where the VA decoder finds its application, it is necessary to choose larger values for L to get a reasonable signal-to-noise ratio at the decoder input. This can be satisfied by a receiver using a DFE and its performance may be better than a low complexity VA decoder working on a short DIR. This situation has been met in simulations made in [10] and in our simulation results reported in **Fig. 4.**

A high-distortion channel as dealt with in [5] has been simulated. It has the coefficients (—0.1, —0.1, —0.3, —0.15, —0.33, 0.16, 0.33, 1, 0.33, 0.16, —0.33, —0.3, 0.15, —0.1, —0.1). 10 000 symbols taking values in an alphabet $\{\pm 1, \pm 3\}$ are emitted for different values of noise variance. A spectrum shaping filter of 31 tap coefficients is used. A DIR of length 3 is used with the VA decoder, while the DFE works on a DIR of length 10. Although the VA works with 16 survivors, the large amount of residual ISI causes its performance to be lower than of the DFE. The larger value of DIR length given to the DFE has considerably reduced the residual ISI. If this value of L had been given to the VA, it would have required 4^9 survivors and the VA decoder would have then given a better performance than the DFE.

The DDFE has the same receiver structure of Fig. 3 where the decoder used is a reduced-state VA working with 2 survivors as it has been explained in section 2. Since the complexity is reasonable and is nearly independent of the DIR memory length L, a value of L of the same order as that of the DEF can be chosen and thus a good signal-to-noise ratio can be delivered to the decoder. If the same filters $\{p_i\}$ and $\{h_i\}$ of the DFE are used, a better performance than that of the DFE is guaranteed due to the more likely decisions made by the two-survivor decoder. In addition, error propagation is also lower with respect to the DFE.

Simulation results are illustrated in Fig. 4, where like in the DFE, a 31-coefficient prefilter and a 10-coefficients DIR are used.

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Active Compensated Bandpass Filter with Reduced Sensitivity to Op Amp Gain Bandwidth Product

Aktiv kompensierter Bandpaß mit reduzierter Empfindlichkeit für das Verstärkungs-Bandbreiten-Produkt von Operationsverstärkern

By Ahmed M. Soliman*)

Abstract:
A new configuration for realizing an inverting bandpass filter is introduced. The network is canonic and has extremely low sensitivities to the op amp unity gain bandwidth product. Design equations are given. Comparison with other well known bandpass filters is included.

Übersicht:
Es wird eine neue Anordnung zur Realisierung eines invertierenden Bandpasses vorgestellt. Das kanonische Netzwerk hat extrem niedrige Empfindlichkeiten für das Verstärkungs-Bandbreiten-Produkt von Operationsverstärkern. Es werden Ausführungsformen gezeigt. Vergleiche mit anderen bekannten Bandpaßfiltern werden angestellt.

Für die Dokumentation:
Bandpaßfilter / Operationsverstärker / Verstärkungs-Bandbreite-Produkt / Empfindlichkeit

1. Introduction

The design of selective active RC filters by activating passive RC building blocks usually leads to very low active sensitivities [1—6]. Recently a single op amp active RC bandpass filter based on this idea was given [5]. The circuit is shown in **Fig. 1** and requires two identical passive RC bandpass building blocks for N_1 and N_2. It has the attractive advantage of extremely low sensitivities to the op amp unity gain bandwidth product which is due to

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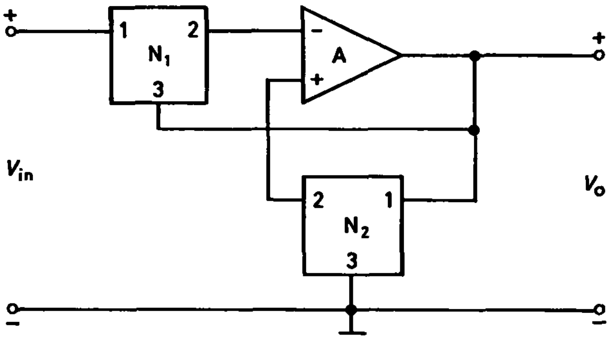

Fig. 1: The passive compensated noncanonic bandpass filter [5]

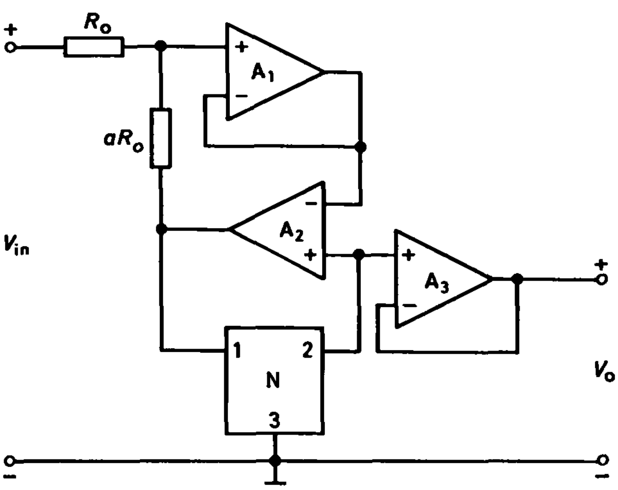

Fig. 2: The new active compensated canonic bandpass filter

the additional feedback obtained from the passive network N_2 and thus may be considered as a passive compensation method.

The purpose of this paper is to introduce a new canonic active RC bandpass network having the same frequency limitation equations as the previous circuit [5] but using active compensation instead of passive compensation.

2. The general configuration

Fig. 2 represents the proposed building block which requires three op amps. Assuming the op amps to be ideal except for the finite frequency dependent open loop gain A. By direct circuit analysis the transfer function of the circuit in Fig. 2 is obtained as

$$G(s) \equiv \frac{V_0}{V_{in}} =$$

$$= \frac{-aT}{\left(1+\frac{1}{A_3}\right)\left[1-(a+1)\left(1+\frac{1}{A_1}\right)\left(T-\frac{1}{A_2}\right)\right]}, \quad (1)$$

where $T(s)$ is the transfer function of the passive RC network N. Taking $a = 1$, and assuming matched op amps are used, that is

$$A_1 = A_2 = A_3 = A$$

the transfer function becomes

$$G(s) = \frac{-T}{(1-2T) + \frac{1}{A}(3-4T) + \frac{2}{A^2}(2-T) + \frac{2}{A^3}}. \quad (2)$$

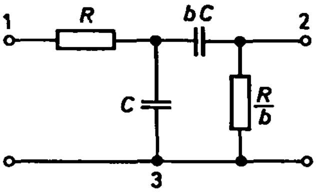

Fig. 3: The passive RC network N

3. Bandpass filter realization

Using a passive RC bandpass network for N thus

$$T(s) = \frac{K\omega_p s}{s^2 + \left(\frac{\omega_p}{q_p}\right)s + \omega_p^2}, \quad (3)$$

where $0 < q_p < 0.5$ and $K q_p \leqslant 1$.

Substituting (3) in (2) and as $A \to \infty$ gives

$$G(s) = \frac{-H\frac{\omega_0}{Q}s}{s^2 + \left(\frac{\omega_0}{Q}\right)s + \omega_0^2}, \quad (4)$$

where

$$\omega_0 = \omega_p, \quad (5)$$

$$Q = \frac{q_p}{1-2Kq_p}, \quad (6)$$

$$H = |\text{Gain}|_{\omega_0} = KQ. \quad (7)$$

For the proposed passive RC network N shown in **Fig. 3**, the parameters K, ω_p and q_p are given by

$$K = 1, \quad \omega_p = \frac{1}{CR}, \quad q_p = \frac{1}{2+b}. \quad (8)$$

Thus the ω_0 and the Q of the active filter are given by

$$\omega_0 = \frac{1}{CR}, \quad (9)$$

$$Q = \frac{1}{b}. \quad (10)$$

It is seen that the parameter b controls the selectivity of the filter without affecting ω_0.

The design equations for $\omega_0 = 1$ are given by

$$C = 1, \quad R = 1, \quad a = 1 \quad \text{and} \quad b = 1/Q. \quad (11)$$

4. Effect of the rolloff of the op amp gain

It is well known that the op amp unity gain bandwidth product is likely to be a major limiting factor in the performance of active filters. Here the frequency limitation equations of the network are given based on the one pole rolloff model of the op amp which is characterized by

$$A = \frac{A_0\omega_1}{s+\omega_1} \simeq \frac{GB}{s}, \quad (12)$$

where

- A_0 is the open loop dc gain of the op amp,
- ω_1 is the open loop 3-dB bandwidth, and
- $GB = A_0\omega_1$ is the gain bandwidth product.

When (12) and (3) are substituted in (2), using (9) and (10) and neglecting higher order terms, the denominator of $G(s)$ becomes

$$D(s) = \left(s^2 + \frac{\omega_0}{Q} s + \omega_0^2\right) + + \frac{3s}{GB}\left(s^2 + \left(\frac{1}{Q} + \frac{2}{3}K\right)\omega_0 s + \omega_0^2\right). \tag{13}$$

Following Budak-Petrela analysis [7] it follows that the relative change in ω_0 and Q due to the limited frequency response of the op amps are given by

$$\frac{\Delta\omega_0}{\omega_0} = -K\frac{\omega_0}{GB}, \tag{14}$$

$$\frac{\Delta Q}{Q} = K(2K-1)\frac{\omega_0}{GB}. \tag{15}$$

For the special passive RC network shown in Fig. 3 for which $K = 1$ the above equations reduce to

$$-\frac{\Delta\omega_0}{\omega_0} = \frac{\Delta Q}{Q} = \frac{\omega_o}{GB}. \tag{16}$$

5. Conclusions

Table 1 includes comparison of the given network with other well known low sensitivity active RC bandpass filters. It is seen that the proposed network has identical fractional shifts in ω_0 and Q as those of the passive compensated bandpass filter introduced recently by the author [5].

It should be noted that the voltage follower at the output may be replaced by a voltage controlled voltage source to provide independent control on the gain without affecting the filter selectivity.

Table 1

Bandpass network type	Circuit components			Fractional shifts in ω_0 and Q	
	Op Amps	R	C	$\frac{\Delta\omega_0}{\omega_0}$	$\frac{\Delta Q}{Q}$
Wilson-Bedri-Bowron circuit [6]	2	5	3	$-\frac{4\omega_0}{GB}$	$\frac{4\omega_0}{GB}$
Soliman passive compensated non-canonic circuit [5]	1	4	4	$-\frac{\omega_0}{GB}$	$\frac{\omega_0}{GB}$
New active compensated canonic circuit	3	4	2	$-\frac{\omega_0}{GB}$	$\frac{\omega_0}{GB}$

The proposed filter is very practical utilizing the modern low cost matched op amps integrated circuits which are currently available in dual and quad packages. It is worth noting that the configuration in Fig. 2 is very general and it may employ other passive RC bandpass circuits for N.

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Der aktuelle Tagungsbericht

International Symposium „Information Theory and Systems Theory in Digital Communications Technology", Berlin, Germany, Sept. 18—20, 1978.

Parallel to the data-processing area digital technology is becoming more and more important in the processing of video and sound signals as well as in transmission and data switching techniques. The basic sciences for digital communications technology are information and system theory. Therefore the NTG/FA 1 in cooperation with the Commission C of the U. R. S. I. Local Commission in the FRG and the German section of the IEEE and ETV Berlin sponsored an international meeting on the above-mentioned subject with the aim of showing and discussing the ways these theories can be used and their advantages. The focus was on the following topics at the meeting:

1. source models and source encoding
2. channel models and channel encoding
3. digital transmission
4. synchronization and stability of networks
5. optimization, simulation and networks

The scientific programm included 77 lectures held by 23 foreign and 51 German scientists. The large number of some 350 participants showed how relevant the subject is.

D. Wolf and *D. Preuss* dealt with statistic models of voice and video signals. Not only the source statistics but also the perception characteristics of the receiver determine the optimization and effectiveness of a coding procedure. A general lecture by *H. Marko* and a paper by *J. O. Limb* were dedicated to this topic. Models of visuel perception and their use for video coding were discussed in these contributions.

In the field of voice coding the contributions on subband coding by *J. E. Stjernvall* and *C. Mourikis* et al. as well as the paper on coding by *H. G. Fehn* and *P. Noll* represented a certain focal point.

As an example for the papers on video coding, reference should be made to the coding method of *J. Klie* with which for the first time a moving image transmission was demonstrated with the very low transmission bit rate of 64 kbit/s.

In a concluding lecture by *J. B. O'Neal* the main points involved in the mathematical treatment of source encoding with the methods of the rate-distortion theory were discussed.