Dual-Output All-Pass Filter Employing Fully-Differential Operational Amplifier and Current-Controlled Current Conveyor

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Abstract

First-order voltage-mode all-pass filter providing high input and low output impedances is described. The proposed filter employs a fully-differential operational amplifier, a current-controlled current conveyor CCCII, and one capacitor, with the possibility of electronic control of natural frequency. Two voltage outputs are simultaneously available for non-inverting and inverting transfer functions, with the possibility of utilizing them as a differential output.

1. Introduction

Since 1996, many active filters, employing the non-inverting and inverting first-order all-pass filters, have been published [1-28]. The reason consists in the importance of these building blocks in various applications, particularly for on-chip analog signal processing and generation.

The required features of voltage-mode topologies include high input and low output impedances, simplicity of circuit topology, which predetermines low-power operation, low sensitivities to parameter variations, and particularly the absence of matching conditions among the component parameters which would be necessary for the proper operation. Some applications require electronic control of natural frequency.

Fulfilling the above requirements simultaneously is a difficult task [28]. Many researchers searched for various circuit structures which employ assorted types of active elements, starting from conventional operational amplifiers (OpAmps) [1-5], continuing with commercially available Current-Feedback OpAmps [6] and Operational Transconductance Amplifiers [7-8], and utilizing also modern building blocks, namely various types of current conveyors [9-15], Differential Difference Amplifier (DDA) [16], Current Differencing Buffered Amplifier (CDBA) [17-21], Current Differencing Transconductance Amplifier (CDTAs) [22-27], Voltage Differencing – Differential Input Buffered Amplifier (VD-DIBA) [28] and others.

In addition to the above-mentioned requirements, also economical versions of all-pass filters are needed, which include both the non-inverting and inverting versions in a single device. In [29], a fully-differential OpAmp-based all-pass filter provides both outputs, enabling also differential voltage output. However, the natural frequency of this filter is determined by a fixed resistor, and its direct electronic control is thus problematic.

In Section II of this paper, a fully-differential OpAmp-based all-pass filter is proposed which also employs a Current-Controlled Current Conveyor CCCII+ and one capacitor. Then the natural frequency of the filter is determined by the conveyor intrinsic resistance, which can be simply controlled via a bias current. The capacitor is of the floating type. However, its one outlet is connected to low-impedance OpAmp output, which eliminates the corresponding parasitic impedance. The influence of the impedance associated with the second outlet is analyzed in Section III. Section IV contains experimental verification of the proposed circuit architecture via SPICE simulations, utilizing a transistor-level model of CCCII+ and behavioral model of commercially available OpAmp.

2. Proposed all-pass filter

The proposed all-pass filter employing a fully-differential OpAmp and a CCCII+ is shown in Fig. 1.

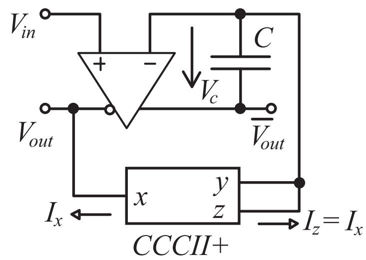

Fig. 1. Proposed all-pass filter with high-impedance input V_{in} and two low-impedance outputs V_{out} and $\overline{V}_{out} = -V_{out}$.

Due to the negative feedback in the circuit, the OpAmp differential input voltage is close to zero, and the input voltage V_{in} appears at the y terminal of CCCII+. The current I_x, flowing out of the x terminal, is given by the voltage at the intrinsic resistance R_x of this terminal, thus

$$I_x = \frac{V_{in} - V_{out}}{R_x}. \tag{1}$$

This current is conveyed to the z terminal and then it flows through the capacitor C, causing the voltage drop

$$V_c = V_{in} - \overline{V}_{out} = V_{in} + V_{out} = \frac{I_x}{sC}. \tag{2}$$

Combining (1) and (2) yields

$$V_{in} + V_{out} = \frac{V_{in} - V_{out}}{sCR_x}. \tag{3}$$

The filter transfer function is then

$$\frac{V_{out}}{V_{in}}=-\frac{\overline{V}_{out}}{V_{in}}=\frac{1-sCR_x}{1+sCR_x}. \tag{4}$$

Note that R_x can be electronically adjusted via the bias current I_B according to the equation [30]

$$R_x=\frac{V_T}{2I_B} \tag{5}$$

where V_T is the thermal voltage. Thus (4) represents typical transfer functions of the non-inverting and inverting first-order all-pass filter with the natural frequency

$$\omega_0=\frac{1}{CR_x}=\frac{2}{V_TC}I_B \tag{6}$$

which is directly proportional to the bias current. The temperature dependence can be advantageously utilized in various temperature sensors such as temperature/phase shift converters or temperature/frequency converters (when the all-pass cell is a part of sinusoidal oscillator).

3. Non-ideal case

The theoretical small-signal behavior of the filter, described by Eq. (4), is affected by several non-idealities. Transfer function (4) can be influenced by the following real properties of the active and passive elements: OpAmp input and output impedances, parasitic impedances of conveyor terminals, ESR (Equivalent Serial Resistance) and parasitic pin impedances of the capacitor, "alpha" and "beta" factors of the current conveyor, defined by the equations

$$V_x=\beta V_y-R_xI_x\,,\; I_z=\alpha I_x\,, \tag{7}$$

and the frequency dependence of key parameters, particularly the OpAmp gain and current conveyor "alpha" and beta" factors.

The above parameters were included in the model of the filter in Fig. 1, and the approximate symbolic analysis of transfer function (4) was performed via SNAP symbolic program [31]. Numerical values of the component parameters considered are from Section IV. Dominant sources of the imperfections in the frequency responses within a given frequency range, revealed via such an analysis, are indicated in Fig. 2: "alpha" and "beta" factors of the current conveyor, and R_p and C_p parasitic elements, which model the parasitic impedances of the terminals of current conveyor, OpAmp, and capacitor C. Other parasitic impedances do not manifest themselves due to low-impedance OpAmp outputs.

Taking into account the above non-idealities, filter transfer function (4) is in a more general form:

$$\frac{V_{out}}{V_{in}}=-\frac{\overline{V}_{out}}{V_{in}}=\frac{\beta-\dfrac{R_x'}{\alpha R_p}-s\dfrac{C+C_p}{\alpha}R_x'}{1+s\dfrac{CR_x'}{\alpha}}, \tag{8}$$

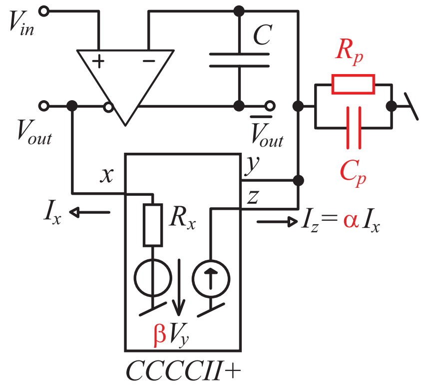

Fig. 2. Simplified model for error analysis of filter from Fig. 1.

where R_x' is the resistance R_x modified by the influence of OpAmp output resistance R_o:

$$R_x'=R_x+R_o. \tag{9}$$

Note that the low-frequency gain (LFG) and the high-frequency gain (HFG) are as follows:

$$LFG=\beta-\frac{R_x'}{\alpha R_p},\; HFG=1+\frac{C_p}{C}. \tag{10}$$

Their ratio, which is equal to one in the ideal case, can serve as a rough measure of parasitic passband ripple. Note that the high-frequency gain has a tendency to be higher than one, which can be eliminated via selecting large capacitance $C >> C_p$. On the other hand, the low-frequency gain has a tendency to be smaller than one, particularly for $\beta < 1$. This tendency can be reduced via selecting small intrinsic resistance $R_x << R_p$.

Note that the above conclusions about the high-frequency gain do not take the high-frequency behavior of active elements into account.

After a simple arrangement of Eq. (8) we obtain

$$\frac{V_{out}}{V_{in}}=-\frac{\overline{V}_{out}}{V_{in}}=\frac{1-s\dfrac{C}{\alpha}R_x'}{1+s\dfrac{C}{\alpha}R_x'}+\frac{\beta-1-\dfrac{R_x'}{\alpha R_p}-s\dfrac{C_p}{\alpha}R_x'}{1+s\dfrac{CR_x'}{\alpha}}. \tag{11}$$

The first term on the right side of (11) represents the transfer function of ideal all-pass filter with the natural frequency

$$\omega_0'=\frac{\alpha}{CR_x'}=\alpha\frac{R_x}{R_x'}\omega_0. \tag{12}$$

The second term can be interpreted as an additional error term. At the natural frequency (12), its absolute value is

$$err=\sqrt{\frac{\left(\beta-1-\dfrac{R_x'}{\alpha R_p}\right)^2+\left(\dfrac{C_p}{C}\right)^2}{2}}. \tag{13}$$

This value (err) can serve for the estimation of a global error which is caused by the real influences. Since the "beta" factor is normally declined from its ideal value 1 by several per cent points, it can be a dominant error source in comparison to R_p and C_p, which normally differ by several decades from R_x and C. For example, when β=α=0.95, $R'_x/R_p=C_p/C=10^{-4}$, then

$$err \doteq 0.035,\ LFG \doteq 0.950,\ HFG = 1.0001. \qquad (14)$$

The first value means that the error term in (11) is by 29 dB less than the first term, which represents the ideal all-pass filter with natural frequency (12). This error is mainly caused by the imperfection of the "beta" term, which is also responsible for the decreased low-frequency gain. The high-frequency gain is almost ideal, since it is not influenced by the "beta" factor.

At the natural frequency ω_0 (6) of ideal non-inverting all-pass filter with transfer function (4), the phase shift is exactly -90°. The impact of the real properties can be also measured by comparing this frequency with the frequency ω''_0, at which the same phase shift appears in the non-ideal case. An analysis of Eq. (8) leads to the following result:

$$\frac{\omega''_0}{\omega_0} = \alpha \sqrt{\frac{\beta - \dfrac{R'_x}{\alpha R_p}}{1+\dfrac{C_p}{C}}}. \qquad (15)$$

It should be noted that there are several reasons for decreasing the frequency of the -90° phase shift due to real influences: Decreasing the alpha and beta factors below 1, large parasitic capacitance C_p, and the R_x/R_p ratio when it cannot be neglected.

4. Spice simulations

In order to verify the feasibility of the proposed all-pass filter, a PSPICE simulation of the structure in Fig. 1 was performed.

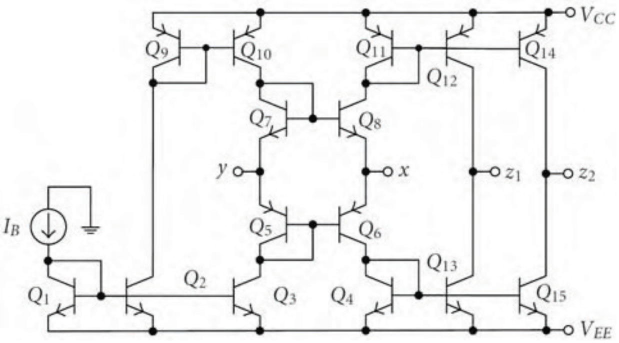

Fig. 3. Transistor-level model of CCCII, adopted from [30].

The transistor-level modeling of the CCCII was performed according to [30], see Fig. 3. Since only the single-output conveyor is used in Fig. 1, transistors Q14 and Q15 are omitted in the model. As in [30], the PNP and NPN transistors were modeled with the parameters of the PR200N and NR200N bipolar transistors of ALA400 transistor array from AT&T [32], see Table I. Symmetrical supply voltages of ±1.5 V were used. The bias current $I_B = 50\ \mu A$ corresponds to the intrinsic resistance $R_x = 258.6\ \Omega$ for a temperature of 27 °C (see Eq. 5). For the capacitance $C = 6.16$ nF, the theoretical natural frequency from (6) is 100 kHz.

With regard to the low-voltage model of CCCII, the low-voltage fully differential rail-rail OpAmp LTC6403-1 was also used in the filter, together with its SPICE model from [33]. Its supply voltages were chosen the same as for the CCCII.

Table 1. Transistor models, adopted from [30].

.model PX PNP
+RB=327 IRB=0 RBM=24.55 RC=50 RE=3
+IS=73.5E-18 EG=1.206 XTI=1.7 XTB=1.866 BF=110
+IKF=2.359E-3 NF=1 VAF=51.8 ISE=25.1E-16 NE=1.650
+BR=0.4745 IKR=6.478E-3 NR=1 VAR=9.96 ISC=0 NC=2
+TF=0.610E-9 TR=0.610E-8 CJE=0.180E-12 VJE=0.5
+MJE=0.28 CJC=0.164E-12 VJC=0.8 MJC=0.4 XCJC=0.037
+CJS=1.03E-12 VJS=0.55 MJS=0.35 FC=0.5
.model NX NPN
+RB=524.6 IRB=0 RBM=25 RC=50 RE=1
+IS=121E-18 EG=1.206 XTI=2 XTB=1.538 BF=137.5
+IKF=6.974E-3 NF=1 VAF=159.4 ISE=36E-16 NE=1.713
+BR=0.7258 IKR=2.198E-3 NR=1 VAR=10.73 ISC=0 NC=2
+TF=0.425E-9 TR=0.425E-8 CJE=0.214E-12 VJE=0.5
+MJE=0.28 CJC=0.983E-13 VJC=0.5 MJC=0.3 XCJC=0.034
+CJS=0.913E-12 VJS=0.64 MJS=0.4 FC=0.5

In the first step, a DC analysis of the designed filter from Fig. 1 was performed with the input voltage swept within the range of -1V to +1V, see Fig. 4. For $V_{in} = 0$, the differential DC gain is approximately ±1.001 for both outputs. Linear operation is guaranteed approximately for abs(V_{in}) < 0.63V. For larger input voltages, the nonlinearity of CCCII becomes dominant.

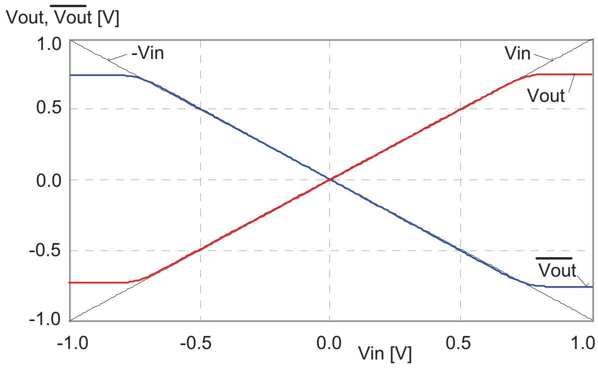

Fig. 4. DC analysis of proposed all-pass filter from Fig. 1.

The frequency responses of the filter from Fig. 1 are shown in Fig. 5, demonstrating the natural frequency control via the bias current. For $I_B = 50\ \mu A$, the frequency measured for -90° phase shift of the non-inverting cell is 92 kHz, which is ca 8 per cent below its theoretical value. This decrease is due to the real properties modeled by Eq. (15).

The transient analysis confirmed the filter stability. For I_B=500μA and sinusoidal excitation with a frequency of 1 MHz, the THD factor was 0.88% for an amplitude of 10 mV and 5.48% for 100 mV.

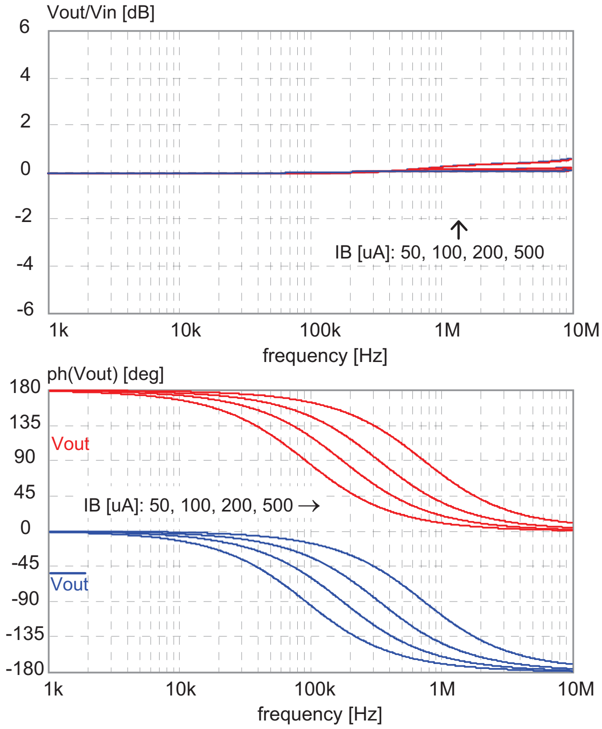

Fig. 5. Simulated amplitude and phase frequency responses for various values of bias current I_B.

5. Conclusions

The proposed first-order all-pass filter provides both non-inverting and inverting outputs simultaneously. It can be also used for generating a differential voltage output with double voltage swing in comparison with the single-ended solution. Due to the utilization of fully differential operational amplifier, the filter topology exhibits high-input and low-output impedances, which enables an easy cascade synthesis. The natural frequency can be tuned electronically via modifying the intrinsic resistance of the CCCII x terminal. As a certain drawback can bee seen in that the floating capacitor with the capacitance C is employed. However, one of its outlets is connected directly to the OpAmp low-impedance output. The influence of the parasitic impedance of the second outlet is analyzed in Section III, with the conclusion that it causes a modification of the natural frequency. This modification is negligible when the parasitic capacitance is much smaller than C. If necessary, this phenomenon can be easily compensated via a modification of C.

6. Acknowledgment

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