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A NEW ACTIVE COMPENSATED DIFFERENTIAL INTEGRATOR WITHOUT MATCHED OPERATIONAL AMPLIFIERS

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INTRODUCTION

The single operational amplifier (opamp) differential integrator[1] finds wide use in many circuit applications. It is well known that the finite and complex gain nature of the opamp degrades significantly the performance of the differential integrator.[2] Recently, an active compensated differential integrator has been reported in the literature.[2] In this paper a novel active phase compensated differential integrator is introduced. The proposed network has the advantage that the phase compensation condition is independent of the gain bandwidth of the opamps employed in the circuit. Experimental results are included.

NEW CIRCUIT

The proposed phase compensated differential integrator is shown in Figure 1. Straighforward analysis of the circuit yields the following expression for the output voltage V_0 in terms of the inverting input voltage V_1 and the non-inverting input voltage V_2.

$$V_0 = \frac{1}{sCR}\,\frac{-V_1\left[1+\dfrac{K_1(K_2+1)(1+sCR)}{(K_1+1)A_2}\right]+V_2\left[1+K_2\dfrac{(1+sCR)}{A_2}\right]}{1+\left[\dfrac{(K_2+1)}{(K_1+1)}\right]\left[\dfrac{1}{A_2}+\dfrac{K_1+1}{A_1A_2}\right]\left[1+\dfrac{1}{sCR}\right]} \tag{1}$$

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Received 9 December 1981 *Revised 1 June 1983*

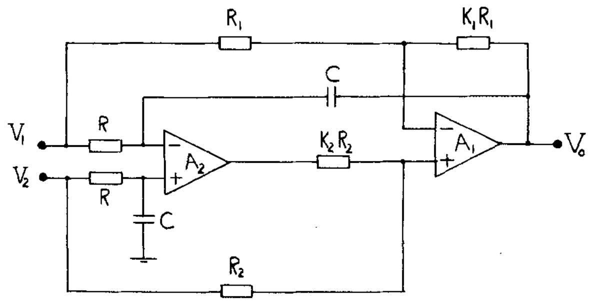

Figure 1. The proposed active-compensated differential integrator

From the above equation it is seen that for the differential mode of operation it is necessary to have

$$K_1 = K_2 \tag{2}$$

In this case equation (1) simplies to:

$$V_0 = \frac{\omega_0}{s}(V_2 - V_1)\varepsilon(s) \tag{3}$$

where

$$\omega_0 = \frac{1}{CR} \tag{4}$$

$$\varepsilon(s) = \frac{1 + \dfrac{K_1}{A_2} + \dfrac{K_1 s}{\omega_0 A_2}}{1 + \dfrac{1}{A_2} + \dfrac{(K_1+1)}{A_1 A_2} + \dfrac{\omega_0}{sA_2} + \dfrac{(K_1+1)\omega_0}{sA_1A_2}} \tag{5}$$

$\varepsilon(s)$ is the error function of the compensated circuit.

Let the open loop gain of the two opamps be characterized by the single pole model given by

$$A_i(s) \simeq \frac{\omega_{ti}}{s}, \qquad (i = 1, 2) \tag{6}$$

where ω_t is the unity gain bandwidth of the opamp.

Using (6) in (5), one has

$$\varepsilon(s) \simeq \frac{1 + \dfrac{K_1}{\omega_{t2}} s + \dfrac{K_1}{\omega_0 \omega_{t2}} s^2}{1 + \dfrac{1}{\omega_{t2}} s + \dfrac{(K_1+1)}{\omega_{t1}\omega_{t2}} s^2} \qquad \begin{matrix} \omega_0 \ll \omega_{t_2} \\ (K_1+1)\omega_0 \ll \omega_{t1} \end{matrix} \tag{7}$$

From the above equation, it is seen that the phase compensation condition is given by

$$K_1 = 1 \tag{8}$$

That is, the phase compensation condition is independent of the ω_t of the opamps used, and it is not necessary to use matched opamps with this differential integrator.

The compensated error function reduces to:

$$\varepsilon_c(s) \simeq \frac{1+\dfrac{s}{\omega_{t2}}+\dfrac{s^2}{\omega_0\omega_{t2}}}{1+\dfrac{s}{\omega_{t2}}+\dfrac{2s^2}{\omega_{t1}\omega_{t2}}} \tag{9}$$

The remaining phase error of the compensated differential integrator is given by

$$\phi_c(\omega_0) \simeq \left(\frac{\omega_0}{\omega_{t2}}\right)^2 \simeq \left|\frac{1}{A_2(j\omega_0)}\right|^2, \qquad 2\omega_0 \ll \omega_{t_1} \tag{10}$$

Comparing the above equation with that of the uncompensated differential integrator which is given by[2]

$$\phi(\omega_0) \simeq -\frac{1}{|A(j\omega_0)|}, \qquad \omega_0 \ll \omega_t \tag{11}$$

it is seen that the reduction in the phase error is quite evident.

STABILITY ANALYSIS

Taking the second opamp pole into consideration and assuming that it occurs at a frequency ω_2 ($\omega_2 > \omega_t$), the open loop gain can be expressed as[3]

$$A(s) \simeq \frac{\omega_t}{s(1+s/\omega_2)} \tag{12}$$

Assuming that matched opamps are used and substituting for $A(s)$ from (12) into (5), it follows that a necessary condition for the stability of the differential integrator is given by:

$$\frac{\omega_2}{\omega_t} > 2 - \frac{1}{2(K_1+1)} \tag{13}$$

From equation (8), it is seen that the phase compensated differential integrator requires opamps having $\omega_2 > 1{\cdot}75\omega_t$.

EXPERIMENTAL RESULTS

A third-order lowpass Chebyshev filter is realized using the leap-frog[4] simulation of the passive prototype of Figure 2. The uncompensated active realization is shown in Figure 3. Note that the output circuitry reduces to the simple voltage follower A_3 to simulate a 1-Ω load resistor. The circuit is implemented using resistor tolerances of ±0·5 per cent and capacitor tolerances of ±0·2 per cent. LM 348 opamps are used. The opamps have a ratio ω_2/ω_t of 5·5 and ω_t of 1 MHz.

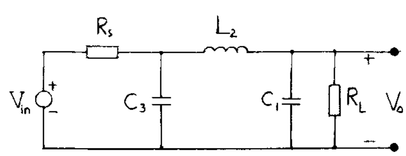

Figure 2. Passive prototype

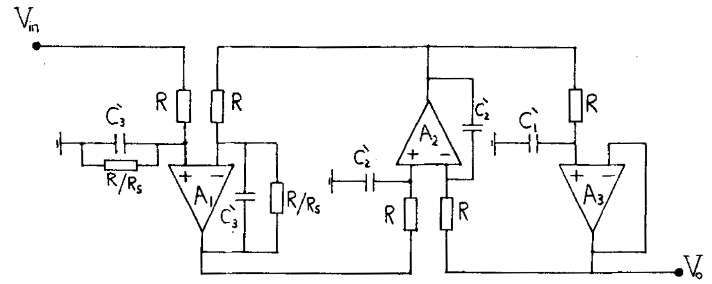

Figure 3. Leap-frog active realization

For the compensated case, it is found that one needs to alter only the circuitry of opamp A_2 as in Figure 1 in order to obtain a frequency response that is identical to the theoretical frequency response as shown in Figure 4.

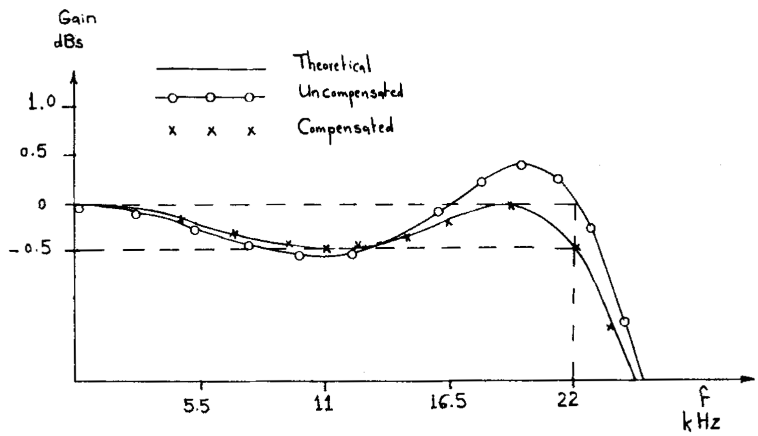

Figure 4. Experimental and theoretical frequency responses

CONCLUSIONS

A new active phase compensated differential integrator is given. To the authors' knowledge, this is the first active phase compensated differential integrator which has a phase compensation condition independent of the gain bandwidth of the opamps.

It is worth noting that the circuit is also suitable for realizing a phase lead differential integrator by taking $K_1 > 1$. In this case the phase lead is given by:

$$\phi(\omega) \simeq (K_1 - 1)\left(\frac{\omega}{\omega_{t2}}\right) \tag{14}$$

The new compensated circuit is tested in a third-order leap-frog realization. The resulting compensated performance is in agreement with the theory.

ACKNOWLEDGEMENT

The authors wish to thank K. M. Ansari and J. C. Muller of the Department of Electrical Engineering, University of Manitoba, for implementing the third order filter as part of their B.S. graduation thesis under the supervision of one of the authors (M. Ismail).

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ON THE SYNTHESIS OF A CLASS OF IMMITTANCES AND FILTERS USING GROUNDED CAPACITORS

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INTRODUCTION

It is known that active circuit structures employing grounded capacitors are suitable for IC implementation in monolithic[1] as well as in hybrid[2] IC fabrication techniques. In view of the current research interest towards microelectronic implementation of *RC* active filters,[3–5] the evolution of network topologies containing grounded capactiors is of contemporary relevance. The present letter shows that using operational transconductance amplifiers (OTA) or DVCCS[6] as active elements, it is possible to evolve active circuit structures for a class of immittances and filters, employing 'all grounded capacitors'; a feature which is not available in the corresponding operational-amplifier based methods/circuits presently known.

NEW IMMITTANCE SIMULATION CIRCUITS

We first present a new active realization of the floating FDNC ($Z(s) = Ms^2$) shown here in Figure 1. All other immittance simulation circuits described in this paper are generated from this circuit. Recall that the usual method of implementing a floating FDNR[7] ($Z(s) = 1/Ds^2$) or FDNC is to use a cascade connection of two generalized impedance converters (GIC) with a resistor embedded between them,[8] resulting in a four op-amp circuit employing four 'floating' capacitors. Such a circuit has the obvious drawback of being 'non-canonic', and an excessive number of capacitors would be required if such a circuit were to be used in active simulation of higher order passive *LC* filters. By contrast, the new circuit [which also uses only four active elements (OTAs)] requires only two capacitors and is hence 'canonic'. Analysis of the circuit of Figure 1 shows that the **Y**-matrix of this circuit is given by

$$[\mathbf{Y}] = \frac{G_1G_2G_3}{s^2C_1C_2}\begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix} \tag{1}$$

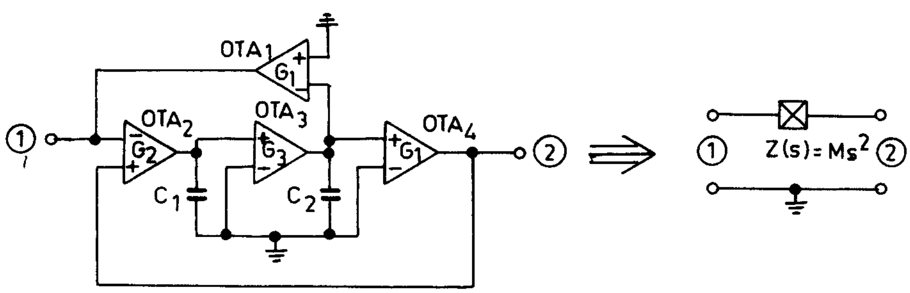

Figure 1. The new circuit for simulating a floating FDNC ($Z(s) = Ms^2$; $M = C_1C_2/G_1G_2G_3$)

Received 18 October 1982 *Revised 14 March 1983*