Active Compensation of Opamps

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Abstract—**A novel active compensation method for the finite gain bandwidth of operational amplifiers, when used in voltage-controlled voltage source structures, is given. In this method, a universal active compensation scheme is proposed which can be used with operational amplifier voltage-controlled voltage sources, when looked upon as a general three-port active building block. It is shown that phase shifts contributed by imperfect operational amplifiers are virtually eliminated over an extended frequency range. Experimental results are presented.**

I. Introduction

It is well known that the finite and complex gain nature of the operational amplifiers (opamps) degrades the performance of RC-active filters significantly. Therefore, many authors have looked at this problem, see, e.g., [1] and [2]. They have proposed means to improve the performance of active networks with respect to the use of imperfect amplifiers.

In fact, the problem of high-frequency effects in opamp circuits can be handled via two main compensation methods namely: "passive compensation" and "active compensation."

The "passive compensation" method produces an amount of phase lead that compensates for the phase lag of the imperfect amplifiers by using some additional passive components [3] and [4]. These passive components have to be adjusted at specific ambient temperature and power supply voltages to match the opamp unity gain bandwidth according to some design constraints. This requires that the opamp unity gain bandwidth be precisely measured. If the ambient temperature or the power supply voltage is changed the compensation will no longer be satisfactory.

On the other hand, with the introduction of low-cost "dual" opamps having closely matched characteristics which track with changes in temperature and voltage, Bracket and Sedra have considered "active compensation" and have used this method with opamp integrators [5].

The object here is to describe an active compensation scheme that can be used with the generalized three-port opamp voltage-controlled voltage source (VCVS) active building blocks which find wide use in many applications.

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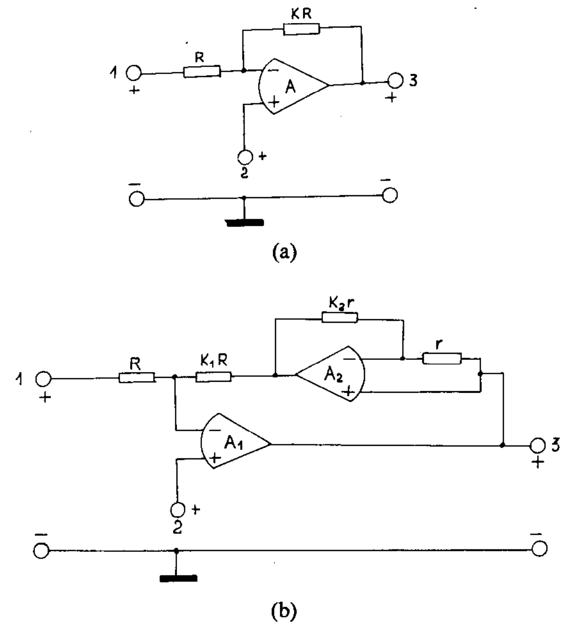

Fig. 1 (a) The uncompensated three-port opamp VCVS active building block. (b) The proposed compensated three-port opamp VCVS.

II. The Proposed Active Compensation Scheme

The three-port VCVS of Fig. 1(a) is generally used in one of the following ways in active-RC networks:

1) Port 1 is shorted to ground. This mode of operation is known as the noninverting VCVS.

2) Port 2 is shorted to ground. This mode of operation is known as the inverting VCVS.

3) Neither port 1 or 2 is shorted to ground. In this case it can be said that the VCVS appears in a double-input arrangement.

The opamp voltage follower of Fig. 2(a) can be considered as a special case of the double-input VCVS.

Let the open loop gain of the opamp be represented in the form

$$A(s)=\frac{\omega_t}{s+\omega_p}$$

with ω_t being the unity gain bandwidth.

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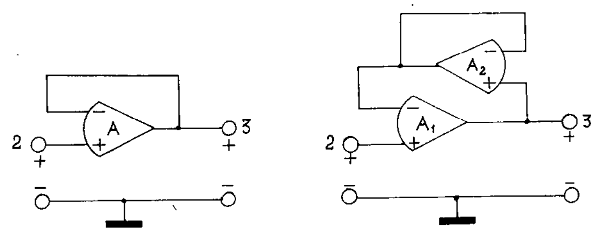


Fig. 2 (a) The uncompensated voltage follower. (b) The compensated voltage follower.

In the frequency range of interest, $\omega \gg \omega_p$, so

$$A(s)=\frac{\omega_t}{s}.$$

For the three-port VCVS of Fig. 1(a), the generalized uncompensated expression is

$$V_3=[(K+1)V_2-KV_1]\epsilon_1(s) \tag{1}$$

where

$$\epsilon_1(s)=\frac{1}{1+s\tau} \tag{2}$$

and

$$\tau=\frac{K+1}{\omega_t} \tag{3}$$

$\epsilon_1(s)$ is the error function contributed by the finite ω_t of the opamp. Ideally $\epsilon_1(s)$ must be frequency independent, i.e., it must have a unity magnitude and a zero phase. From (2) however it is seen that the phase and the magnitude of $\epsilon_1(s)$ are given by

$$\left.\begin{aligned}&\arg[\epsilon_1(j\omega)]\simeq-\omega\tau\\ \text{and}\quad&|\epsilon_1(j\omega)|\simeq 1-\tfrac{1}{2}\omega^2\tau^2\end{aligned}\right\}\ \omega\tau\ll 1. \tag{4}$$

The above expressions indicate that the magnitude error is a second-order term equal to $-\frac{1}{2}\omega^2\tau^2$, whereas the phase error is of first-order magnitude is equal to $-\omega\tau$. In other words, the VCVS's structures require only phase compensation.

For the proposed compensated three-port VCVS of Fig. 1(b), the generalized compensated expression is

$$V_3=[(K_1+1)V_2-K_1V_1]\epsilon_2(s) \tag{5}$$

where

$$\epsilon_2(s)=\frac{1+s\tau_2}{1+s\tau_1+s^2\tau_1\tau_2} \tag{6}$$

and

$$\tau_i=\frac{K_i+1}{\omega_{t_i}}\qquad (i=1,2) \tag{7}$$

$\epsilon_2(s)$ is the remaining error of the new compensated circuit of Fig. 1(b). Its phase is expressed as

$$\arg[\epsilon_2(j\omega)]\simeq\omega\tau_2-\omega\tau_1-\omega^3\tau_1^2\tau_2,\qquad \omega\tau_i\ll 1\quad (i=1,2). \tag{8}$$

By examining the preceeding equation for the remaining phase error it is seen that the order of the first two terms is much larger than that of the last term. Therefore, it is clear that by taking $\tau_1=\tau_2=\tau$ will yield relatively negligible phase error $(-\omega^3\tau^3)$ over a prescribed frequency range. Equations (5) and (6) reduce to

$$V_3=[(K+1)V_2-KV_1]\epsilon_3(s) \tag{9}$$

where

$$\epsilon_3(s)=\frac{1+s\tau}{1+s\tau+s^2\tau^2} \tag{10}$$

and

$$\tau=\frac{K+1}{\omega_t}. \tag{11}$$

The magnitude of $\epsilon_3(s)$ is expressed as

$$|\epsilon_3(j\omega)|\simeq 1+\omega^2\tau^2,\qquad \omega\tau\ll 1. \tag{12}$$

Thus with $\tau_1=\tau_2=\tau$ the phase error is reduced to a negligible level at the expense of a simple doubling $(\omega^2\tau^2)$ of the normally very low magnitude error.

The efficiency of the proposed compensation technique is comparatively dependent upon the realized parameter K and the used opamp ω_t (since it is assumed that $\omega\tau=(\omega(K+1)/\omega_t)\ll 1$, throughout the above analysis).

Therefore, it is of importance to clarify to what extent this dependence is strong. Fig. 3 shows a set of calculated curves for $|\epsilon_3(j\omega)|$ and $\arg[\epsilon_3(j\omega)]$ plotted versus ω/ω_t with K as a parameter without any approximations. This figure indicates the remaining error of the new compensated circuit as a function of the used opamp ω_t and of the realized K.

It is worth noting that the error function $\epsilon_3(s)$ was obtained by Wilson in [3] for compensation of opamp VCVS's using "passive compensation" techniques (by adding a single capacitor for the noninverting structure and two capacitors for the inverting one).

A. The Compensated Voltage Follower

Now, it is clear that the compensated inverting VCVS and the compensated noninverting VCVS are obtained by shorting to ground port 2 or port 1 of Fig. 1(b), respectively.

As for the voltage follower of Fig. 2(a), which is a positive unity voltage gain amplifier, if K is set equal to zero in (9), the following transfer function is obtained:

$$\frac{V_3}{V_2}=\epsilon_{30}(s)=\frac{1+(s/\omega_t)}{1+(s/\omega_t)+(s^2/\omega_t^2)} \tag{13}$$

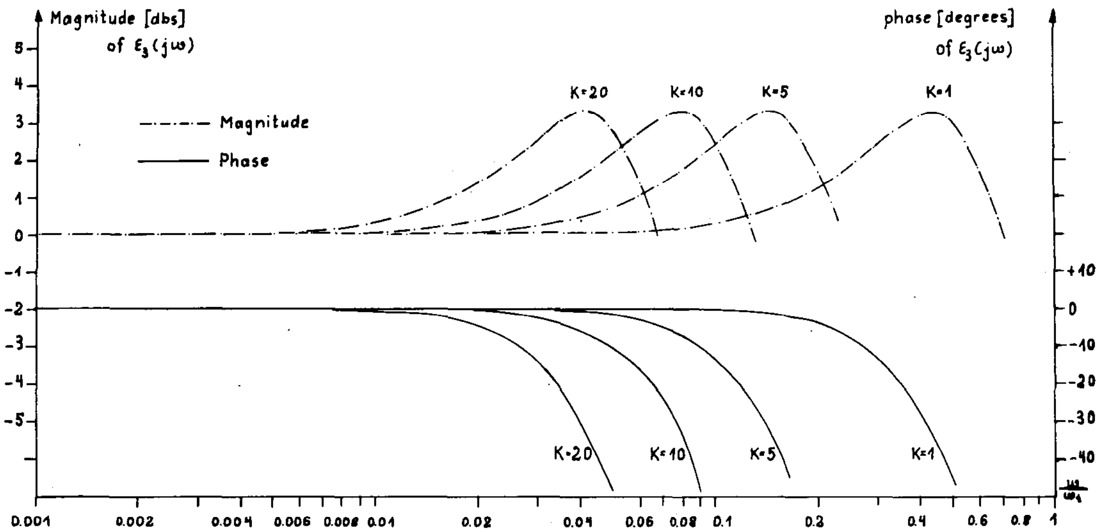


Fig. 3 Calculated magnitude and phase characteristics of $\epsilon_3(j\omega)$ plotted versus ω/ω_t, with K taken as a parameter.

where $\epsilon_{30}(s)$ is the remaining error function $\epsilon_3(s)$ when K is set equal to zero, i.e., with $\tau = 1/\omega_t$.

The above transfer function has simply an ideal gain factor of unity. Therefore, it represents the transfer function of the compensated voltage follower.

However, $K=0$ in (9) means that in Fig. 1(b) the resistors K_1R and K_2r are short circuited. This will make the transfer function between V_2 and V_3 independent of the remaining two resistors r and R. Then, r and R are taken open circuited and this results in the compensated follower of Fig. 2(b). In other words, the voltage follower of Fig. 2(a) needs simply another voltage follower in the feedback path to improve its high-frequency properties. It is required only that the two opamps used (A_1 and A_2) must have identical unity gain bandwidth ω_t.

III. Experimental Results

The compensation technique described above has been applied to the building block of Fig. 1(a) with port 2 shorted to ground to realize an inverting VCVS taken as an example.

First, the magnitude and the phase characteristics of the uncompensated inverting VCVS designed for a gain of 20 dB ($K=10$) have been measured. Precise resistors of $R=1.8$ kΩ and $KR=18$ kΩ have been used. Then, the magnitude and the phase characteristics of the compensated inverting VCVS (Fig. 1(b)) with port 2 shorted to ground and with $K_1=K_2=10$) have been measured. Precise resistors of $r=R=1.8$ kΩ and $K_2r=K_1R=18$ kΩ have been used.

For carrying out the experiment above a ± 15-V dual opamp type ML 747 (Microsystems International Ltd) has been used which has a unity gain bandwidth of $\omega_t/2\pi = 800$ kHz.

Fig. 4 shows plots for the magnitude and the phase characteristics of the compensated and the uncompensated inverting VCVS's, as well as the corresponding theoretical characteristics (calculated from (1) and (9) with $V_2=0$, $K=10$, and $\omega_t/2\pi=800$ kHz). This figure demonstrates to what extent the useful frequency range is improved.

It is worth noting that the absolute values of r and R are arbitrary since the compensation technique requires only that $K_1=K_2$. However, it is better to take $r=R$ in order to limit the spread of resistors values.

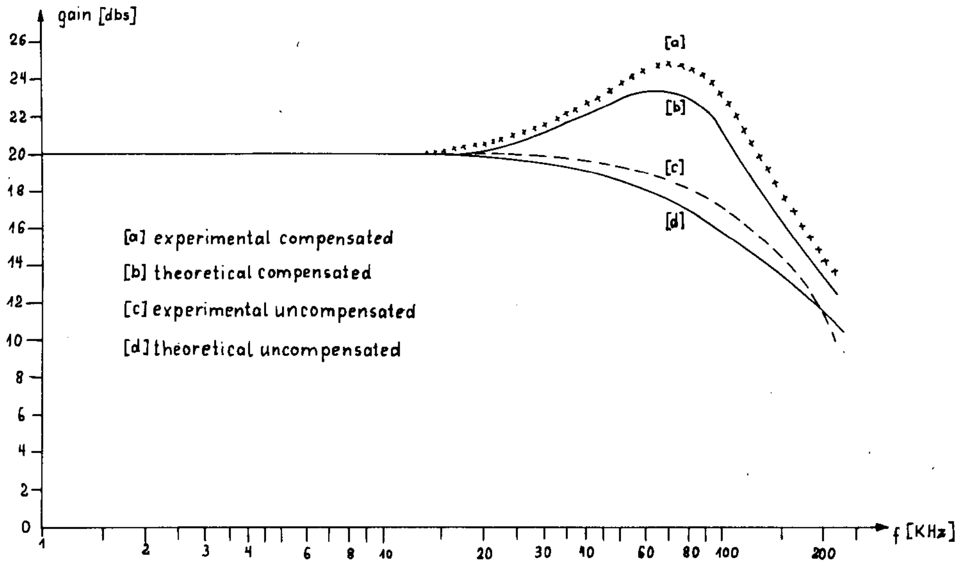
A. Application of the Proposed Compensated VCVS's

The proposed compensation scheme has been applied to the second-order Wilson's all-pass filter section [6] of Fig. 5(a) in which a VCVS appears in a double-input arrangement (nodes N_1 and N_2 are nongrounded). This all-pass filter section has the advantage of having a unity gain constant. The filter section of Fig. 5(a) was built in the laboratory (designed for a selectivity and natural frequency of 5 and 22.607 kHz, respectively). First, the phase frequency response of the circuit of Fig. 5(a) (the uncompensated performance) was measured. Then, the modified circuit of Fig. 5(b) was built in which the voltage follower is replaced by the compensated voltage follower of Fig. 2(b) and the double-input VCVS is replaced by its compensated scheme. Then, the phase frequency response of the circuit of Fig. 5(b) (the compensated performance) was also measured.

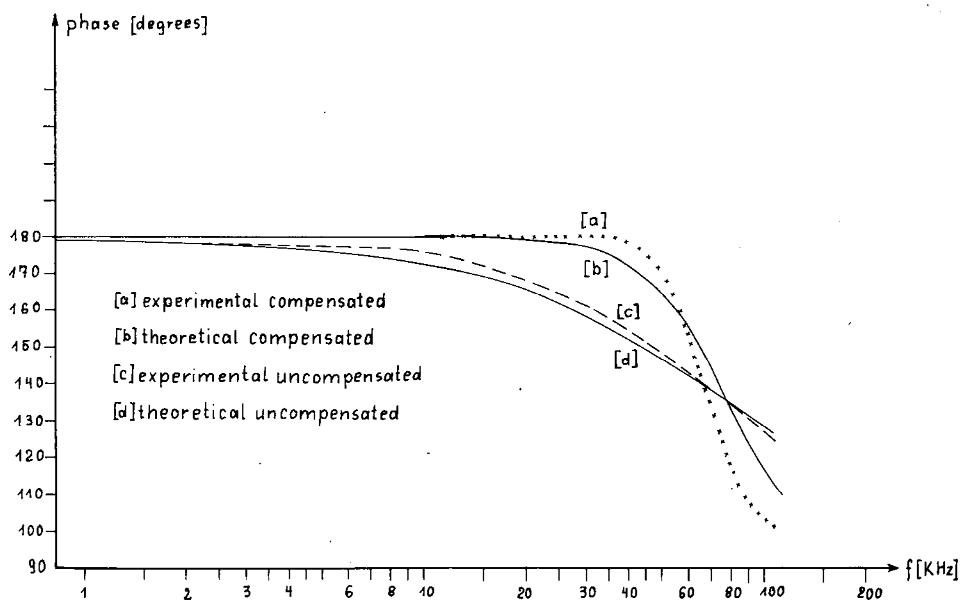
Fig. 6 shows plots for the uncompensated, the compensated, and the theoretical phase performances.

Close agreement is clearly observed between the compensated and the theoretical performance (natural frequency error of 0.19 percent). The uncompensated performance is clearly shifted from the theoretical performance (natural frequency error of -10.028 percent).

Experimental observations showed also that the voltage



(a)



(b)

Fig. 4 (a) Gain characteristics of an inverting VCVS (designed for $K=10$). (b) Phase characteristics of an inverting VCVS (designed for $K=10$).

gain of both the compensated circuit and the uncompensated one was approximately constant at the value of unity.

For carrying out the preceeding experiment, the following component list was used:

$\pm$ 15-V dual opamps type ML 747 (Microsystems International LTD) having a unity gain bandwidth of $\omega_t/2\pi=$ 800 kHz.

$$C=2.2 \text{ nF} \qquad R=3.2 \text{ k}\Omega$$

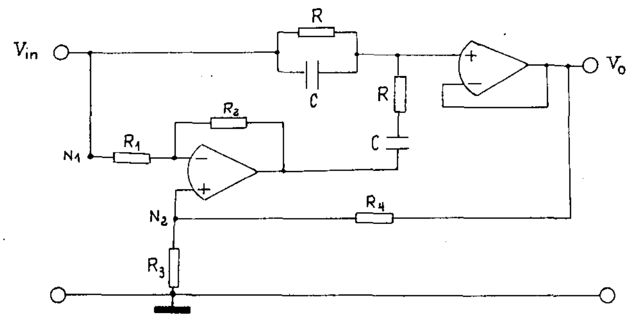
$$R_1=1 \text{ k}\Omega \qquad R_2=2.2 \text{ k}\Omega$$

$$R_3=7 \text{ k}\Omega \qquad R_4=1 \text{ k}\Omega\,.$$

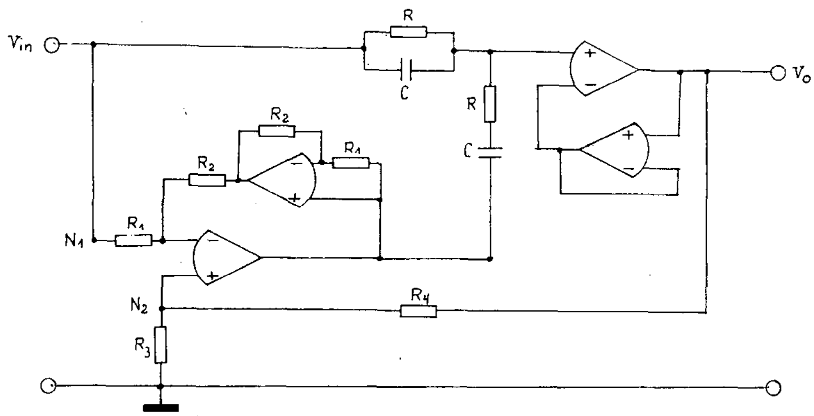
Five-percent capacitors and precise resistors were used.

IV. Conclusions

Main advantages of "active compensation" method for the finite gain bandwidth of opamps over that of "passive compensation" have been surveyed. A universal active compensation building block for opamp VCVS structures



(a)



(b)

Fig. 5 (a) Wilson's all-pass filter section. (b) The modified all-pass filter section.

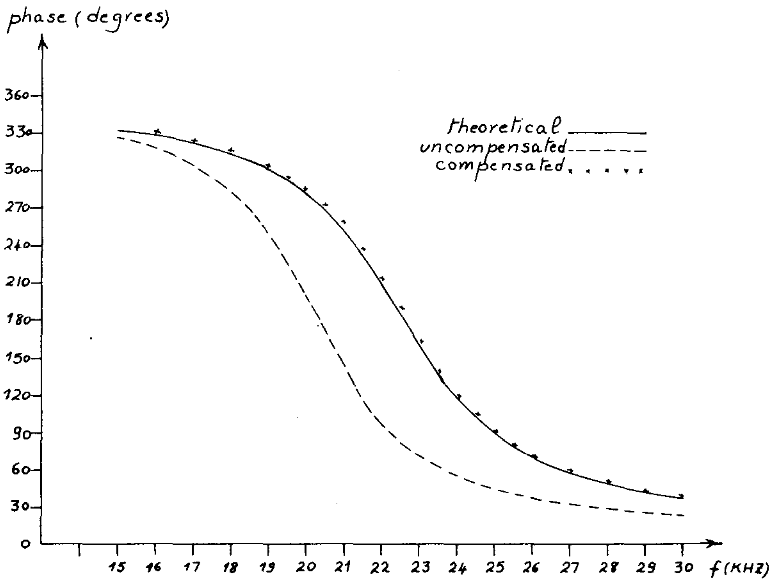


Fig. 6 Phase frequency response of the all-pass filters of Fig. 5(a) and (b).

is given in which a dual opamp is used without any design constraints. The universal building block finds wide use in active-RC networks specially those in which the VCVS appears in a double-input arrangement. Experimental results are included.

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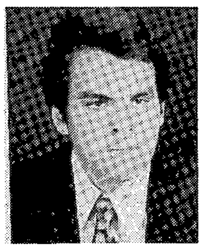

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Sequence Discriminators and Their Use in Frequency Division Multiplex-Communication Systems

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***Abstract*—A simple theory of networks exhibiting sequence discrimination and their use for producing quadrature signals is first presented. A channel unit employing sequence discriminators (SD's), namely, passive (*RC*), passive with feedback, and active designs are discussed. A channel unit employing SD's and the associated modulation and demodulation schemes are suggested. Statistical sensitivity results and experimental results using thick film resistor and NPO chip capacitor implementations are also given. These indicate that the (SD) approach for single-sideband generation and detection is very promising in FDM applications. This technique which uses resistors, capacitors, and commercial integrated circuits is simple to implement and requires no specially developed technologies.**

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I. Introduction

SEVERAL basic contributions have been reported for the transfer function approximation as well as the realization of 90° phase shift difference networks [1]–[12]. Published literature indicated that this approach has potential applications in the instrumentation as well as the communication area [2], [8]–[12].

The objective is to obtain, from the input signal, two signals V_1 and V_2 to be ideally equal in magnitude and different in phase from one another by 90°. These properties of V_1 and V_2 are required only over the frequency range of interest. Most of the work started by restricting V_1 and V_2 to be equal in magnitude at all frequencies. Consequently, the signals V_1 and V_2 were produced by

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