Current-Mode Integrator using OA and OTAs and Its Applications

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Abstract: A circuit building block for realizing a continuous-time active-only current-mode integrator is presented. The proposed integrator is composed only of internally compensated type operational amplifier (OA) and operational transconductance amplifiers (OTAs). The integrator is suitable for integrated circuit implementation in either bipolar or CMOS technologies, since it does not require any external passive elements. Moreover, the integrator gain can be tuned through the transconductance gains of the OTAs. Some application examples in the realization of current-mode network functions using the proposed current-mode integrator as an active element are also given.

1. Introduction

In the last decade, the realizations of various active circuits utilizing the operational amplifier (OA) pole have received considerable attention for their potential advantages such as attractive for monolithic IC integration, ease to design, and suitable for high frequency operation [1-2]. Several OA-based active-R capacitor-less circuits for realizing analog transfer functions have been reported [3-4]. Presently from the above reasons, there is the strong motivation to design resistor-less and capacitor-less filter circuits utilizing the finite and complex gain natures of internally compensated OAs and OTAs. Due to the active only nature, the resistor-less and capacitor-less active filter would be attractive for its integratability, programmability and wide frequency range of operation. Many implementations in active-only filter design are available in the literature [5-7].

It is well-known fact that an integrator is an important circuit building block, which are widely used in analog signal processing applications, such as, filter design, waveform shaping, process controller design, and calibration circuit, etc. However, the implementation of a continuous-time current-mode integrator that employs only active elements has not yet been reported. Therefore, a circuit configuration for realizing active-only current-mode integrator is proposed in this paper. The proposed integrator consists of one OA and two OTAs. Since no passive element is required, the integrator can be implemented in integrated circuit form in both bipolar and CMOS technologies. The integrator gain can be electronically tuned by adjusting the transconductance gains of the OTAs. The various realizations of active-only analog signal processing circuits employing the proposed integrator will also be presented. Finally, the workabilities of the proposed integrator and its applications have been simulated based upon a LM741 type IC OA and a CA3080 type IC OTA.

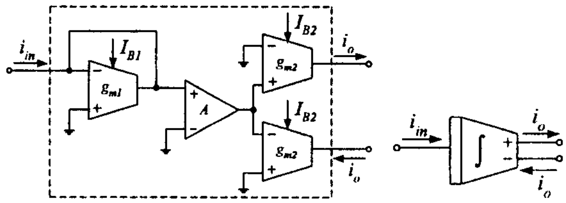


Fig.1 : The proposed active-only current-mode integrator (a) circuit implementation (b) circuit representation

2. Circuit Description

The proposed active-only dual-output current-mode integrator implementation and its representation are shown in Fig.1. It consists of only an OA and OTAs, where the dual-current-output OTA2 is implemented by using single-ended output OTAs in parallel connection [8]. If ω_a is the 3-dB bandwidth of the OA and by considering the OA for the frequencies $\omega >> \omega_a$, the open-loop gain $A_{OA}(s)$ of the OA can be approximately given by

$$A_{OA}(s) = \frac{A_o\omega_a}{s+\omega_a} \cong \frac{B}{s} \tag{1}$$

where B represents the gain-bandwidth product (GBP) of the OA, which is the product of the dc gain A_o and the 3-dB bandwidth ω_a. Let assume that g_{m1} and g_{m2} denote the transconductance gains of the OTA1 and OTA2, respectively, then the current transfer function of the current-mode integrator can be expressed as:

$$\frac{I_o(s)}{I_{in}(s)} = \frac{B}{s}\left[\frac{g_{m2}}{g_{m1}}\right] = \frac{B}{s}A_G \tag{2}$$

where A_G denotes the integrator gain, which is the ratio between g_{m2} and g_{m1}. Eqn.(2) indicates that the relationship of the currents I_o and I_{in} is in the form of the integrating action as required. It should be noted that, for ordinary bipolar OTAs, $g_{m1} = I_{B1}/2V_T$ and $g_{m2} = I_{B2}/2V_T$, where V_T is the thermal voltage and I_{B1} and I_{B2} are the bias currents of the OTA1 and OTA2, respectively. Thus, eqn.(2) becomes

$$\frac{I_o(s)}{I_{in}(s)} = \frac{B}{s}\left[\frac{I_{B2}}{I_{B1}}\right] = \frac{B}{s}A_G \qquad (3)$$

Now A_G is the current gain that is the bias current ratio between I_{B2} and I_{B1}. In this case, the temperature dependence of the transconductance gains g_{m1} and g_{m2} of the bipolar OTAs are also compensated.

Deviation from the ideal performance that predicted from the eqn.(2) is due to the parasitic effects of the non-ideality characteristics of the OA and OTAs. If the second dominant pole ω_b of the OA is taken for consideration, the OA open-loop gain $A_{OA}(s)$ can be rewritten by

$$A_{OA}(s) = \frac{B}{s}\frac{\omega_b}{(s+\omega_b)} = \frac{B}{s}\frac{1}{(1+\tau_b s)} \qquad (4)$$

where $\tau_b = 1/\omega_b$. For the OTAs, let $\omega_c = 1/\tau_c$ represents the effective transconductance internal-pole of the OTA and g_{m0} is the low frequency transconductance gain. The OTA open-loop gain $g_m(s)$ for general case can be described by

$$g_m(s) = \frac{g_{m0}}{\left(1+ {s}/{\omega_c}\right)} \cong g_{m0}\left(1- {s}/{\omega_c}\right) \qquad (5)$$

Therefore, the frequency response of the current-mode integrator in Fig.1 that including the second dominant pole of the OA and the transconductance internal-poles of the OTAs can now be given by

$$\frac{I_O(s)}{I_{in}(s)} = \left[\frac{B}{s}\right]\left[\frac{1}{1+\tau_b s}\right]\left[\frac{g_{m02}}{g_{m01}}\right]\left[\frac{\omega_{c2}-s}{\omega_{c2}}\right]\left[\frac{\omega_{c1}}{\omega_{c1}-s}\right] \qquad (6)$$

where ω_{c1} and ω_{c2} are the effective transconductance internal-poles of the OTA1 and OTA2, respectively. It can be seen that if the conditions ($\omega_{c1} \cong \omega_{c2}$) and ($\omega_{c1}$, $\omega_{c2} >> \omega$) are satisfied, then eqn. (6) becomes frequency independent. Let us define that $A_{G0} = g_{m02}/g_{m01}$ is the dc integrator gain, as a result, it can be reasonably reduced to

$$\frac{I_o(s)}{I_{in}(s)} = \left[\frac{A_{G0}B}{s}\right]\left[\frac{1}{1+\tau_b s}\right] \qquad (7)$$

One can see that the frequency characteristic of the proposed current-mode integrator has a dc current gain equaled to eqn.(2) and has a high frequency dominant pole located at ω_b. Hence, the OA pole ω_b should be the major high-frequency limitation of the proposed current-mode integrator.

3.Application examples

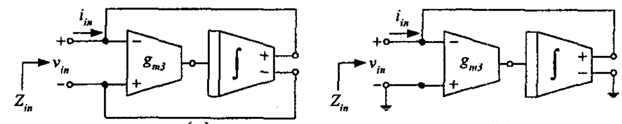
The following sections will concentrate on the usefulness of the proposed current-mode integrator. Some application examples to realize driving-point impedance function elements, current-mode biquadratic filter and current-mode nth-order filter will be demonstrated.

3.1. Inductance simulations

Fig. 2(a) shows the circuit diagram of a tunable floating inductance simulation. From this circuit, it can easily be shown that the value of the floating simulated inductance is

$$L_{eq} = \left[\frac{1}{g_{m3}A_G B}\right] \qquad (8)$$

It should be noted that the equivalent inductance L_{eq} can properly be tuned by electronic means through the current ratio A_G and/or the transconductance gain g_{m3}. In addition, the circuit in Fig.2(a) can easily be modified to simulate a grounded inductor by replacing the dual-output OTA2 with the single-output OTA2 as shown in Fig.2(b).



(a) (b)

Fig.2 : Active-only inductance simulations

3.2. Electronically tunable active-only current-mode biquadratic filter

Fig.3 shows the circuit diagram of the tunable current-mode filter based on the use of the proposed current-mode integrator, where i_{LP} , i_{BP}, and i_{HP} are the lowpass, bandpass and higpass current-output terminals, respectively. The circuit parameters ω_o and Q-factor of this filter can be written by

$$\omega_o = \sqrt{A_{G1}A_{G2}B_1B_2} \qquad (9)$$

and

$$Q = \frac{g_{mc}}{g_{md}}\sqrt{\frac{A_{G2}B_2}{A_{G1}B_1}} \qquad (10)$$

One can see that the filter also enjoys orthogonal tuning of ω_o and Q-factor via the transconductance gains of the OTAs and it's also temperature independent.

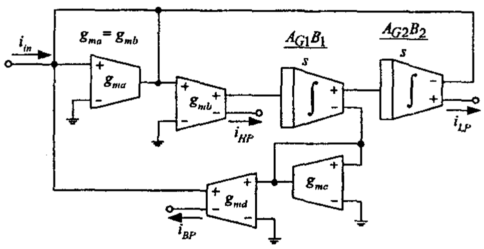
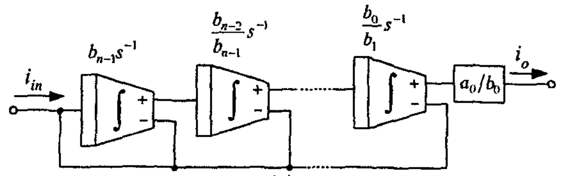
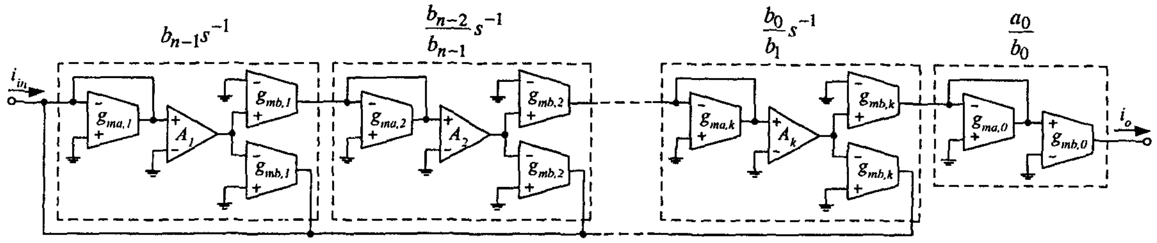


Fig.3 : Active-only current-mode biquadratic filter



(a)



(b)

Fig.4 : (a) nth-order current-mode filter representation
(b) nth-order current-mode filter realization using only active elements

3.3 nth-order current-mode filters

The standard current transfer function of an nth-order lowpass filter is often expressed as the following form :

$$\frac{I_o(s)}{I_{in}(s)} = \frac{a_0}{s^n + b_{n-1}s^{n-1} + \ldots + b_1 s + b_0} \qquad (11)$$

One can straightforwardly realize the nth-order current transfer function of equation (11) by cascading the proposed current-mode integrator of Fig.1. The system diagram thus obtained can be shown in Fig.4(a) and the coefficient of the standard function in terms of the circuit parameters can be expressed as follows :

$$b_{n-1} = A_{G1}B_1 \quad ; \quad A_{G1} = \frac{g_{mb,1}}{g_{ma,1}} \qquad (12)$$

$$\frac{b_{n-k}}{b_{n+1-k}} = A_{Gk}B_k \quad ; \quad A_{Gk} = \frac{g_{mb,k}}{g_{ma,k}} \qquad (13)$$

and

$$\frac{a_0}{b_0} = \frac{g_{mb,0}}{g_{ma,0}} \qquad (14)$$

where A_{Gk} and B_k represent the current gain and GBP of the k-th integrator (for k = 2, 3, ... , n). For this implementation, n active-only current integrators and two additional transconductance elements that employed for realizing the output proportional gain block ($g_{mb,0}/g_{ma,0}$) are required. Fig.4(b) shows the nth-order current-mode filter realization based on the use of only active components. In addition, if coefficients a_0 and b_0 are equaled, then two additional transconductance elements will also be eliminated.

4. Design examples and Simulation results

In order to verify the theoretical study of the proposed current-mode integrator, PSPICE simulation results are included. In this simulation, the OTA is modeled by employing CA3080 type OTA with a macro model [8] and LM741 type OA with the gain-bandwidth product B = 5.906 Mrad/s is used [5]. Fig.5 shows the simulated frequency responses of the proposed current-mode integrator. The results show that the circuit acts as an integrating function with a slope -20 dB per decade for the frequency range from 10 Hz to 1 MHz and has less than 10% phase error from the frequency range of 30 Hz to 500 kHz.

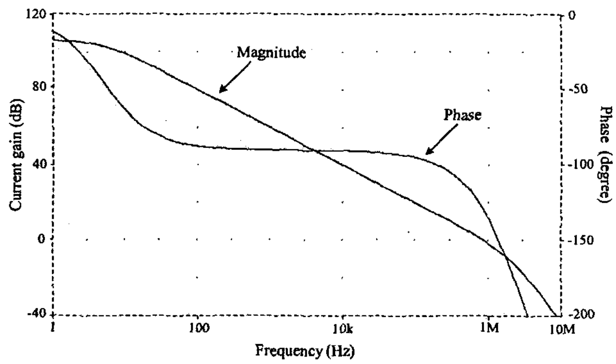
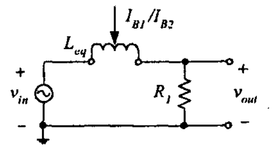


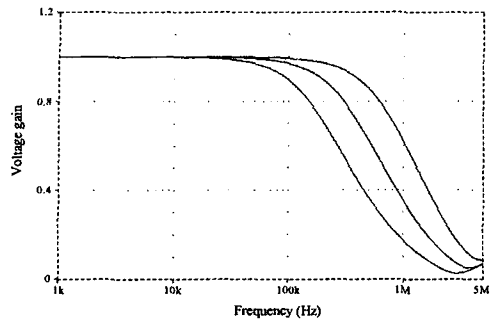
Fig.5 : Frequency responses of the proposed integrator

The performance of the floating inductance circuit of Fig.2(a) is demonstrated through the use of an electronically tunable active RL low-pass filter in Fig.6(a) with the external resistor R_1 = 1 kΩ. The bias current ratio $A_G = I_{B2}/I_{B1}$ (= g_{m2}/g_{m1}) is set to 0.5, 1 and 2, while g_{m1} and g_{m3} are respectively set to constant at 1 mS and 0.4 mS ; thus the cut-off frequencies f_C are approximately equal to 200 kHz, 400 kHz and 800 kHz, respectively. The frequency responses of the low-pass filter are shown in Fig.6(b).

Fig.7 shows simulated responses of the tunable current-mode multifunctional filter of Fig.3, when $A_{G1} = A_{G2}$ = 0.05 and g_{mc}/g_{md} = 0.1. This filter is designed for $\omega_o/2\pi$ = 50 kHz at the unity Q-factor. All the simulated results shown above imply that the proposed integrator exhibit reasonably good agreement with the predicted values.



(a)



(b)

Fig.6 : (a) first-order RL lowpass filter
(b) frequency responses of the simulated RL lowpass filter

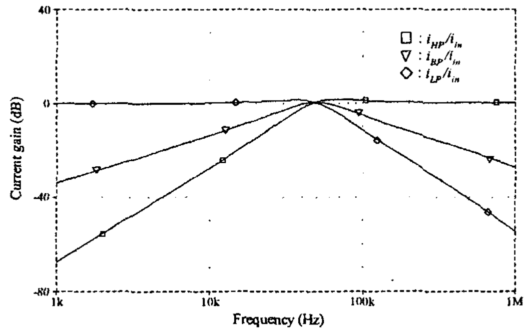


Fig.7 : Simulated frequency response of current-mode filter of Fig.3

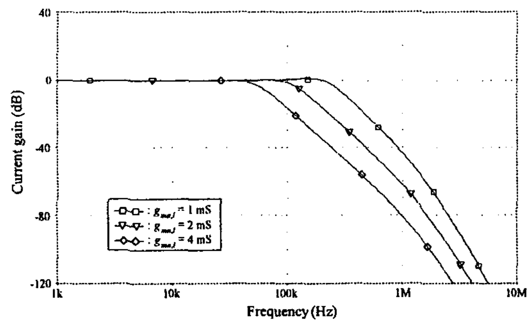


Fig.8 : Simulated frequency response of nth-order current-mode filter of Fig.4

To illustrate the current-mode filter using all active elements of Fig.4(b), a design of a third-order Butterworth filter with a cut-off frequency of 100 kHz is an example. The normalized transfer function for this filter can be written by

$$\frac{I_o(s)}{I_{in}(s)} = \frac{1}{s^3 + 2s^2 + 2s + 1} \quad (15)$$

The active-only filter realization based on the circuit of Fig.4 is constructed, in which the transconductance elements $g_{ma,0}$ and $g_{mb,0}$ will be eliminated due to the coefficient values a_0 and b_0 equal to unity. Thus, by calculating the circuit parameters we obtain : $g_{mb,1}$ = 0.424 mS, $g_{mb,2}$ = 0.212 mS and $g_{mb,3}$ = 0.106 mS, while $g_{ma,i}$ (i = 1, 2, 3) are set equal to 2 mS. In addition, the values of $g_{ma,i}$ can be used to change the cut-off frequency. The simulated frequency responses are shown in Fig.8 that exhibit reasonably close agreement with theoretical results. It is also shown that the cut-off frequency can be tuned electronically through adjusting the transconductance gains ($g_{ma,1} = g_{ma,2} = g_{ma,3}$).

5. Conclusions

This paper presented an alternative scheme for realizing continuous-time active-only current-mode integrator. The proposed integrator is realizable with only internally compensated type OA and OTAs, and does not require any external passive elements. Because of their active-only nature, the integrator can be easily employed to realize active network functions and are suitable for implementing in monolithic integrated form in both bipolar or CMOS technologies. Since the proposed circuit utilizes an OA pole, it is also suitable for high frequency operation. The simulated results have been used to verify the theoretical analysis.

6. Acknowledgment

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