediate applications in voltage-controlled Butterworth, Chebyschev, and Elliptic designs.

A simple second-order filter structure is shown in Fig. 2.7(a) [17], [19]. This structure is canonical in the sense that only four components are needed to obtain second-order transfer functions. The output voltage, *Vo*, is given by the expression

$$V_{01} = \frac{S^2 C_1 C_2 V_c + SC_1 g_{m2} V_B + g_{m1} g_{m2} V_A}{S^2 C_1 C_2 + SC_1 g_{m2} + g_{m1} g_{m2}} \qquad (2.5.1)$$

The transfer function for the specific excitations at V_A, V_B, and V_C are listed in the Table 2.1. Note that for $g_{m1} = g_{m2} = g_m$, the low-pass, band-pass, high-pass, and notch versions of this circuit all behave as ω_o adjustable circuits with fixed pole Q's. The pole Q's are determined by the capacitor ratio, which can be accurately maintained in monolithic designs. It is interesting to note that the zeros of the notch circuit also move in a constant-Q (i.e. along the jω axis) manner with the poles, as g_m is adjusted.

Occasionally, it is desirable to have circuits in which ω_o and Q of the poles can be independently adjusted. Two circuits with these characteristics are shown in Fig. 2.7(b) [18], [19], [24] and Fig. 2.7(c) [18], [24]. The output voltages for these circuits are, respectively,

$$V_{02} = \frac{S^2 C_1 C_2 V_c + SC_1 g_{m2} V_B + g_{m1} g_{m2} V_A}{S^2 C_1 C_2 + SC_1 g_{m2} g_{m3} R + g_{m1} g_{m2}} \qquad (2.5.2)$$

and

$$V_{03} = \frac{S^2 C_1 C_2 V_c + SC_1 g_{m2} V_B + g_{m1} g_{m2} V_A}{S^2 C_1 C_2 + SC_1 g_{m3} + g_{m1} g_{m2}} \qquad (2.5.3)$$

The circuits of Figs. 2.7(b) and 2.7(c) can be also used to implement low-pass, band-pass, high-pass, and notch transfer functions through the proper selection of the inputs as for the circuit of Fig. 2.7(a).

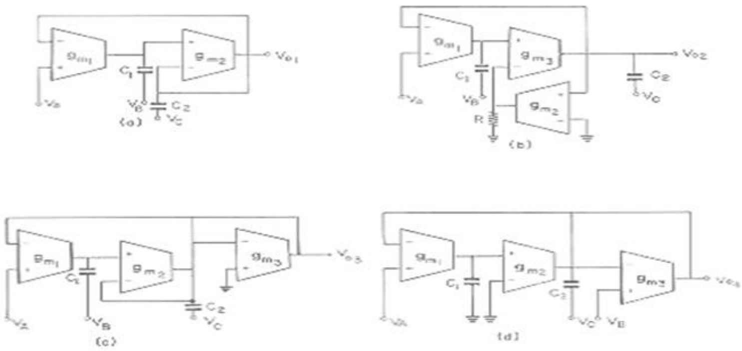


Fig. 2.7 Second-order filter structures [17],[19].

In the circuit of Fig. 2.7(b), the expressions for ω_o and Q of the poles of the circuit are given by

$$\omega_o = \sqrt{\frac{g_{m1}g_{m2}}{C_2C_1}} \tag{2.5.4}$$

and

$$Q = \frac{1}{g_{m3}R}\sqrt{\frac{C_2g_{m1}}{C_1g_{m2}}} \tag{2.5.5}$$

The poles can be moved in a constant-Q manner if g_{m3} is fixed, and if $g_{m1} = g_{m2} = g_m$ is adjusted; whereas movement in a constant ω_o manner is attainable if g_{m3} is adjusted when $g_{m1}$and g_{m2} remain constant. The independent adjustment of ω_o, and Q is apparent.

For the circuit of Fig. 2.7(c), the expressions for ω_o and Q of the poles become

$$\omega_o = \sqrt{\frac{g_{m1}g_{m2}}{C_2C_1}} \tag{2.5.6}$$

and

$$Q = \left(\sqrt{\frac{C_2}{C_1}}\right)\frac{\sqrt{g_{m1}g_{m2}}}{g_{m3}} \tag{2.5.7}$$

ω_o can be adjusted linearly with $g_{m1} = g_{m2} = g_m$ and *gm*3 constant. Such movement is often termed constant bandwidth movement. If g_{m1}, g_{m2}, and g_m are adjusted simultaneously, constant-*Q* pole movement is possible. Adjusting *gm*3 (for Q > 1/2) moves the poles along vertical lines parallel to the jω axis in the s-plane.

The circuit of Fig. 2.7(d) has an output given by

$$V_{04} = \frac{S^2C_1C_2V_c+SC_1g_{m3}V_B+g_{m1}g_{m2}V_A}{S^2C_1C_2+SC_1g_{m3}+g_{m1}g_{m2}} \tag{2.5.8}$$

The wo and Q of the poles are, respectively,

$$\omega_o = \sqrt{\frac{g_{m1}g_{m2}}{C_2C_1}} \tag{2.5.9}$$

$$Q = \frac{1}{g_{m3}} \sqrt{\frac{C_2 g_{m1} g_2}{C_1}} \qquad (2.5.10)$$

Although the transfer function is similar to that above, note that since the coefficient of the s term in the numerator equals that in the denominator, adjustment of the band-pass version of this circuit with $g_{m1} = g_{m2} = g_m$ will result in a constant bandwidth, constant gain response.

For monolithic structures, it may prove useful to replace the resistor in Fig. 2.7(b) with the OTA structure of Fig. 2.3(a). Likewise, if the bandwidth adjustment with g_{m3} is not needed, it may be desirable to replace the third OTA shown in Fig. 2.7(c) with a fixed resistor in some applications.

Circuit Type	Input Conditions	Transfer Function	If $g_{m1} = g_{m2} = g_m$	
			ω_o	Q
ω_o Adjustable Low-pass	$V_1 = V_A$ V_B and V_C Grounded	$\frac{g_{m1} g_{m2}}{s^2 c_1 c_2 + s c_1 g_{m2} + g_{m1} g_{m2}}$	$\frac{g_m}{\sqrt{C_1 C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
ω_o Adjustable Band-pass	$V_1 = V_B$ V_A and V_C Grounded	$\frac{s c_1 g_{m2}}{s^2 c_1 c_2 + s c_1 g_{m2} + g_{m1} g_{m2}}$	$\frac{g_m}{\sqrt{C_1 C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
ω_o Adjustable High-pass	$V_1 = V_C$ V_B and V_A Grounded	$\frac{s^2 c_1 c_2}{s^2 c_1 c_2 + s c_1 g_{m2} + g_{m1} g_{m2}}$	$\frac{g_m}{\sqrt{C_1 C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
ω_o Adjustable Notch	$V_1 = V_A = V_C$ V_B Grounded	$\frac{s^2 c_1 c_2 + g_{m1} g_{m2}}{s^2 c_1 c_2 + s c_1 g_{m2} + g_{m1} g_{m2}}$	$\frac{g_m}{\sqrt{C_1 C_2}}$	$\sqrt{\frac{C_2}{C_1}}$

Table 2.1 Transfer functions for bi-quadratic structure of Fig. 2.7(a).

Phase equalizers are also possible with the structures shown in Fig. 2.7. For example, interchanging the " + " and " - " terminals of the first two OTAs in Fig. 2.7(c), setting $V_A = V_B = V_C = V_i$,and making $g_{m1} = g_{m2} = g_{m3} = g_m$ results in a second-order g_m adjustable phase equalizer.

The circuit of Fig. 2.8 has both poles and zeros that can be adjusted simultaneously in a constant-Q manner. The circuit is similar to those shown in Fig. 2.7(a) with the exception that the capacitor C_2 in the previous circuits has been split to allow for adjusting the pole-zero ratios. The transfer function of the circuit is given by

$$\frac{V_o}{V_i} = \left(\frac{C_2}{C_2+C_3}\right)\left(\frac{S^2 + {g_{m1}}/{C_1 C_2}}{S^2 + S\,{g_{m2}}/{(C_2+C_3)} + {g_{m1} g_{m2}}/{C_1 (C_2+C_3)}}\right) \tag{2.5.11}$$

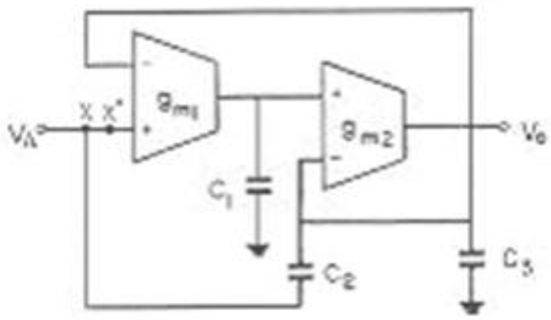


Fig. 2.8 Elliptic Filter structure.

This circuit has applications in higher-order voltage controlled elliptic filters. For higher-order structures obtained by cascading these second-order blocks, all g_m's would be made equal and adjusted simultaneously. Buffering between stages using a standard unity gain buffer is required to prevent interstage loading. Modifications of the other circuits in Fig. 2.7 to obtain wo and Q adjustable features is also possible. Although the ratio of the zero location to pole location can be controlled with the C_2/C_3 ratio in discrete designs, this may pose some problems in monolithic structures. One convenient way to control the pole-zero ratio is to insert the voltage-controlled amplifier of Fig. 2.2(g) between the points x and x' in Fig. 2.8 and use the transconductance gain of either of these additional OTAs as the control variable.

The final second-order structure considered here is the general biquad of Fig. 2.9. The output for this circuit is given by

$$V_0 = \frac{S^2 C_1 C_2 V_c + S C_1 g_{m4} V_B + g_{m2} g_{m5} V_A}{S^2 C_1 C_2 + S C_1 g_{m3} + g_{m1} g_{m2}} \qquad (2.5.12)$$

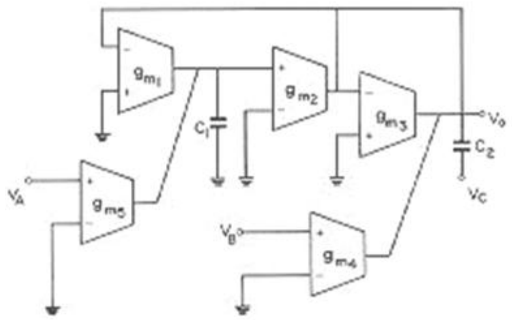


Fig. 2.9 General bi-quadratic structure.

The potential for tuning the w. and Q for both the poles and zeros (when $V_A = V_B = V_C = V_i$) to any desired value should be apparent. Although somewhat component intense, it can be argued that if there is to be capability for completely arbitrary location of a pair of poles and a pair of zeros via adjustment of the transconductance gain of the OTA, then at least 4 degrees of freedom and, hence, 4 OTAs are required. This circuit uses only one more than the minimum! The capability for various types of pole and/or zero movement through the simultaneous adjustment of two or more of the transconductance gains should also be apparent.

Emphasis in this section has been placed entirely upon second-order structures in which the desired filter characteristics depend directly upon the transconductance gain of the OTA. Very simple structures in which the filter characteristics are adjustable through the parameter g_m resulted. As stated in the introduction, g_m is readily controllable by a dc bias current over a wide range of values, thus making these circuits directly applicable to voltage controlled applications. Several of the more recent works on OTA applications [18]-[24] have followed this approach. Most of the earlier works [7]-[16] and the circuits presented in the manufacturer's application notes [3]-[5] concentrated upon topologies in which the filter characteristics are independent or only mildly dependent upon the transconductance gain. Most of these structures are very complicated, very component-intense, and require tuning algorithms that are unwieldy. Alternatives to these earlier designs using conventional operational amplifiers have proven to be much better.

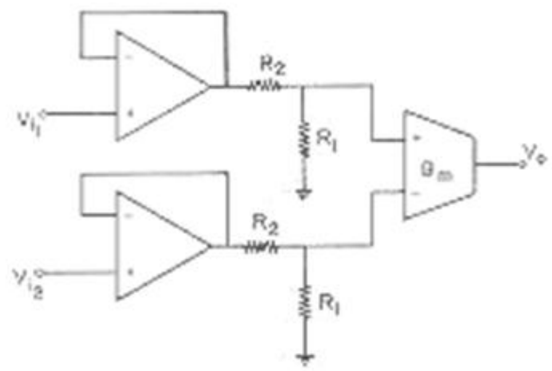


Fig. 2.10 Signal conditioner for OTAs.

2.6 PRACTICAL CONSIDERATIONS

Although all circuits presented up to this point in this paper are practical with ideal operational transconductance amplifiers, existing discrete OTAs are far from ideal. As mentioned in the introduction, the major limiting factor with commercially available OTAs is the limited differential input voltage swing. Recent activity in the literature has concentrated upon designing OTAs with improved input characteristics [27]-[28]. Significant improvements in performance over what is currently available with discrete OTAs have been demonstrated. An alternative is to use voltage attenuators and buffers at the input of existing OTAS. This technique is often suggested in the manufacturer's application notes and is illustrated in Fig. 2.10. This technique can be used to obtain reasonable signal swings with all circuits discussed up to this point. Although such circuits are useful, a rather high price is paid for this modification. First**,** the circuit requires many more components. Second, the finite bandwidth of the op amps will limit the frequency response of the OTA structures. Finally, the attenuation of the input signal to the OTA causesa serious loss in dynamic range. From a topological point of view, some OTA based structures are inherently more susceptible to differential voltage limitations than others. This parallels the concern for op amp based active *RC* and switched-capacitor structures that the signal amplitudes at the output of internal op amps assume acceptable values. These considerations become more serious for high Q and high dynamic range applications.

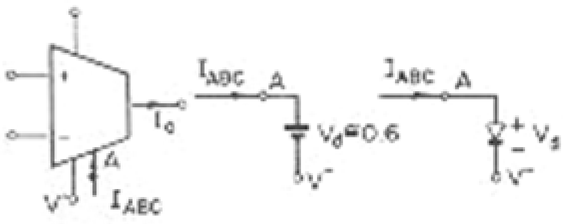


Fig. 2.11 Macromodel of bias current port on bipolar OTA [27-28].

A macro model of the bias current (I_{ABC}) input port of a typical bipolar OTA is shown in Fig. 2.11. This actually forms part of an internal current mirror that is discussed later. Several schemes for controlling the current (I_{ABC}) and, thus, the g_m of the OTA by an external control voltage, *Vc*, are shown in Fig. 2.12.

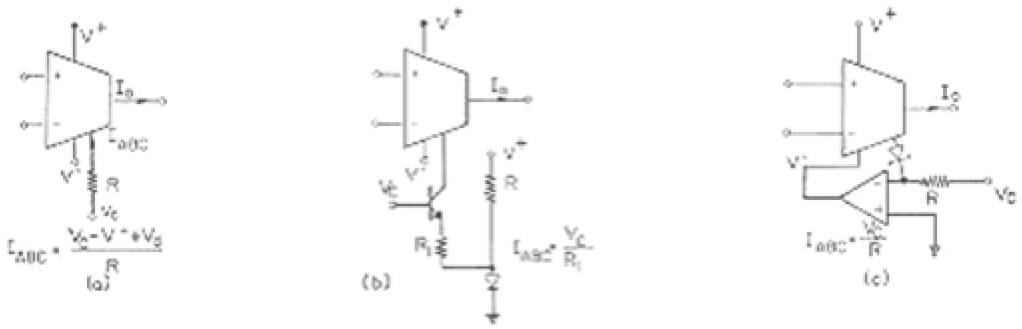


Fig. 2.12 Schemes for obtaining voltage control with the OTA.

The first circuit is the simplest but is very sensitive to small changes of V_c *as* V_c approaches .6*v* + *V*-. In the second circuit, the control voltage is referenced to zero but the small V_c is sensitive to mismatches between the B-E voltage of the transistor and the forward diode voltage drop. In the circuit of Fig. 2.12(c), the control voltage is also referenced to ground and is not dependent upon the matching or cancellation of voltages across external forward biased *pn* junctions. The zener diode is used to maintain the common mode voltage at a reasonable level. The frequency response of the op amp is not of concern here since it is used only in the dc control path. It should be noted that the amplifier bias current is proportional to V_c for all schemes shown in Fig. 2.12. Since I_{ABC} can typically be adjusted over several decades, all schemes will be very sensitive to small changes in V_c toward the low current end of the I_{ABC} range. Logarithmic amplifiers are often used to control I_{ABC} with an external control voltage if the wide adjustment range of I_{ABC} is to be effectively utilized.

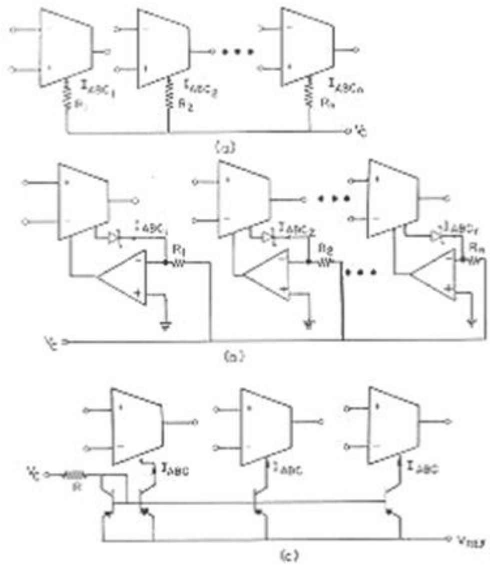


Fig. 2.13 Schemes for simultaneous gm adjustment.

Many of the filter circuits discussed in the previous sections of this paper require the simultaneous adjustment of matched g_m's. Several schemes for achieving this are shown in Fig. 2.13. In the first circuit, it is easy to adjust the g_m's by trimming the resistors for a fixed g_m. The circuit is quite sensitive to the slight differences in the voltage V_d of Fig. 2.11(a) for small values of I_{ABC}. The circuit of Fig. 2.13(b) again has V_C referenced to ground and is essentially independent of the matching of V_d for the individual OTAs. The scheme of Fig. 2.13(c) is useful if an external single package *pnp* current mirror with *n* outputs is available. A discrete component version of this mirror would not be practical.

For integrated circuit applications, the amplifier bias currents of several OTAs are particularly easy to match and control. For monolithic applications, the simultaneous adjustment of the gain of a large number of OTAs with a single dc bias current can be easily attained by using a single input-multiple output current mirror such as is shown in Fig. 2.14. This structure actually *replaces* the bias current mirrors on each of the OTAS. The transconductance gains can be ratioed, if desired, by correspondingly ratioing the emitter areas (or width length ratio for MOS structures) in the outputs of the current mirror.

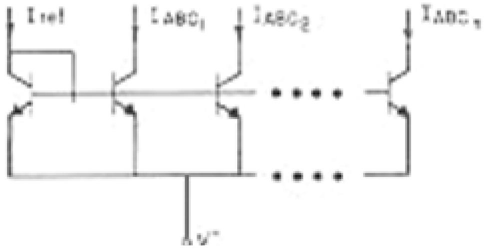


Fig. 2.14 Single input-multiple output bias current generator for monolithic applications.

With conventional operational amplifiers, the slew rate, input impedance, output impedance, and maximum output current are essentially fixed at the design stage. For OTAS, it is generally the case that these parameters are either proportional or inversely proportional to I_{ABC}. Thus adjusting g_m via I_{ABC} causes all of these parasitic parameters to change accordingly. Although the user should be cognizant of the changes in these parameters, the problems they present are manageable. The output capacitance of an OTA does cause concern at low output currents and high frequencies.

Much as in the design of conventional op amp based circuits, a dc bias current path must be provided for both input terminals of the OTA. Although the amplifier of Fig. 2.15 serves as an effective g_m attenuator, which will prove useful in some applications, the circuit is useless since the required input bias current will cause an accumulation of charge on the capacitors and eventual saturation of the OTA. It may be mentioned here that more complicated circuits with the same problem are suggested in the literature [17].

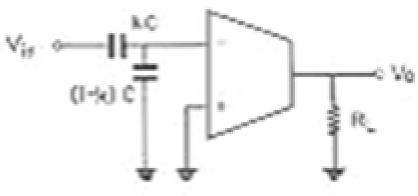


Fig. 2.15 g_m attenuator [17].

Numerous nonlinear applications of OTA structures exist. Suffice it to say that since the amplifier bias current, I_{ABC} can be considered as a third signal input, simple multipliers, modulators, and a host of other nonlinear circuits are possible. The reader is referred to the application notes for a discussion of some of the nonlinear applications. Some of the structures that use only OTAs and capacitors show promise for monolithic applications in MOS or bipolar processes. The circuits should offer high frequency continuous-time capabilities. Either external voltage-control or an internal reference circuit to compensate for process and temperature variations will be necessary to make these circuits practical in demanding applications.

Finally, it should be noted that some of the filter structures presented earlier in this paper have a non-infinite input impedance, and that the output impedance is generally quite high. Cascading of such structures will require interstage buffer amplifiers, which will tend to degrade the bandwidth of the overall filter structures. Output buffers are also generally required to drive external loads.

2.7 CONCLUSIONS

In the present chapter, a detailed and comprehensive review of OTA based signal processing circuits has been presented. The characteristics of these circuits are adjusted with the externally accessible dc amplifier bias current. Most of these circuits utilize a very small number of components. Applications include amplifiers, controlled impedances, and filters. Higher-order continuous-time voltage-controlled filters such as the common Butterworth, Chebyschev, and Elliptic types can be obtained. In addition to the voltage control characteristics, the OTA based circuits show promise for high-frequency applications where conventional op amp based circuits become bandwidth limited. The major factor limiting the performance of OTA based filters using commercially available OTAs is the severely limited differential input voltage capability inherent with conventional differential amplifier input stages. Recent research results suggested significant improvements in the input characteristics of OTAs can be attained [27]-[28]. Several groups have utilised OTAs in continuous time monolithic filter structures [28-40]. With recent advancements in the semiconductor manufacturing technology and current mode signal processing OTAs have an important place.

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