CHAPTER-IV

INDUCTORLESS FILTER REALIZATION SCHEME : A GENERAL APPROACH

4.1 INTRODUCTION

In the preceding chapter a floating inductor realization scheme and also its application in filter design has been demonstrated. The ideal floating inductor requires quite a few resistors and active blocks making the final filter topology sufficiently elaborate. Instead integrated development of inductorless filters through active RC form yields economic structures. In this direction individual schemes or particular forms have been produced[1-3] with success. This chapter presents a generalized appreach to realize inductorless filter schemes with a stress on allpass response characteristics. Allpass filters have large scale applications in the areas of communication and control systems. A typical example of its use is as a delay network in a communication channel or in a control loop where a time delay may be required to be introduced without reduction in the signal strength. The presented scheme uses interconnection of one three terminal network with another four terminal one, constrained by a single operational amplifier (OA). The scheme is sufficiently general in the sense that it can realize allpass, low, high and bandpass and also band elimination filter functions. It obtains these functions with complex poles such that high-Q realization is implied. As pointed out, the method starts with an approach to generalize the active RC-allpass schemes with a single OA.

4.2 THE GENERALIZATION OF THIS ALL-PASS FILTER[4*]

For obtaining a complete generalization scheme of the active RC allpass networks[5-16] with a single differential input operational amplifier, a 5-terminal RC network constrained by an operational amplifier as shown in figure 4.1 would be required. The terminal 4 and 5 are the output terminals and 1 is the input terminal. Employing only the constraints (a) $E_2 = E_3$, (b) $I_2 = I_3 = 0$ and (c) I_4 being arbitrary, dictated by the ideal OA to the admittance matrix of this 5-terminal network, it is possible to realize the voltage transmission matrix as

$$\left[t \right] = \left[t_{41},\ t_{51} \right] \tag{4.1}$$

This generalization covers almost the entire class of single OA allpass realization schemes. However, the constraints (a) to (c) can reduce the admittance matrix order from 5x5 to 4x3 at the most, but as such this reduction offers little advantage in evaluating the matrix $\left[t \right]$. Furthermore, if the OA has a finite gain A, constraint (a) changes to (a′) $(E_2 - E_3)A = E_4$. Many of the above schemes[3-6, 11-15] have, however, used a 4-terminal constrained network to realize an all pass transmission function t_{41} In fact, the constraint provided by the OA, both for finite and infinite gain configurations, in a 4-terminal network make the evaluation of t_{41} considerably easier. The other schemes[7-10] have also used a 4-terminal (including a grounded one) network, but terminal 2 has been grounded in these, and the output has been obtained at 5 realizing t_{51}. The latter schemes were further simplified by making $y_{13} = y_{31} = g_{31}$ and the realization then

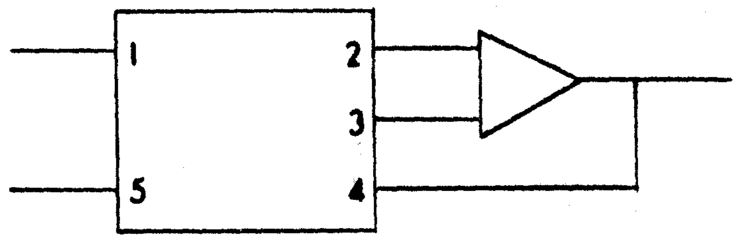

Fig. 4.1 The 5-Terminal network scheme

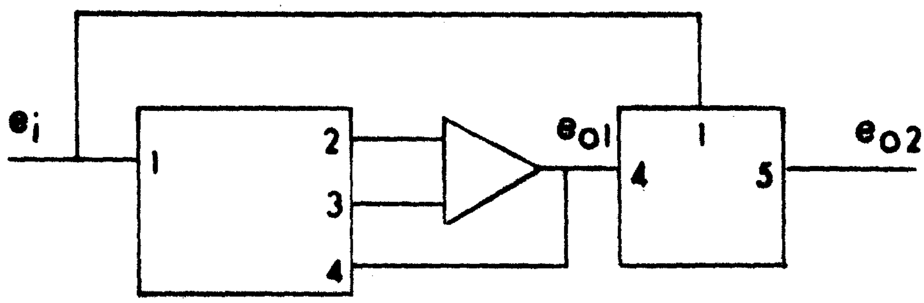

Fig. 4.2 The proposed scheme of generalisation

actually involves the synthesis of 3-terminal ladder or parallel ladder network. Combining the above two schemes, therefore, a generalized scheme can be attempted where the simplicity of analysis and synthesis of both the schemes can be retained. In the following, such a scheme, interconnecting a 3-terminal network with a 4-terminal network constrained by a differential input OA is presented. The evaluation of the overall transfer function in this network is considerably simplified. The proposed generalization scheme is shown in figure 4.2 where the voltages at the input (terminal 1), first output terminal (terminal 4) and second output terminal (terminal 5) are denoted by e_i, e_{01} and e_{02} respectively. The voltage transfer function t_{51} as defined earlier would now have a changed significance as indicated in figure 4.2. So, for brevity, t_{41} and t_{51} of equation (4.1) are replaced by t_{01} and t_{02} respectively, which are defined in equations (4.4) and (4.6) below. Applying the constraints (a) and (b) to the admittance matrix of the 4-terminal network one easily obtains

$$\begin{bmatrix} \frac{I_1}{y_{11}} \\ 0 \\ 0 \\ \frac{I_4}{y_{44}} \end{bmatrix} = \begin{bmatrix} 1 & \frac{1}{y_{11}A}(y_{12} - y_{13} + Ay_{14}) \\ \frac{y_{21}}{y_{22}} & \frac{1}{y_{22}A}(y_{22} - y_{23} + Ay_{24}) \\ \frac{y_{31}}{y_{33}} & \frac{1}{y_{33}A}(y_{32} - y_{33} + Ay_{34}) \\ \frac{y_{41}}{y_{44}} & \frac{1}{y_{44}A}(y_{42} - y_{43} + Ay_{44}) \end{bmatrix} \begin{bmatrix} e_i \\ e_{01} \end{bmatrix} \qquad (4.2)$$

from which the voltage transfer function $\frac{e_{01}}{e_i}$ is obtained as

$$\frac{e_{01}}{e_i} = \frac{A\left(\frac{y_{31}}{y_{33}} - \frac{y_{21}}{y_{22}}\right)}{A\left(\frac{y_{24}}{y_{22}} - \frac{y_{34}}{y_{33}}\right) - \left(\frac{y_{23}}{y_{22}} + \frac{y_{32}}{y_{33}} - 2\right)} \tag{4.3}$$

For the infinite gain OA realization, as is mostly the case, equation (4.3) simplifies to

$$t_{01} = \left.\frac{e_{01}}{e_i}\right|_{A \to \infty} = \frac{\frac{y_{31}}{y_{33}} - \frac{y_{21}}{y_{22}}}{\frac{y_{24}}{y_{22}} - \frac{y_{34}}{y_{33}}} \tag{4.4}$$

For obtaining t_{02} , we define for the 3-terminal network

$$t_{51} = \frac{e_{02} - e_{01}}{e_i - e_{01}} \tag{4.5}$$

and hence from figure 4.2

$$t_{02} = \left.\frac{e_{02}}{e_i}\right|_{A \to \infty} = t_{01} + t_{51}\,(1 - t_{01}) \tag{4.6}$$

The transmittance t_{51} can be written as the ratio of two second-order cofactors of the admittance matrix of the 3-terminal network.

$$\left[y_3\right] = \begin{bmatrix} y_{11} & y_{14} & y_{15} \\ y_{41} & y_{44} & y_{45} \\ y_{51} & y_{54} & y_{55} \end{bmatrix} \tag{4.7}$$

following the usual definition[17], and is given as

$$t_{51} = \frac{y_{54}^{14}}{y_{14}^{14}} \tag{4.8}$$

t_{02} can also be obtained from t_{54}, given as

$$t_{54} = \frac{e_{02} - e_i}{e_{01} - e_i} = \frac{y_{51}^{41}}{y_{41}^{41}} \tag{4.9}$$

when

$$t_{02} = 1 + t_{54}\,(t_{01} - 1) \tag{4.10}$$

t_{01} becomes identical with t_{02} if $t_{51} = 0$ or $t_{54} = 1$, which from equation (4.5) or (4.9) implies $e_{02} = e_{01}$.

Following equation (4.6) or (4.10) almost all the allpass networks are obtainable from figure 4.2 with the proper choice of y_{ij}'s in equation (4.4) and y_{mn}^{ki}'s (which are, again, y_{pq}'s only) in equation (4.8) or (4.9). We choose equation (4.6) and show by way of example how some of the important allpass realization schemes are derived from it.

The circuit of Dutta Roy[6] is easily obtained if

$$y_{34} = 0$$

$$\frac{y_{21}}{y_{22}} = \frac{1/R_1}{1/R_1 + 1/R_2} = \frac{\lambda}{1 + \lambda}$$

and

$$y_{24} = y_{22} - y_{21}$$

$$t_{51} = 0$$

These give

$$t_{(DR)} = \frac{y_{31}(1+\lambda) - y_{33}\lambda}{y_{33}} \tag{4.11}$$

The Deliyannis[11] circuit is obtained for

$$y_{34} = 0$$

$$\frac{y_{31}}{y_{33}} = \frac{g_a}{g_a + g_b} = \frac{1}{1+k}$$

$$y_{22} = y_{21} + y_{22}'$$

and

$$t_{51} = 0$$

The transfer function is given as

$$t_{(D)} = \frac{y_{22}' - ky_{21}}{(1+k)y_{24}} \tag{4.12}$$

The circuits of Schoonaert and Kretzschmar[7] and that of Patranabis[1(
requires the component arrangement such that

$$y_{22} = \alpha \; ; \; y_{34} = g_{34} \; ; \; y_{31} = g_{31} \; ; \; y_{33} = g_{31} + g_{34} \; ; \; m = \frac{g_{31}}{g_{34}}$$

and $t_{51} = \beta_o\,(s, R, C)$ giving

$$t_{(P)} \text{ or } t_{(SK)} = \beta_o\,(1 - m) + m \tag{4.13}$$

β_o must be synthesised in the ladder or parallel ladder form[10].
In an earlier generalization scheme[18], which required one 3-termina

network less than the Liberatore generalization[19], the Holt and Gray circuit was not realizable. In the present scheme this also is possible with the following identification

$$y_{31} = s + 1 \; ; \quad y_{21} = y_{24} = y \text{ (say)}; \; y_{34} = \frac{2s}{s + 1}$$

$$y_{33} = s + 1 + \frac{2s}{s + 1} + \frac{s}{s + 1}$$

$$y_{22} = 2y \text{ and } t_{51} = 0$$

This gives

$$t_{(HG)} = \frac{1 - s + s^2}{1 + s + s^2} \tag{4.14}$$

Most of the other circuits are only special cases of the above schemes and can similarly be shown to be realizable by this scheme including a circuit[17] with a finite gain OA.

The generalization has thus been shown to cover the allpass realization scheme almost fully, except perhaps the cases where the OA is used as a voltage-controlled current source[20]. It would further be noted that this structure with appropriate choice of the passive network could be adapted to realize any arbitrary transfer functions besides the allpass ones already discussed.

4.3 HIGH, LOW, BAND-PASS AND NOTCH FILTER REALIZATION

Some examples from second order, high, low, band-pass and notch filters are given below for confirming the generalization of the second scheme.

4.3.1 HIGH-PASS FILTER REALIZATION

From equation (4.4) and (4.10) for $t_{54} = 1$, we get

$$t_{02} = \frac{\dfrac{y_{31}}{y_{33}} - \dfrac{y_{21}}{y_{22}}}{\dfrac{y_{24}}{y_{22}} - \dfrac{y_{34}}{y_{33}}} \tag{4.15}$$

Letting now,

$$y_{21} = s + 1 \;;\quad y_{24} = s + 3 \;;\quad y_{22} = 3s + 4$$

$$y_{31} = s + 2 \;;\quad y_{34} = \tfrac{1}{2}(s+1);\quad y_{33} = 2s + 8$$

equation (4.15) yields

$$t_{02} = \frac{2s^2}{s^2 + 18s + 40} \tag{4.16}$$

This is the transfer function of a High-pass filter with real poles.

However for highpass filters with complex poles the following selections are made

$$y_{21} = s + 1 \;;\quad y_{24} = s + 3 \;;\quad y_{22} = 3s + 4$$

$$y_{31} = s + 2 \;;\quad y_{34} = \tfrac{1}{2} \;;\quad y_{33} = 2s + 8$$

The transfer function is then obtained as

$$t_{02} = \frac{2s^2}{4s^2 + 25s + 44} \tag{4.17}$$

4.3.2 BAND-PASS FILTER REALIZATION

In equation (4.15) assuming

$$y_{21} = s + 1 \; ; \quad y_{24} = 2s + 1 \; ; \quad y_{22} = 3s + 2$$

$$y_{31} = s + 2 \; ; \quad y_{34} = s + 1 \; ; \quad y_{33} = 3s + 4$$

the band-pass filter transfer function with real poles is obtained as

$$t_{02} = \frac{s}{3s^2 + 6s + 2} \tag{4.18}$$

If however one identifies,

$$y_{21} = y_{24} = s + 1 \; ; \quad y_{22} = 3s + 2$$

$$y_{31} = s + 2 \; ; \quad y_{34} = 1 \; ; \quad y_{33} = 3s + 4$$

a band-pass filter with complex poles is obtained,

$$t_{02} = \frac{s}{3s^2 + 4s + 2} \tag{4.19}$$

4.3.3 LOW-PASS FILTER REALIZATION

A real pole second order filter can be easily synthesised with appropriate choice of y_{ij}'s. An example of complex pole second order filter is given below with

$$y_{21} = s + 5 \; ; \quad y_{24} = 3s + 11 \; ; \quad y_{22} = 4s + 16$$

$$y_{31} = s + 3 \; ; \quad y_{34} = 2s + 3 \; ; \quad y_{33} = 4s + 8$$

These yield

$$t_{02} = \frac{2}{s^2 + 6s + 10} \tag{4.20}$$

4.3.4 NOTCH FILTER REALIZATION

If one makes $t_{51} = \frac{1}{2}$ instead of 0 in the Holt and Gray realization procedure given above, then from equation (4.6)

$$t_N(s) = \frac{s^2 + 1}{s^2 + s + 1} \tag{4.21}$$

indicating that a notch network function is realizable by the scheme.

4.4 CONSIDERATION OF THE FINITE GAIN-BANDWIDTH PRODUCT OF THE OPERATIONAL AMPLIFIER

Since the constraints of infinite gain was not used in the beginning for obtaining the voltage transfer function from the 4-terminal network, equation (4.2) can be conveniently used to evaluate the effect of the finite gain-bandwidth product of the OA in the realization schemes. It may be shown that a simple-pole model of the OA introduces an additional pole in all these realizations with t_{01} now given as

$$t_{01f} = \frac{\dfrac{y_{31}}{y_{33}} - \dfrac{y_{21}}{y_{22}}}{\dfrac{y_{24}}{y_{22}} - \dfrac{y_{34}}{y_{33}} - \dfrac{s}{\omega_B}\left(\dfrac{y_{23}}{y_{22}} + \dfrac{y_{32}}{y_{33}} - 2\right)} \tag{4.22}$$

where ω_B is the unity gain-bandwidth of the amplifier.

4.5 CONCLUSION

The generalization scheme of active RC filters with a single operational amplifier has been described in this chapter. The scheme may be used as a single stock of active RC filter function realization schemes.

In Chapter-II a simulated inductor has been shown to generate sinusoidal oscillations. High, low and band-pass filters realized through integrated approach can be equally effective in producing harmonic generation. Further through a suitable choice, single element frequency controllability of such oscillators is easily obtained. This is borne out in the next chapter with associated details and applications.

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* Chapter-IV is based mainly on this publication.