Current-mode BiCMOS sliding-mode controller circuit for AC signal generation in switching power DC-DC converters

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Abstract - **This communication describes the design of a microelectronic BiCMOS analog circuit, intended to be applied to the sliding-mode control of switching DC-DC power converters for the application of generating a sinusoidal signal with externally adjustable amplitude, frequency and offset. The circuit implements a sliding law over a nonlinear autonomous switching surface. The proposal of this analog design, which operates in current-mode, represents good performance as long as operation speed, power consumption, suitability for low-voltage operation and interference robustness are concerned. The main circuit is composed of externally linearized transconductors based on current conveyors, current-mode full-wave rectifiers and transimpedance high-speed comparators, as well as bipolar translinear squaring cells. HSPICE transistor-level simulations for a BiCMOS 1.2μm technology validate the functionality of the proposed sliding-mode controller implementation.**

I. INTRODUCTION

The idea of considering multidimensional nonlinear functions for Sliding-mode control techniques ([1], [2]) appears as a plausible alternative to the PWM control strategies in the area of switching DC-DC regulators, since they improve the robustness of these power conversion circuits against perturbations, both input voltage and load variations. The work presented herein considers the implementation of an autonomous (independent of the time variable) nonlinear switching surface, and its associated sliding control law in order to obtain the generation of sinusoidal signals in a Buck converter [3]. Usually, in regulator applications, the sliding surface is generated as a linear combination of the converter state variables and a set of reference values ([1], [4]), hence constituting an hyperplane in the space state. The surface proposed in [3], however, is nonlinear and does not require external reference. In the aforementioned previous work [3], and, considering as power plant the Buck converter depicted in figure 1(a), (where an input switching bridge is added in order to assure bipolarity in the generated signal), it is demonstrated that, when the output sinusoidal signal is described by

$$V_{out}(t) = A sin(\omega t) + B \qquad (A > 0) \qquad (1)$$

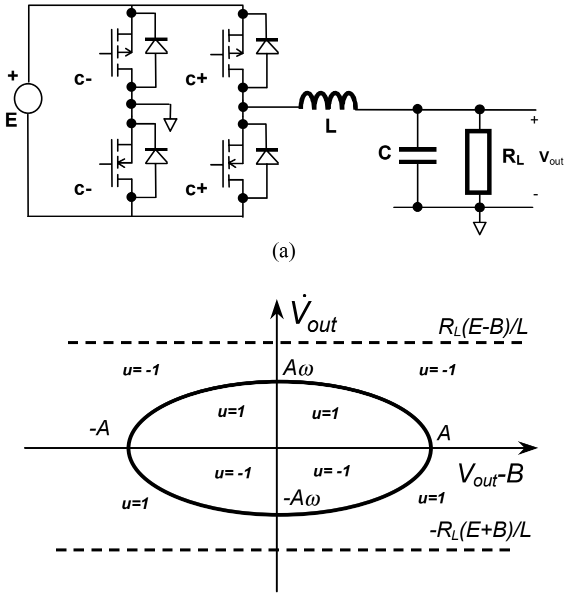
this oscillating output signal can be obtained as a sliding regime when forcing the following sliding surface:

$$\sigma(V_{out}, \dot{V}_{out}) := \dot{V}_{out}^2 + \omega^2 (V_{out} - B)^2 - \omega^2 A^2 = 0 \qquad (2)$$

In addition to this, in [3], the state-space zones that constitute the sliding domain are derived, as well as the control switching law, defined by

$$u = \begin{cases} +1 & if \quad \sigma \cdot \dot{V}_{out} < 0 \\ -1 & if \quad \sigma \cdot \dot{V}_{out} > 0 \end{cases} \qquad (3)$$

where it is considered that $u=\pm 1$ implies $\pm E$ as the converter input-filter voltage. All this results are summarized in figure 1(b), where there are shown, in the state-plane, the sliding surface –ellipsoidal curve-, the sliding domain, as well as the value that the control variable *u* should take in accordance to the law defined by expression (3).



(a)

(b)

Fig 1. (a) Buck converter with input bridge (b) State-plane representation of the control law and the sliding surface.

After reviewing the theoretical aspects, what follows is both the synthesis of the controller circuit and the analysis of the subcircuits which conform it. Before considering design aspects, we proceed to discuss the different technology and design options, namely, analog versus digital, and current-mode versus classical voltage-mode design.

On one hand, being the sliding control an instantaneous control technique with multiple inputs, which does not require excessive processing complexity, digital signal processing presents the disadvantages of the associated computation time, together with the necessity of including A/D and D/A converters. Nevertheless, there exist implementations based on look-up-table mapped on EPROMs [3], or by means of microcontrollers [5] that have shown to be effective for low switching frequencies.

On the other hand, the few existing analog implementations of sliding control laws [6] make use of classical voltage-mode processing, *i.e.*, high-gain operational amplifiers are operated in

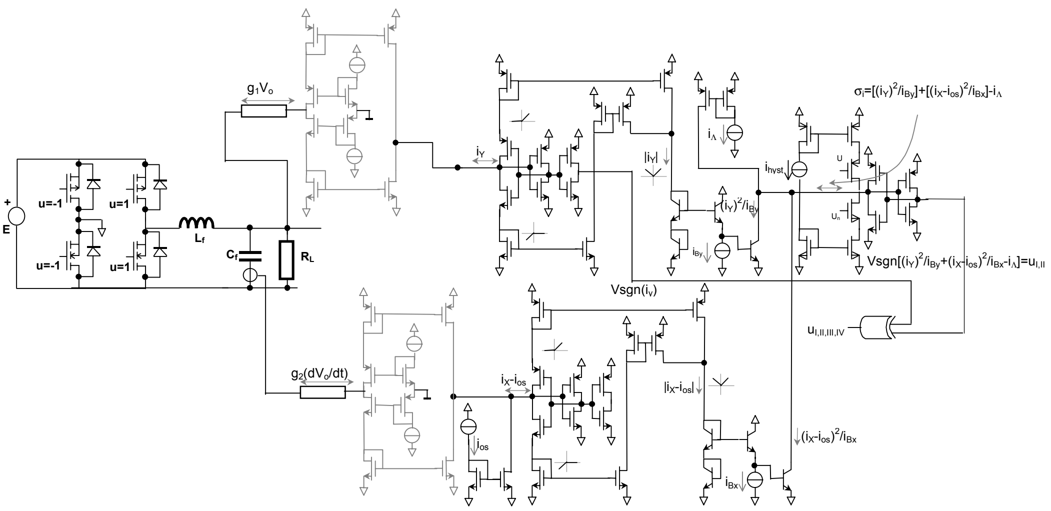


Fig. 2. (a) Diagram of the current-mode sliding AC generation controller.

closed-loop negative-feedback configurations, which results in bandwidth limitations, both small-signal and large signal -slew rate-. In front of this limitations, current-mode design appeared (see [7]) in order to overcome, mainly in the area of microelectronic design, the inherent limitations of voltage-mode processing. The use of currents as information-carrying signals yields reduced voltage variations in the circuit (this is because the processing is done in low-impedance nodes whose voltages are practically constant for nonlinear high-compressing active devices as bipolar transistors or large aspect-ratio MOSTs), and thus, in a reduced impact of parasitic capacitances which represents an increment in the maximum operation speed of these circuits. Apart from this key effect, other advantages of current-mode processing are the local open-loop high-bandwidth processing and the capability of directly implementing signal aggregations, by virtue of Kirchoff's current law. As opposite to this, some disadvantages are related to current-mode processing, namely, the accuracy errors due both to the loss of precision associated to the loss of negative feedback, and the need to use mismatched current replicas when copying signals. Note, however, that this loss of precision reduces its impact in controller implementation applications, compared to the design of open-loop signal processors (*e.g.* filters), since in the former the global feedback lowers the precision requirements. Last but not least, a standing feature of current-mode operation is that, due to the operation between low-impedance nodes, it is more robust to interferences or crosstalks in a mixed digital/analog or switching microelectronic environment, as it is the case of an ON/OFF controller intended to control switching converters, which delivers a pulsating output signal.

Recently, current-mode design has been used in the area of switching power converter controllers, as for instance in [8], where the 'true' current-mode translinear principle associated to bipolar transistors is considered for the design of power-factor-correction (PFC) circuits.

II. BICMOS ANALOG PROCESSING BLOCK IMPLEMENTATION DETAILS

The first stage of the processor consists of sensing the converter's output voltage signal (V_o), and the voltage signal due to the Buck converter capacitor current (coming this either from a Hall sensor or a low-ohmic sensing resistor), which is proportional to dV_o/dt, and the subsequent conversion into current-mode representation via transconductors. The processing required to implement the nonlinear equation described by equation (2), (3) and figure 1(b) consists in a squaring relation, signal aggregations and hysteretic comparison. An important fact that the controller under study considers is the symmetry that the shape of the sliding surface shows in the phase-plane. Hence, the controller, after transconducing external voltage signals into current, must detect the sign (or polarity) of incoming signals –so as to detect the operation quadrant-, and rectify them before obtaining the ellipsoidal comparation edge. Subsequently, and after the hysteretic comparison, the information about the phase-plane quadrant position is used to recover the correct value of the discrete ON/OFF output signal. The aforementioned processing is realized in a compact way after the circuit depicted in figure 2.

From this diagram, it is straightforward to obtain, as a function of signals and references of the circuit, as well as their relationship with the sinusoidal waveform defined in (1), the equation which models the ellipsoidal segment (sliding surface σi) due to just one quadrant, yielding

$$\sigma_i = \frac{(i_Y)^2}{i_{B_Y}} + \frac{(i_X - i_{os})^2}{i_{B_X}} - i_\Lambda = 0$$

$$B = \frac{i_{os}}{k_X}, \quad \omega = \left(\frac{k_X}{k_Y}\right)^2 \left(\frac{i_{B_Y}}{i_{B_X}}\right), \quad A = \frac{1}{k_X} i_{B_X} \sqrt{\frac{i_\Lambda}{i_{B_Y}}} \tag{4}$$

$$\begin{cases} i_Y = k_Y \dot{V}_{out} \\ i_X = k_X V_{out} \end{cases}$$

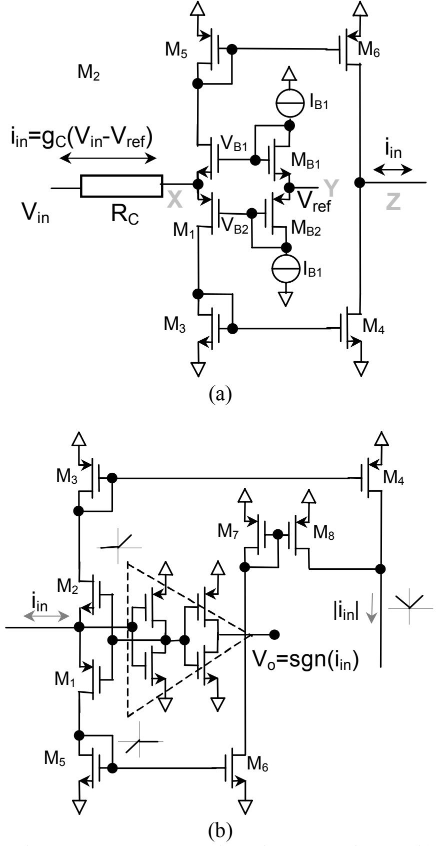
Below, the operation and characteristics of each subblock constituting the sliding controller are described.

A. Current-Conveyor-Based transconductors

As long as the signals internally processed by the controller are currents, the external signals in the converter, in voltage-mode, are converted into currents by means of transconductor circuits. The circuit option chosen is depicted in figure 3(a). In contraposition to the design of linearized transconductors through the use of compensation techniques of the nonlinear characteristics of the transistors in the active block (see, for instance, chapter 5 in [7]), for the circuit depicted in figure 3(a) the transconductance action is inherently linear and due to an external resistor, R_C, an element which is already required to adapt signal levels in the converter to the dynamic range of the controller. Instead of using negative feedback over a voltage-gain element (as is the case of classical opamp-based differential amplifiers), the V_{ref} voltage is imposed to one of the terminals of the R_C resistor by a circuit structure of the current conveyor (CCII) type [9], and the sensed current iin is steered and recovered at the output node (Z) by virtue of two complementary current mirrors (M3-6). If the whole circuit is interpreted as a class AB current mirror, the voltage levels $V_{B1,2}$ forced by transistors $M_{B1,2}$ are to be shared for each transconductor stage. This use of a resistor as an off-chip transconductor has been recently used as front-end for current-mode signal processors (mainly switched-current circuits [12], [13], or as high-speed, high-linear low-voltage continuous-time V-I converters [14]).

B. Current rectifiers

The design of the current-mode full-wave rectifiers -fig 3(b)- is based on the so-called current switch, a circuit whose core is a high-speed current-input voltage-output comparator [10], which exhibits a nonlinear low-impedance input node that senses and detects the sign of the input current. Taking into account that the half-wave rectified currents through each branch –M3,4 y M5,6- are steered and recovered at the output node, and accounting for the polarity inversion due to an extra current mirror –M7,8- , full-wave rectification is achieved for the input current signal. The capacitive-input offset-free transimpedance comparator avoids dead-zone in the nonlinear signal transfer, while showing response times well under 100ns. If a certain mismatch between MOS transistors is considered (mostly in current-steering mirrors), the rectifier would show a certain asymmetry in its transfer characteristics.



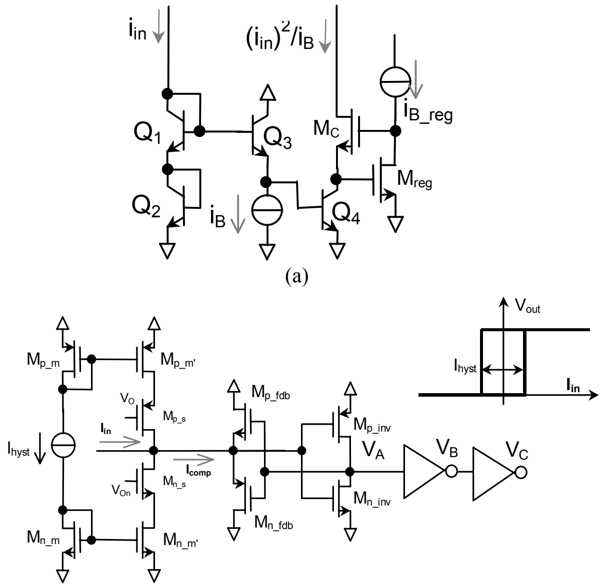
(a)

(b)

Fig. 3. (a) Class AB CMOS CCII-based transconductor (b) CMOS current-mode full-wave rectifier

C. Translinear Squaring Cells

Taking advantage of the translinear principle connatural to active elements with exponential voltage-current characteristics (bipolar transistors or MOS transistors operated in weak inversion), we have considered the squaring cell depicted in figure 4(a), which produces an output current after $i_o=i_{CQ4}=(I_{in})^2/i_B$ [7], [11]. The doubled base-emitter voltage of the input diode branch results in a quadratic transfer for the output current of the cell. The selection of a circuit based on bipolar transistors, as opposed to a structure based on the quadratic relationship of MOS transistors, is due to the better characteristics of the former circuit in front of the latter, in terms of dynamic range and bandwidth. MOS-based regulated cascode techniques are applied at the squarer's output transistor to lower output conductance to a negligible level. This effect is further reduced by the low-impedance input node of the following stage, the transimpedance comparator.



(a)

(b)

Fig. 4. (a) Bipolar Translinear squaring cell (b) Hysteretic current-input transimpedance comparator.

D. Hysteretic current-input transimpedance comparator

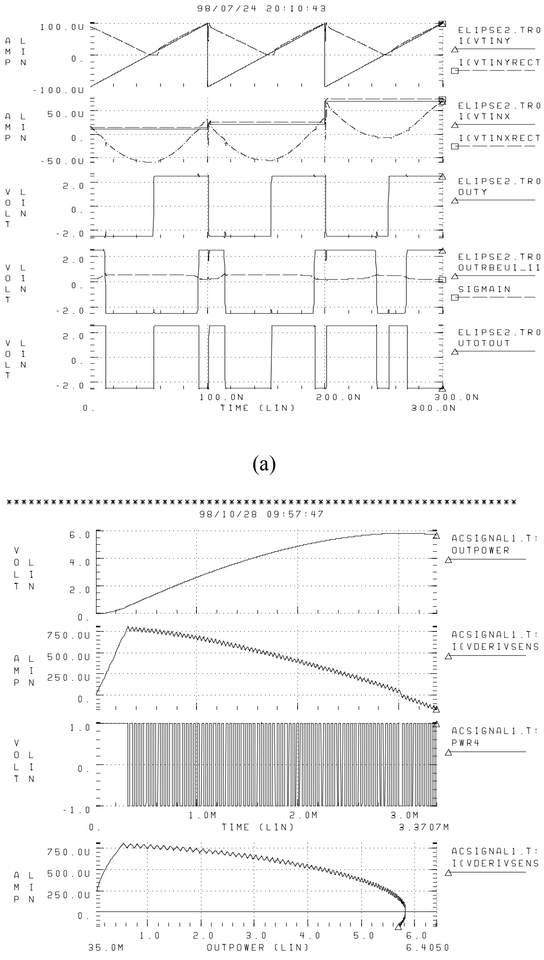
It is well known that actual implementations of sliding control require the addition of a certain hysteresis level to the output comparator which is in charge of continuously evaluating the σ_i combination, so as to avoid sliding regimes with infinite switching frequencies. In order to achieve this effect, the hysteretic current-mode comparator shown in figure 4(b) is considered as the output stage of the controller. The core of this circuit is constituted by a current-input voltage-output comparator which includes a CMOS inverter (M_{n,p_inv}) that confers it zero offset. A pair of added transistors (M_{n,p_fdb}) provide a class-AB feedback path, which results in a reduction of the input impedance as well as a much faster response. The subcircuit responsible of adding hysteresis is based on a bidirectional current mirror ($M_{n,p\_m,m'}$), whose output is switched as a function of the output state V_C of the comparator, yielding a very compact circuit implementation with short response times.

III. SIMULATION RESULTS

In order to validate both the functionality of the approach and the feasibility of the proposed microelectronic implementation, figure

5(a) shows a global transistor-level simulation of the whole controller, using the technology parameters and models supplied by the manufacturer for a standard BiCMOS technology (AMS 1.2µm). The controller circuit is excited in open-loop by means of a sawtooth input signal (i_Y) and step currents (i_X), so as to explore the bidimensional input space. Even though the frequency of the input signal i_Y is high (10Mhz) in order to emphasize the advantages of the current-mode design approach, it can be observed the capability of the circuit to rectify the signal (plot 1) as well as obtaining a bit coding of its sense (plot 3), to obtain the squaring conformation (plot 2), implement the hysteretic comparison ON-OFF signal (plot 4), and finally to obtain the control signal as a function of the operation state space quadrant (plot 5).

Figure 5(b) shows the response of the Buck power converter (modeled with ideal switches), when the sliding controller circuit is inserted in the control loop. The evolution of the output voltage signal of the converter (plot 1), and the current through the filter capacitor (plot 2), as well as their related state-plane representation (plot 4), follow the expected behavior of sinusoidal variations as a function of time and ellipsoidal variations in the state-plane. Plot 3 in this figure depicts the output ON-OFF state of the hysteretic comparator driving the state of the power transistors in the input bridge of the switching DC-DC converter.



(a)

(b)

Fig. 5. Simulation results (a) Open-loop operation of the controller fin=10Mhz (b) Closed-loop operation of a controller and the Buck power converter.

IV. CONCLUSIONS

This work proposes a microelectronic implementation of a sliding-mode controller for the generation of variable-parameter sinusoidal waveforms. This proposal finds application in the design of power conditioning systems as uninterruptible power supplies (UPS). In order to overcome the difficulties imposed by high switching frequencies, as well as the requirements of the inherently instantaneous sliding-control mode –compared to low-frequency linearized PWM control-, the use of current-mode analog design approach is presented. Global transistor-level simulation results confirm both the correct operation of the circuit in terms of bandwidth and also its functionality for the control of a Buck power converter. Work in progress includes the design of the circuit at full-custom layout level and its implementation as an ASIC sliding controller.

ACKNOWLEDGEMENT

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