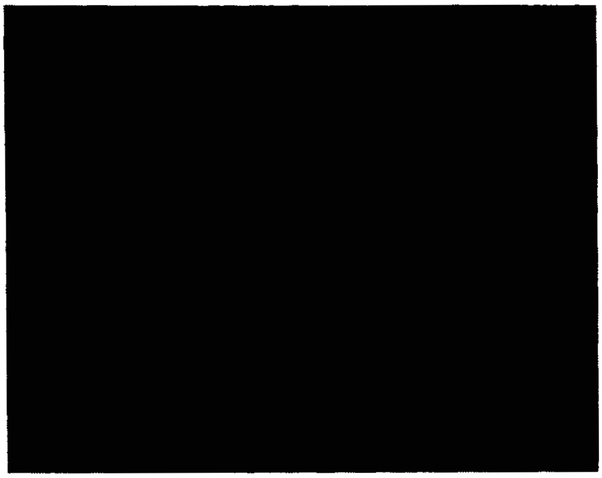

Fig. 4. Experimental results of Fig. 3(b). Horizontal scale: 5-kHz/div. Vertical scale: 2.3-dB/div.

this notch filter is given by

$$H_N(s) = \frac{s^2 + 1/Lc_2}{s^2 + s/R_e c_2 + 1/Lc_2} \tag{7}$$

where $R_e = R_L // R$. Using the switched-capacitor circuit of Fig. 1 to replace L, and replacing R_e by a resistor equivalent [5] switched-capacitor c_L, we obtain the switched-capacitor notch filter shown in Fig. 3(b). The following z-transfer function can be obtained using the switched-capacitor transformation [1], [6], the resistor equivalents [5], and by assuming that $R > R_L$ ($c_L > c$)

$$H_N(z) = \frac{(1 - z^{-2})^2 + A_0 cc_1/c_0 c_2}{(1 - z^{-2})^2 + \frac{c_L}{c_2}(1 - z^{-2}) + \frac{A_0 cc_1}{c_0 c_2}}. \tag{8}$$

IV. Experimental Results

A notch filter with $f_0 \cong 12.5$ kHz and $Q = 23$ was designed using two CA3140 op amps with measured GB products $(A_0\omega_1)_1 = 2\pi \times 3.88 \times 10^6$ rad s^{-1}, $(A_0\omega_1)_2 = 2\pi \times 3.72 \times 10^6$ rad s^{-1}, and $f_s = 100$ kHz. The component values used were $c = 149$ pF, $c' = 154$ pF, $c_2 = 0.09$ μF, and $c_L = 0.06$ μF. The measured notch characteristic using these components is shown in Fig. 4, for an input voltage of 16 mV.[1] The small differences (about 4 percent) between the experimental results and the predicted response are due basically to the effects of the excess phase [7] of the op amps and the parasitic capacitances.

Acknowledgment

The authors are indebted to Dr. P. C. T. Roberts for helpful technical assistance.

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[1] The spectrum analyzer is an HP3580A with the input sensitivity adjusted to give the display in the most convenient form and, hence, the vertical scale is 2.3 dB/div.

Novel Generalized Differential Integrator with Controlled Phase Lead

AHMED M. SOLIMAN

Abstract—**A new active compensation method for the differential integrator is given. The proposed integrator has a phase lead which is controlled by a single resistor. The circuit realizes a novel phase lead inverting (noninverting) integrator by shorting one of the two input ports to ground. A differential integrator with a very small phase lag which includes the compensated Miller [1] and the compensated balanced time constant [2] integrators as special cases is also given.**

I. Introduction

Several active compensation methods for the inverting and noninverting integrators to reduce the effect of the finite unity-gain bandwidth of the op amp have been reported [1]-[4]. The op amp integrators available up to date may be classified as follows.

1) The Phase Lag Inverting Integrator: This includes the well-known Miller integrator and its compensated version which employs a voltage follower in the feedback path [1], [4]. Another integrator which belongs to this class has been reported by Reddy [5] for the purpose of magnitude correction in a double integrator loop.

2) The Phase Lag Noninverting Integrator: This includes the Miller-inverter cascade, the modified Miller-inverter cascade [4], the Deboo integrator [6] as well as its compensated versions reported most recently [2], [3], and the balanced time constant integrator and its modified version [2] which employs a voltage follower in the feedback path.

3) The Phase Lead Noninverting Integrator: This includes the Vogel [1], Akerberg-Mossberg [7], and the Reddy [5] integrator. They are used together with the phase lag inverting integrators to realize good quality double integrator filters.

From the above classification it is clear that there is a need for inverting integrators with a phase lead, so that they may be used together with the available phase lag noninverting integrators to realize excellent biquad filters. This is one of the objectives of this contribution.

II. The Uncompensated Differential Integrator

First a brief review of the effect of the finite unity-gain bandwidth of the op amp on the performance of the differential integrator is given. If the op amp is compensated to have a single-pole open-loop response with a unity-gain bandwidth ω_t, then its gain is approximately given by

$$A(s) \simeq \frac{\omega_t}{s}. \tag{1}$$

Thus by analysis to the circuit in Fig. 1,

$$V_0 = (V_2 - V_1)\frac{\omega_0}{s}\epsilon(s) \tag{2}$$

where

$$\omega_0 = \frac{1}{CR} \tag{3}$$

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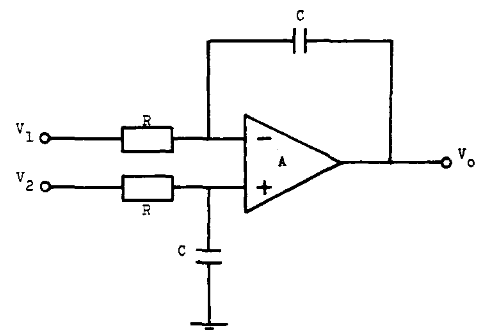

Fig. 1. The uncompensated differential integrator.

and

$$\epsilon(s) = \frac{1}{(1 + \omega_0/\omega_t) + s/\omega_t}. \tag{4}$$

$\epsilon(s)$ is the error function contributed by the finite ω_t of the op amp. From (4) it is seen that the phase shift due to the finite ω_t is

$$\phi \simeq -\frac{\omega}{\omega_t} = -\frac{1}{|A(j\omega)|}, \quad \omega_t >> \omega_0. \tag{5}$$

III. The Phase Lead Differential Integrator

Fig. 2 represents the proposed active compensated differential integrator. The circuit uses the active building block used before for the compensation of the voltage-controlled voltage-source (VCVS) structures [8].

The generalized compensated response for the circuit in Fig. 2 is

$$V_0 = (V_2 - V_1)\frac{\omega_0}{s}\epsilon_m(s) \tag{6}$$

where

$$\epsilon_m(s) = \frac{1 + (K+1)/A_1}{1 + (1/A)(1 + \omega_0/s)(1 + (K+1)/A_1)}. \tag{7}$$

Thus assuming matched op amps are used (available now at low cost in dual packages),

$$\epsilon_m(s) = \frac{1 + s/\omega_t (K+1)}{(1 + \omega_0/\omega_t) + (s/\omega_t)(1 + (\omega_0/\omega_t)(K+1)) + (s/\omega_t)^2 (K+1)}. \tag{8}$$

The phase shift in this case is leading and is given by

$$\phi_m \simeq \frac{K\omega}{\omega_t} = \frac{K}{|A(j\omega)|}, \quad \omega_t >> (K+1)\omega_0. \tag{9}$$

The resistor KR_1 controls the amount of phase lead obtained. This phase lead differential integrator may also be used in one of the following special cases.

1) Inverting Phase Lead Integrator: This is obtained by shorting port 2 to ground as shown in Fig. 3. This is the first available inverting integrator with a controlled amount of phase lead up to this author's knowledge. Applications of this integrator in the realization of filters with zero Q sensitivity to the op amp unity-gain bandwidth will be reported in a future article.

2) Noninverting Phase Lead Integrator: This is obtained by shorting port 1 to ground. In the special case of $K = 1$, the realizable integrator is equivalent to the well-known phase lead noninverting integrator [4], [7].

IV. The Compensated Differential Integrator with Reduced Phase Lag

From (8) if K is set to zero, the compensated error function reduces to

$$\epsilon_0(s) = \frac{1 + s/\omega_t}{(1 + \omega_0/\omega_t) + s/\omega_t(1 + \omega_0/\omega_t) + (s/\omega_t)^2}. \tag{10}$$

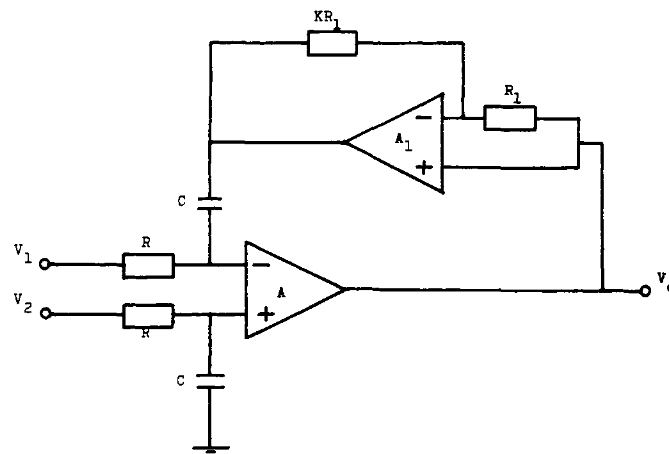

Fig. 2. The proposed phase lead differential integrator.

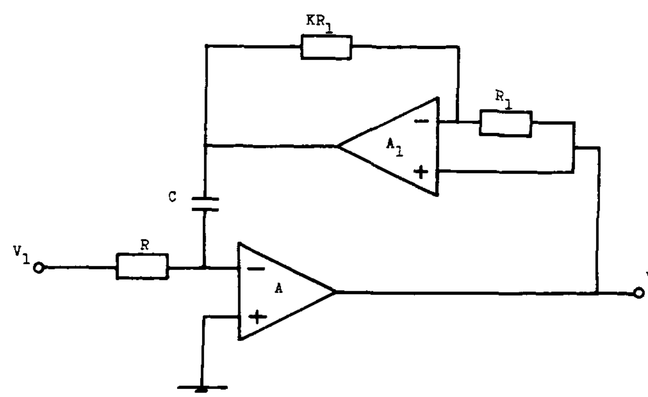

Fig. 3. The novel phase lead inverting integrator.

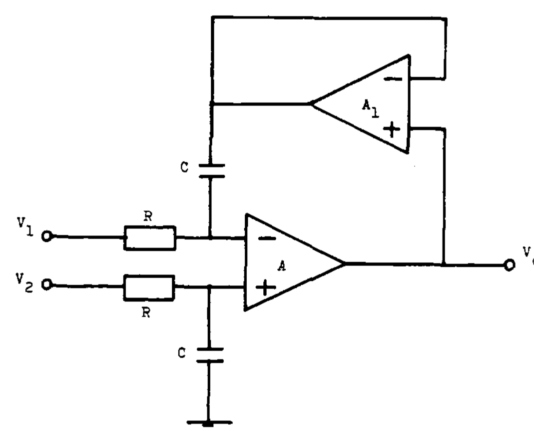

Fig. 4. The active compensated differential integrator with a small phase lag.

The phase shift in this case is

$$\phi_0 \simeq -\left(\frac{\omega}{\omega_t}\right)^3 = \frac{-1}{|A(j\omega)|^3}, \quad \omega_t >> \omega_0. \tag{11}$$

Comparing (5) and (11) the reduction in the phase lag is evident. Setting $K = 0$, however, implies that the resistor KR_1 is shorted and the response in this case is independent of the remaining resistor R_1. Thus R_1 is taken as open circuit and the circuit reduces to that in Fig. 4, which employs a voltage follower for compensation. It is worth noting that this compensated differential integrator includes the compensated Miller [1], [4] and the compensated balanced time constant [2] integrators as special cases by shorting to ground port 2 or port 1, respectively.

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Active Phase Compensation of Op Amp VCCS Structures

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***Abstract*—A generalized active compensated three-port voltage controlled current source (VCCS) is proposed. The compensation is affected by adding to the uncompensated VCCS another op amp and two resistors. The compensated VCCS (which may be used as an inverting or a noninverting VCCS) has a negligible phase error and an improved bandwidth.**

I. Introduction

Recently, an active compensation method for the finite unity gain bandwidth of op amps when used in voltage-controlled voltage-source (VCVS) structures has been reported [1].

The three-port voltage-controlled current-source (VCCS) shown in Fig. 1 has been introduced in the literature [2] as a generalization of the well-known inverting VCCS [3]-[4]. This generalized three-port VCCS can be used as an inverting VCCS by shorting port 2 to ground or as a noninverting VCCS by shorting port 1 to ground.

The purpose of this contribution is to demonstrate the applicability of the recently described method of phase compensation [1] to the VCCS structures.

II. The Uncompensated VCCS

First a brief analysis of the effect of the finite unity gain bandwidth ω_t of the op amp on the performance of the VCCS is given. Let the open loop gain of the op amp be represented by the single pole model; thus

$$A(s) \simeq \frac{\omega_t}{s}. \tag{1}$$

By direct analysis to the circuit in Fig. 1, the load current is given by

$$I_L = -\frac{V_1}{R}\epsilon_1(s) + \frac{V_2}{R}\epsilon_2(s) \tag{2}$$

where

$$\epsilon_1(s) = \frac{1}{1 + (2/\omega_t)(1+x)s} \tag{3}$$

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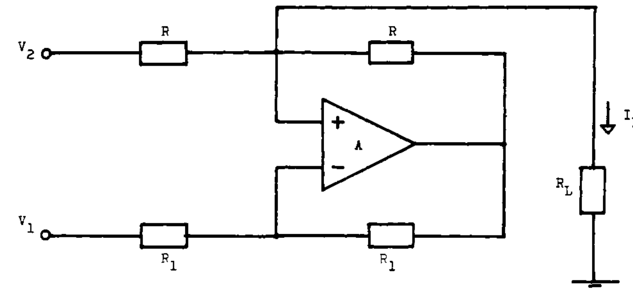

Fig. 1. The generalized uncompensated VCCS.

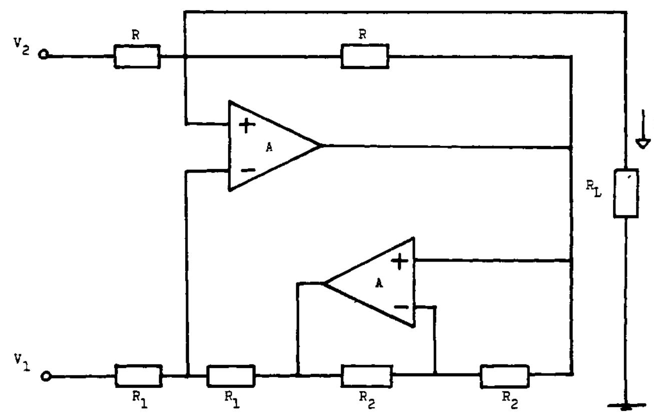

Fig. 2. The new generalized active compensated VCCS, using low cost dual op amps.

$$\epsilon_2(s) = \frac{1 + (2/\omega_t)s}{1 + (2/\omega_t)(1+x)s} \tag{4}$$

and

$$x = \frac{2R_L}{R}. \tag{5}$$

$\epsilon_1(s)$ and $\epsilon_2(s)$ are the inverting and noninverting error functions contributed by the finite ω_t of the op amp. Ideally these error functions must have a unity magnitude and a zero phase. From (3), however, it is seen that the phase error and the magnitude error of the inverting VCCS are given by

$$\left.\begin{aligned} \phi_1 &\equiv \arg \epsilon_1(j\omega) \simeq -\left(\frac{2\omega}{\omega_t}\right)(1+x) \\ \gamma_1 &\equiv |\epsilon_1(j\omega)| - 1 \simeq -2\left(\frac{\omega}{\omega_t}\right)^2 (1+x)^2 \end{aligned}\right\} \frac{2\omega}{\omega_t} << 1. \tag{6}$$

Similarly from (4) for the noninverting VCCS it is seen that

$$\left.\begin{aligned} \phi_2 &\equiv \arg \epsilon_2(j\omega) \simeq -\left(\frac{2\omega}{\omega_t}\right)x \\ \gamma_2 &\equiv |\epsilon_2(j\omega)| - 1 \simeq -2\left(\frac{\omega}{\omega_t}\right)^2 [(1+x)^2 - 1] \end{aligned}\right\} \frac{2\omega}{\omega_t} << 1. \tag{7}$$

The above expressions indicate that the magnitude error is a second-order term, whereas the phase error is of a first-order magnitude. Thus the VCCS structures require only phase compensation.

III. The Active Compensated VCCS

The modern low cost matched op amp integrated circuits which are currently available in dual packages are used here for active phase compensation of the VCCS structures. Fig. 2 represents the new active