GROUNDED IMPEDANCE CURRENT SENSING

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ABSTRACT:

In several topologies of current-mode filters, copies of currents flowing through the grounded working impedances are provided. These currents are then flowing through independent loads, also grounded in most cases. The current approaches, utilizing the so-called multi-output OTAs, lead to the errors which are caused by the tracking errors of multiple current mirrors at the OTA output. In the paper, a method is described which is based on the generation of precise copy of the voltage drop on the grounded impedance. The principle of the method is critically analyzed and tested by computer simulation.

Keywords: OTA, Operational Amplifier, offset, symbolic analysis

1 INTRODUCTION

A drawback of a number of applications whose output signals are currents, consists in the fact that these currents flow through the working, mostly grounded impedances which belong to a given application. As an example, let us name the universal current-mode filters [1, 2] in which the output signals of high-pass sections are currents through grounded capacitors. Actually, these currents cannot be directly used for supplying the independent loads because any attempt to sense them, e.g. by series sensing elements, will affect the operation of such circuits. In several publications, this problem is solved by an auxiliary circuitry which reconstructs on the basis of the Kirchhoff law I: the current is defined as a sum or difference of the other currents which are associated with the node. Than copies of these current are provided e.g. by multiple-output OTAs (Operational Transconductance Amplifiers), and finally they are simply added or subtracted [3, 4]. However, the fundamental drawback of such method consists in a large sensitivity of the resulting current to the errors of producing these copies which are rather high in the current CMOS technologies [5].

An alternative method of solving this problem is based on the sensing of voltage not current of the grounded impedance, with the subsequent current reconstruction on the basis of the copies of voltage and impedance. In the paper, this method is critically analyzed. In the conclusion, its features as well as drawbacks are summarized.

2 PROPOSED CIRCUIT SOLUTION

The circuit schematics is shown in Fig. 1 (a). The current I and the grounded load Z are parts of a concrete applications which requires sensing of a copy of current I. The operational amplifier (OpAmp) senses by its high-impedance input the voltage drop V_z at the impedance and – in the ideal case – copies it to the impedance Z‘, which is designed as an ideal copy of Z. The current through Z‘ is thus equal the excited current I. This current flows thorough R_s, causing a voltage drop which is transferred by Operational Transconductance Amplifier (OTA) back to the current I_L, flowing to an independent load R_L. When the OTA transconductance is equal the reciprocal value of R_s, then $I_L = I$.

One can summarized that for ideal OpAmp and OTA and when considering no parasitic impedances, the output current is as follows:

$$I_L = g_m R_s \frac{Z}{Z'} I \, , \tag{1}$$

or

$$I_L = I \, , \text{ when } g_m R_s = \frac{Z'}{Z} \, . \tag{2}$$

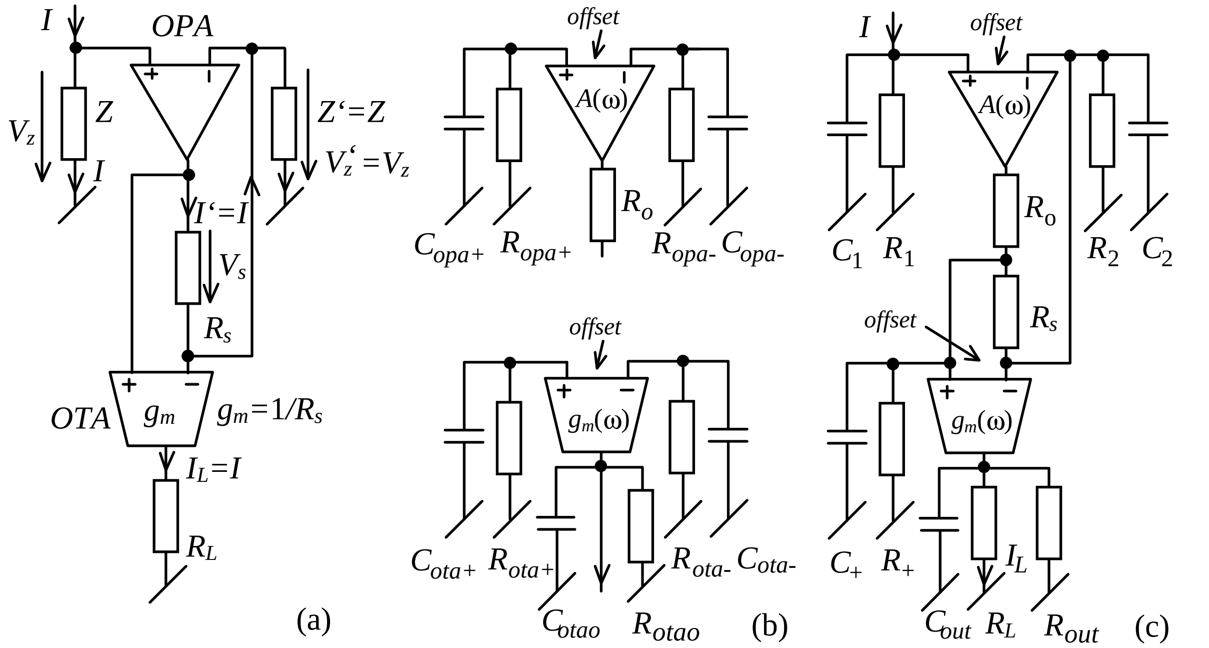

Fig. 1: (a) Proposed circuit for sensing a copy of current I, (b) OpAmp and OTA models, (c) model of the entire circuit, considering real influences.

The following strategies of achieving the unity current gain result from Eq. (2): either choosing the identical impedances Z and Z', and $g_m=1/R_s$, or the impedances can be different, but of the same character such as their ratio would be a real number, and then the product $g_m R_s$ must be equal this ratio.

In the following Section, the analysis of real influences is performed which helps to concretize the above simple conclusion.

3 MODELING REAL INFLUENCES IN THE CIRCUIT

The symbolic analysis of the proposed circuit will be given in Sections 4 and 5, considering dominant real influences. This analysis will be focused on a linear model and nonlinear distortion of the signal will not be studied. The DC characteristics of both amplifiers will be modeled as follows:

$$V_{out,opa} = A(V_{in,opa} + \Delta V_{in,opa}) = AV_{in,opa} + \Delta V_{out,opa} \, , \tag{3}$$

$$I_{out,OTA} = g_m (V_{in,ota} + \Delta V_{in,ota}) = g_m V_{in,ota} + \Delta I_{out,ota} \, , \tag{4}$$

where A and g_m are OpAmp DC open loop gain and OTA DC transconductance, both of them being independent of the signal, and the quantities, specifying by the symbol Δ, are the offsets in question.

Fig. 1 (b) illustrates real influences, considered when modeling small-signal behavior of the active elements: capacitances and resistances of their common inputs, OpAmp output resistance, and OTA output resistance and capacitance. Frequency dependence of the OpAmp voltage gain and OTA transconductance is modeled by single-pole models according to the formulae

$$A(s) = \frac{A}{1 + s/\omega_{opa}}, \; g_m(s) = \frac{g_m}{1 + s/\omega_{ota}}. \tag{5}$$

The circuit operation is also affected by the internal impedance of the source of current I. With respect to the types of applications under investigation and the working frequency region, this impedances will be modeled by a parallel combination of resistor and capacitor with parameters R_i and C_i.

The final model of entire circuit is given in Fig. 1 (c). This circuit has been completed on the assumption that the original impedances Z and Z' are represented by capacitors with the capacitances C and C'. Then

$$C_1 = C + C_{opa+} + C_i, \; R_1 = R_{opa+} \,\&\, R_i, \tag{6}$$

$$C_2 = C' + C_{opa-} + C_{ota-}, \; R_2 = R_{opa-} \,\&\, R_{ota-}, \tag{7}$$

$$C_+ = C_{ota+}, \; R_+ = R_{ota+}, \tag{8}$$

Consider the case which is common for the current mode, i.e. that the circuit operates to the low impedance, thus $R_L \rightarrow 0$. Then the OTA output impedance takes no effect on the current I_L.and we do not have to analyze its influence.

Typical values of parameters of commercial amplifiers have been considered for the purposes of simplified symbolic analysis. Differential-input OTA can be implemented by a pair of "supertransistors" OPA615 [6] with a degenerating resistor for setting the transconductance and increasing the linearity. Fast AD8021 was selected as voltage OpAmp with GBW of 200MHz, adjusted by a compensation 10pF capacitor [7]. The working capacitances $C = C'$ were designed to 100pF. For a simplicity, the internal impedance of the current source I has been chosen such as the input impedance of connected OpAmp. A summary of these parameters is below:

$\Delta V_{in,opa}$ = 0.4mV, $C_{opa+} = C_{opa-}$= 1pF, $R_{opa+} = R_{opa-}$ = 10MΩ, R_o = 100Ω,
A = 86dB ≈ 19953, $GBW = A.f_{opa}$ ≈ 200MHz,
$\Delta V_{in,ota}$ = 4mV, $C_{ota+} = C_{ota-}$= 1.5pF, $R_{ota+} = R_{ota-}$ = 7MΩ, g_m = 1mA/V = 1/R_s, f_{ota} = 440MHz,
$R_i \rightarrow$ 10MΩ, C_i = 1pF.

The following values hold from Eqs. (6) –(8):

C_1 = 102pF, R_1 = 5MΩ, C_2 = 102.5pF, R_2 = 4.1MΩ, C_+ = 1.5pF, R_+ = 7MΩ.

4 DD CHARACTERISTICS OF THE CIRCUIT

The DC symbolic analysis of the circuit in Fig. 1 (c) leads to the following result:

$$I_L = \frac{g_m R_s A(R_1 I + \Delta V_{in,opa})}{R_o \dfrac{R_s + R_2}{R_+} + R_o + R_s + R_2 + AR_2} + g_m \Delta V_{in,ota}. \tag{9}$$

With respect to a large gain A, the terms in the denominator can be vanished compared to the last term. The simple arrangement yields

$$I_L \approx g_m R_s \frac{R_1}{R_2} I + g_m \frac{R_s}{R_2} \Delta V_{out,opa} + \Delta I_{out,ota} \approx 1.220I + 1.947\mu A + 4\mu A = 1.513I + 5.947\mu A\,. \quad (10)$$

The above result is instructive for several reasons. It indicates a doubtful utilization of this circuit in DC regime or in low-frequency region when the working capacitances C and C' do not take effect. Then the current gain depends on the ratio of two parasitic resistances R_1 and R_2. The resulting error, which can be considerable, can be post facto corrected by modifying the OTA transconductance. Other problem is represented by an output offset which also depends on the transconductance value and which can reach large values. In DC regime, the maximum current I is limited by the maximum voltage drop on R_1, which cannot exceed – depending on the supply voltage – several volts. For five volts, the maximum operating current is 1.39μA which is smaller than the estimated offset. That is why one should carefully design such circuit with reference to the offset and its compensation as well as to the compensation of the influence of parasitic resistances on the current gain.

5 SMALL-SIGNAL CIRCUIT CHARACTERISTICS

The full symbolic analysis of circuit in Fig. 1 (c) provides rather blind and complicated equations. A certain simplification can be done, utilizing the fact that the resulting frequency characteristics are virtually insensitive to the variations of OpAmp DC gain, and their sensitivity to the variations of the OpAmp output resistance R_o are low in certain cases. The simplifying conditions

$$A \rightarrow \infty,\ R_o = 0 \quad (11)$$

lead to the final transfer function

$$\frac{I_L}{I} = \frac{R_1}{R_2}\frac{1+sR_2C_2}{1+sR_1C_1}\frac{g_m}{(1+\dfrac{s}{\omega_{ota}})}\frac{R_s}{(1+s\dfrac{R_2+R_s}{\omega_t.R_2}+s^2\dfrac{R_sC_2}{\omega_t})},\ \omega_t = A\omega_{opa}\,. \quad (12)$$

In the first approximation, the dynamic properties of the circuit are are affected by one real zero point, two real poles, and a pair of generally complex poles which come from the quadratic term in the denominator. A partial zero-pole cancellation can be accomplished when fulfilling the condition

$$R_1C_1 = R_2C_2\,. \quad (13)$$

It can be provided theoretically by adjusting C_2 by means of C'.

The next real pole is induced by the frequency dependence of the transconductance (5).

The quadratic term is a model of 2nd order system with the natural frequency ω_0 and the quality factor Q, where

$$\omega_0 = \sqrt{\frac{\omega_t}{R_sC_2}},\ Q = \frac{R_2}{R_2+R_s}\sqrt{\omega_t R_s C_2} \approx \sqrt{\omega_t R_s C_2}\,. \quad (14)$$

After evaluation we get $f_0 \approx$ 11.3MHz, $Q \approx$ 11.3. The step response will suffer by undesirable ringing at frequencies of about 11MHz. Eq. (14) also gives a direction of eliminating this problem: decreasing R_s results in decreasing the quality factor and simultaneously shifting up the natural frequency. Utilizing a slower OpAmp with lower GBW also decreases the quality factor as well as the natural frequency. Both interventions can be combined. In order to provide unity current gain at low frequencies, the transconductance must be increased. In this step, also the affect of the spread of parasitic resistances R_1 and R_2 on the DC gain can be compensated. On the other hand, increasing the transconductance results in increasing the current offset, and decreasing R_s lowers the voltage drop for OTA sensing towards the noise voltage. That is why the trade-off design should be accomplished.

It should be also noted that equations (12)-(14) were derived on the assumption of zero output OpAmp resistance. The experiments show that the output resistance increases the quality factor above values which are generated by Eq. (14).

The results of computer simulation before (a) and after (b) circuit optimization are in Fig. 2. The optimization consists in the correction of the capacitance C_2 to the value of 122pF in order to provide a correct zero/pole cancellation. A slower OpAmp with GBW=10MHz was used. The sensing resistance R_s was decreased to 100 Ohms and the transconductance was set to 0.82mA/V. Then the low-frequency gain was set to 1.

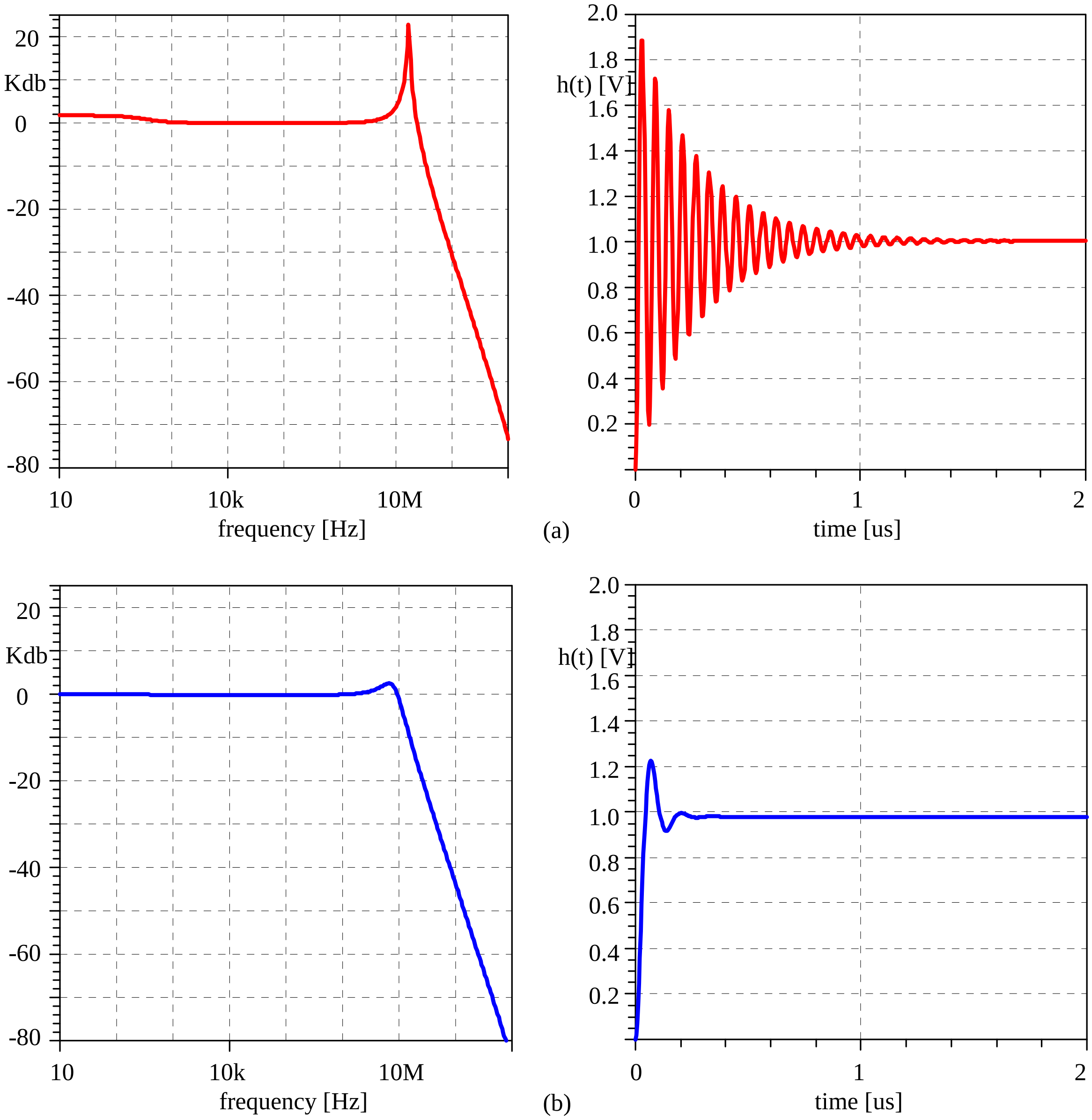

Fig. 2: Frequency responses and step responses of the circuit (a) before, (b) after the optimization.

6 CONCLUSIONS

A circuit for copying the current through a grounded impedance which is a part of an autonomous application is proposed. Compared it to the existing solutions, neither disconnecting of the current circuitry for inclusion of a sensing device is not required nor sensitive sum of current copies is performed. It consists in a single-terminal connection to the ungrounded outlet of the impedance, creating the copy of the impedance voltage, connecting a copy of the impedance, and sensing and replication of the current through this copy. Three related implementation problems are shown in the paper: a necessity of the compensation of the offsets of active devices, a necessity of compensation of current gain variations due to its dependence on parasitic resistances, and a necessity of the optimization of dynamic properties of the circuit.

7 ACKNOWLEDGMENT

This work was supported in part by the Grant Agency of the Czech Republic under grant No. 102/08/0784, and by the research programmes MSM 0021630503, MSM 0021630513, and MO FVT0000403.

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