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A nonlinear macromodel for current-feedback operational amplifiers

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ABSTRACT

This paper proposes a macromodel to emulate the nonlinear behavior of current-feedback operational amplifiers (CFOAs) at low-frequency. The main difference between this macromodel and those reported previously in the literature is that herein, real physical active device performance parameters along with parasitic elements associated to the input–output terminals of the amplifier are considered. To validate the deduced behavioral model, a saturated nonlinear function series (SNFS) based on CFOAs is built and numerical simulations are generated. In this point, the modeling problem is cast in terms of an augmented set of equations but that, unlike a piece-wise linear (PWL) approach, the dynamic behavior of each CFOA is considered. Afterwards, the SNFS is experimentally tested by using commercially available active devices, confirming good agreement among theoretical simulations and experimental tests at two operating frequencies and showing a better accuracy compared with a PWL approach and a linear model for CFOAs. Because the derived nonlinear macromodel for CFOAs is used for generating the behavioral model of the SNFS, one concludes that the latter is also both accurate and efficient with respect to traditional techniques, such as PWL approaches.

1. Introduction

The current-feedback operational amplifier (CFOA) has emerged as one building block in linear and nonlinear circuit applications [1–36]. Basically, the CFOA is a hybrid active device that handles voltage and current signals at its input–output terminals and this is usually modeled with two voltage-controlled voltage sources, one between *y*- and *x*-terminals and another between *z*- and *w*-terminals, whereas a current-controlled current source is used between *x*- and *z*-terminal [1–7]. All controlled sources are assumed to be ideal with unity-gain. Furthermore, on the *z*-terminal, an RC circuit is included to model the dominant pole of the open-loop amplifier [12,13,15,29]. Although this macromodel has widely been used for characterizing analog circuits containing CFOAs, some main performance parameters are not yet considered, such as the slew-rate (*SR*) and the dynamic range (*DR*), along with parasitic elements associated to the input–output terminals. This is a serious drawback, since in practice, the predicted performance of the electronic circuit differs substantially from reality.

Moreover, CFOA-based analog circuits performance characteristics can be evaluated by using a more complex model but with better accuracy, e.g., at the transistor level of abstraction. However, it is well-known that complex models are slower when they are used into network analysis programs, since they require a large memory and a high CPU-time [33]. Otherwise, simple models may, however, compromise the accuracy. In response to this trade-off [33,37], several CFOA macromodels have been reported in the literature and they have evolved according to the circuit needs. A simple macromodel that takes into account current and voltage tracking errors was proposed in [3]. Another macromodel was also developed in [1,7], but unlike of the previous model, basic parasitic elements were included. Fabre and Alami [9] developed a simple macromodel which is fairly accurate into a limited range of frequencies. In [10], two SPICE macromodels were proposed and they use a lower number of two-terminal elements than [1–9,29]. These macromodels are adequate for frequency analysis of linear circuits, although not sufficiently complete for time-domain analysis of nonlinear circuits. However, despite their simplicity, the main drawback lies in the *SR* and *DR* effects that are not included in all macromodels mentioned above. It is worth mentioning that although today computers with several processors working at high speed are available, macromodels are useful since they can be used not only to gain insight on the nonlinear behavior of complex analog circuits, but can also be used into electronic design automation methodologies in order to synthesize and analyze the trade-offs of electronic systems at different levels of abstraction [37].

This paper presents an approach to model the nonlinear behavior of CFOAs, including not only real physical active device parameters, but also parasitic elements associated to the input–output terminals [15,29,34]. By using the proposed macromodel, the dynamical behavior of linear and nonlinear systems at the

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circuit level of abstraction can efficiently be modeled [34,35]. In this context, the behavioral model generation of a saturated nonlinear function series (SNFS) based on CFOAs is chosen as test vehicle [22]. As a consequence, the derived behavioral model of the SNFS could be used to numerically generate multi-scroll chaotic attractors more real than PWL approaches [14,16,17,22,24,31,32,35]. The paper is organized as follows: Section 2 deals with the derivation of the behavioral macromodel for CFOAs. In Section 3, the behavioral model generation for SNFS, by using the CFOA macromodel derived previously, is introduced. In Section 4, experimental results of the SNFS built with commercially available active devices are obtained and compared with a PWL approach, a linear model for CFOAs and the proposed nonlinear macromodel for CFOAs. Finally, conclusions are drawn in Section 5.

2. Behavioral model generation for CFOAs

A macromodel is a model that mimics the real behavior of a circuit or active device without modeling each of its components. The CFOA symbolic representation is depicted in Fig. 1(a) and the proposed nonlinear macromodel is shown in Fig. 1(b), where the most influential performance parameters are included. According to Fig. 1(b), R_y, R_x, R_z and R_w are the parasitic resistances on the y-, x-, z- and w-terminals, whereas C_y and C_z are the parasitic capacitances on the y- and z-terminals, respectively [20,28,30]. Besides, A_{v1}, A_{v2} and A_i model the finite open-loop voltage and current gains of the controlled sources. The behavior of the CFOA, when operates linearly, is governed by the following equation:

$$\begin{aligned}\frac{dv_w(t)}{dt} &= A_{v1}GBv_y(t) - GBv_x(t) - \frac{GB}{A_{DC}}v_w(t) + R_w\frac{GB}{A_{DC}}i_w(t)\\ &\quad + R_x\frac{GB}{A_i}i_z(t) + R_w\frac{di_w(t)}{dt},\\ &\quad -A_{v2}\frac{SR}{GB} \le A_{v1}v_y(t) - v_x(t) \le A_{v2}\frac{SR}{GB}\end{aligned} \tag{1}$$

where $A_{DC} = \frac{A_iA_{v2}R_z}{R_x}$ is the DC gain, $I_{SR} = C_zSR$ is the slew-rate current source, $\omega_{dp} = \frac{1}{R_zC_z}$ is the dominant-pole corner frequency of the open-loop amplifier and $GB = \frac{A_iA_{v2}}{R_xC_z}$ is the gain-bandwidth product of the CFOA [30]. Otherwise, when the CFOA enters in the positive or negative saturation current zone, $v_w(t)$ evolves according to

$$\begin{aligned}\frac{dv_w(t)}{dt} &= -A_{v2}SR - \frac{GB}{A_{DC}}v_w(t) + R_w\frac{GB}{A_{DC}}i_w(t) + R_x\frac{GB}{A_i}i_z(t) + R_w\frac{di_w(t)}{dt},\\ &\quad -A_{v2}\frac{SR}{GB} > A_{v1}v_y(t) - v_x(t)\\ \frac{dv_w(t)}{dt} &= +A_{v2}SR - \frac{GB}{A_{DC}}v_w(t) + R_w\frac{GB}{A_{DC}}i_w(t) + R_x\frac{GB}{A_i}i_z(t) + R_w\frac{di_w(t)}{dt},\\ &\quad +A_{v2}\frac{SR}{GB} < A_{v1}v_y(t) - v_x(t)\end{aligned} \tag{2}$$

Note that $v_x(t)$ and $v_w(t)$ are limited by the saturation voltages and the voltage drops of the parasitic resistances on the x- and w-terminals, as shown in Fig. 1(b) and given by

$$\begin{aligned}&i_x(t)R_x + V_{ns} \le v_x(t) \le i_x(t)R_x + V_{ps}\\ &i_w(t)R_w + V_{ns} \le v_w(t) \le i_w(t)R_w + V_{ps}\end{aligned} \tag{3}$$

where V_{ns} is the negative saturation voltage, V_{ps} is the positive saturation voltage and its difference is the DR of the CFOA. It is important to stress that $i_x(t)$ and $i_w(t)$ depend of the applied signal amplitude on the x- and w-terminals, respectively. Assuming that $R_w \approx 0$ and $i_z(t) \approx 0$, a linear model for CFOA can be deduced from (1) as

$$\frac{dv_w(t)}{dt} \approx A_{v1}GBv_y(t) - GBv_x(t) - \frac{GB}{A_{DC}}v_w(t) \tag{4}$$

Eqs. (1)–(4) will be used to numerically predict the behavior of a SNFS based on CFOAs and simulations will be compared not only with a PWL model, but also with experimental results at two operating frequencies, showing that the real behavior of the SNFS using the nonlinear behavioral model is more accurate than those models mentioned previously [22,24,31–33].

3. Behavioral model generation for SNFS

In order to validate the nonlinear macromodel for CFOAs derived previously, a SNFS is used as test vehicle and built as depicted in Fig. 2(a) [22]. Because Fig. 2(a) is composed by stacked basic n-blocks, we will analyze only one of them using both: the nonlinear macromodel shown in Fig. 1(b) along with its behavioral equations given by (1)–(4) and the linear model given by (4). The equivalent circuit of the n-block is depicted in Fig. 2(b), where it has been assumed that the voltage drop among the terminals of R_s is approximately zero [31,32]. By inspection of Fig. 2(b), we get

$$\begin{aligned}&v_y(t) = \pm Bp_j, \quad i_x(t) = \frac{x(t) - \pm A_{v1}Bp_j}{R_x + R_{an}}, \quad i_z(t) = -\frac{v_z(t)}{R_b}\\ &v_w(t) = V_n(x(t)), \quad i_w(t) = i_n(x(t)) = -\frac{V_n(x(t))}{R_{cn}}, \quad \frac{di_w(t)}{dt}\\ &\quad = -\frac{1}{R_{cn}}\frac{dV_n(x(t))}{dt}\end{aligned} \tag{5}$$

where Bp_j is the jth breakpoint [22,24] and each positive or negative Bp is a DC voltage source which is used to compare if $x(t)$ is larger or smaller than Bp. Substituting each variable of (5) into (1) yields

$$\frac{dV_n(x(t))}{dt} = GB\Big(\pm A_{v1}Bp_j - x(t)\Big) - \frac{GB}{A_{DC}}V_n(x(t)),$$

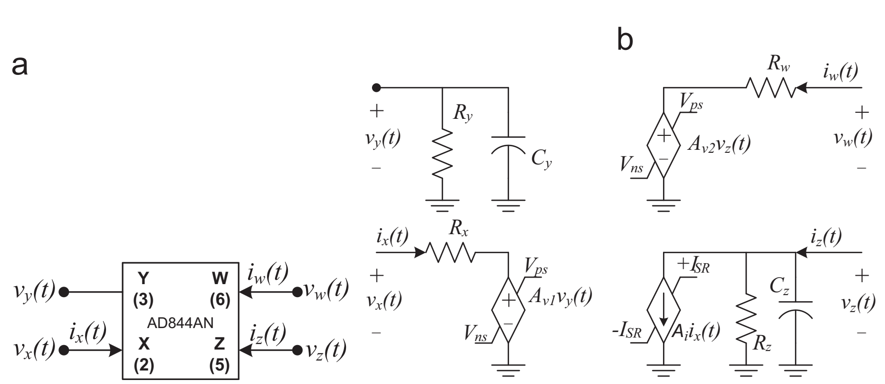

Fig. 1. (a) CFOA circuit symbol. (b) Equivalent nonlinear macromodel.