Novel Variable Phase Inverting Integrator

Ein neuer veränderlicher phaseninvertierender Integrator

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Abstract:

A new variable phase inverting integrator is given. The proposed integrator has the advantage that its phase depends on the gain-bandwidth of only one of the two operational amplifiers employed in the circuit. Application of the integrator for phase correction in the Tow-Thomas biquad and the Kerwin-Huelsman-Newcomb biquad is discussed.

Übersicht:

Es wird ein neuer veränderlicher phaseninvertierender Integrator beschrieben. Dieser hat den Vorteil, daß seine Phase vom Verstärkungs-Bandbreite-Produkt nur eines der beiden in der Schaltung verwendeten Operationsverstärker abhängt. Die Anwendung des Integrators zur Phasenkorrektur in der Tow-Thomas- und in der Kerwin-Huelsman-Newcomb-Filterstufe wird diskutiert.

Für die Dokumentation:

Operationsverstärker / Integrator / aktives Filter / Phasenkorrektur

1. Introduction

Several active compensation methods for realizing inverting integrators suitable for applications at high frequencies have been reported [1–5]. The inverting integrators reported in References [1, 2] are suitable for realizing an adjustable phase. The inverting integrators in [3–5] have the advantage that, the phase compensation condition is independent of the gain-bandwidth of the operational amplifiers (opamps) employed in the circuit. On the other hand, the integrator circuits in [3–5] cannot provide phase lead.

The purpose of this paper is to introduce a variable phase inverting integrator, having a phase which depends on the gain-bandwidth of only one of the two opamps employed in the circuit.

2. The New Integrator

The proposed variable phase inverting integrator is shown in **Fig. 1.** Analysis of the circuit yields the following transfer function:

$$\frac{V_o}{V_i}=-\left(\frac{1}{sCR}\right)\cdot\frac{1+K\left(\frac{a+1}{K+1}\right)\left(\frac{1+sCR}{A_2}\right)}{1+\left(\frac{a+1}{K+1}\right)\frac{1}{A_2}\left(1+\frac{1}{sCR}\right)+\frac{(a+1)}{A_1A_2}\left(1+\frac{1}{sCR}\right)}. \quad (1)$$

Assume that each opamp is characterized by a single pole model with a unity-gain bandwidth ω_t. Thus the open-loop gain A of the opamp is given by

$$A(s)\simeq\frac{\omega_{ti}}{s}\quad (i=1,2). \quad (2)$$

From equation (2) in equation (1) therefore

$$\frac{V_o}{V_i}=-\left(\frac{\omega_o}{s}\right)\cdot\varepsilon(s) \quad (3)$$

where

$$\omega_o=\frac{1}{CR} \quad (4)$$

$$\varepsilon(s)=\frac{1+\frac{K(a+1)}{(K+1)}\frac{s}{\omega_{t2}}+\frac{K(a+1)}{(K+1)}\frac{s^2}{\omega_o\,\omega_{t2}}}{1+\left(\frac{a+1}{K+1}\right)\frac{\omega_o}{\omega_{t2}}+\left(\frac{a+1}{K+1}\right)\frac{s}{\omega_{t2}}+(a+1)\frac{\omega_o s}{\omega_{t1}\omega_{t2}}+(a+1)\frac{s^2}{\omega_{t1}\omega_{t2}}}. \quad (5)$$

$\varepsilon(s)$ is the error function of the integrator. For frequencies such that $\omega_o \ll \left(\frac{K+1}{a+1}\right)\omega_{t2}$ and $\omega_o \ll \frac{1}{(K+1)}\omega_{t1}$, the above equation reduces to

$$\varepsilon(s)\simeq\frac{1+\frac{K(a+1)}{(K+1)}\frac{s}{\omega_{t2}}+\frac{K(a+1)}{(K+1)}\frac{s^2}{\omega_o\omega_{t2}}}{1+\frac{(a+1)}{(K+1)}\frac{s}{\omega_{t2}}+(a+1)\frac{s^2}{\omega_{t1}\omega_{t2}}}. \quad (6)$$

The excess phase of the integrator is given by

$$\Phi\simeq(a+1)\left(\frac{K-1}{K+1}\right)\frac{\omega}{\omega_{t2}}. \quad (7)$$

Thus it is seen that the resistor KR_2 controls the phase of the integrator which can be made leading by taking $K>1$. The resistor aR_1 controls the stability of the integrator.

3. Stability Analysis

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

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$$A(s) \simeq \frac{\omega_t}{s\left(1+\frac{s}{\omega_2}\right)}. \tag{8}$$

Assuming matched opamps are used and substituting from equation (8) into equation (1) and after routine stability analysis, it follows that for a stable operation of the integrator it is necessary that

$$\frac{\omega_2}{\omega_t} > \frac{2(K+1)}{(a+1)} - \frac{1}{2(K+1)}. \tag{9}$$

In order to ensure the integrator stability, it is recommended to take the design value of the parameter a as

$$a = 2K+1. \tag{10}$$

4. Application in Two-Integrator Loop Filters

Recently it has been demonstrated [2], how to realize improved Tow-Thomas (TT) biquad and improved Kerwin-Huelsman-Newcomb (KHN) biquad filters [6] by replacing the uncompensated inverting integrator by a phase lead inverting integrator. The proposed phase lead integrator may also be used to realize improved TT and KHN biquad filters suitable for high Q and high frequency. The design value of the parameter K is obtained from equation (7) by adjusting the phase lead at resonance to $(3\,\omega_o/\omega_t)$ which is the amount necessary for phase correction at ω_o, (assuming matched opamps are used for the biquad).

From equation (7), it follows that the design value of the parameters a and K must be related by the following equation:

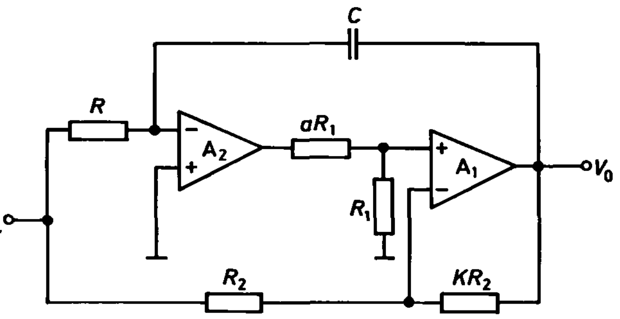

Fig. 1: Novel variable phase inverting integrator

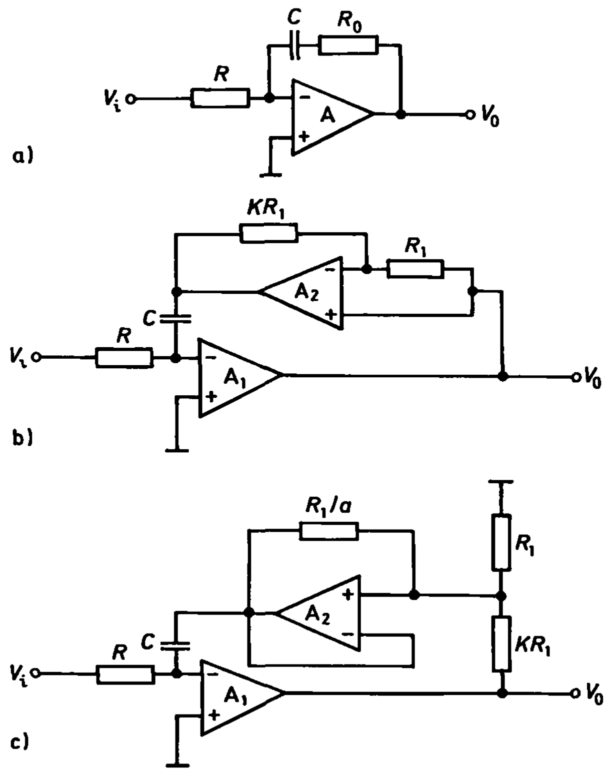

Fig. 2: Variable phase inverting integrators (see also table 1)

Table 1: A comparison table for the phase lead inverting integrators

Inverting Circuit Figure	Integrator Reference	Integrator Phase $\Phi \simeq$	Condition for phase correction in the TT or the KHN biquad	Stability Condition $\frac{\omega_2}{\omega_t} >$
2(a)	[7], [8]	$\omega\left[CR_o - \frac{1}{\omega_t}\right]$	$CR_o = \frac{4}{\omega_t}$	—
2(b)	[1]	$K\frac{\omega}{\omega_t}$	$K=3$	1.875
2(c)	[2]	$K\frac{\omega}{\omega_t}\left[\frac{a}{K+1}-1\right]$	$a=\frac{(K+1)(K+3)}{K}$	$\frac{2}{(K+1)} - \frac{(K+1)}{2[K(a+1)+1]}$
	Recommended Design $K=1$	$\frac{\omega}{\omega_t}\left[\frac{a}{2}-1\right]$	$a=8$	0.9
1	New	$\frac{\omega}{\omega_t}(a+1)\left(\frac{K-1}{K+1}\right)$	$a=\frac{2(K+2)}{(K-1)}$	$\frac{2(K+1)}{(a+1)} - \frac{1}{2(K+1)}$
	Design (1) $a=2K+1$	$\frac{\omega}{\omega_t}\cdot 2(K-1)$	$a=6,\ K=2.5$	0.857
	Design (2) $K=3$	$\frac{\omega}{\omega_t}\cdot\frac{(a+1)}{2}$	$a=5,\ K=3$	1.21

$$a = \frac{2(K+2)}{(K-1)}. \tag{11}$$

From equations (10) and (11) it follows that for phase correction in the TT and the KHN biquads the parameters a and K are given by

$$a = 6 \quad \text{and} \quad K = 2.5. \tag{12}$$

It is worth noting that other designs are possible. For example, taking $K=3$, thus using equation (11), it follows that $a=5$. In this case however, as seen from equation (9), it is necessary to use opamps having $\omega_2 > 1.21\ \omega_t$, (which is satisfied by many general purpose opamps).

5. Conclusions

A new variable phase inverting integrator is given. The integrator has the advantage that its phase depends on the gain-bandwidth of only one of the two opamps, employed in the circuit. Applications of the proposed phase lead integrator in realizing high frequency two integrator loop filters have been considered. **Table 1,** includes the properties of the phase lead inverting integrators shown in Fig. 1 and **Fig. 2,** as well as the condition for phase correction in the Tow-Thomas and the Kerwin-Huelsman-Newcomb biquad filters. It is worth noting that there is a degree of freedom available in the design*) of the new integrator as well as in the integrator of Fig. 2(c).

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*) The magnitude of the phase of the opamp at $\omega=\omega_t$, $\Phi_{opamp}(\omega_t)$ depends on the choice of the degree of freedom. For the new integrator, and with $\omega_2/\omega_t > 0.857$ (design 1), $\Phi_{opamp}(\omega_t) \leq 157°$, with design (2), however, $\Phi_{opamp}(\omega_t) \leq 137°$. On the other hand with the circuit of Fig. 2(c), and using the degree of freedom as recommended in Table 1, $\Phi_{opamp}(\omega_t) \leq 153°$.

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The UN Resolution invites Member States, UN specialized Agencies, governmental and non-governmental organizations and users of communications services to participate actively in the fulfilment of the objectives of the Year and to cooperate closely with the ITU Secretary-General.

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