Analysis of a Soft-Switched PFC Boost Converter Using Analog and Digital Control Circuits

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***Abstract*—This paper presents a comparison between analog and digital (PIC16c73a) control types applied to the boost converter with a nondissipative snubber. Both control types use the bang-bang hysteresis current waveshaping control technique in order to achieve a quasi-unity power factor. The analog control applied presented a high power factor (0.998), high efficiency (92.87%), and low harmonic distortion [total harmonic distortion of current (THDI = 2.84% and total harmonic distortion of current (THDV) = 2.83%)]. The digital control presented a high power factor (0.990), high efficiency (92.46%), and low harmonic distortion (THDI = 5.09% and THDV = 2.84%).**

***Index Terms*—AC–DC power conversion, harmonic distortion, power factor.**

I. Introduction

One of the most interesting areas for researchers in power electronics is the use of microprocessor circuits replacing the existent analog circuits employed in converter control.

Due to the increasing progress in microprocessor circuitry, there have been high technology components available that are perfectly suitable to converter control. The employment of such components reduces the number of electronic devices, implying high switching frequencies, and expanding the converter operation according to the need of the design

One of the most pertinent demands regarding equipment that uses one or more active switches, made by utilities and power quality committees, is that as this equipment introduces a high harmonic content in the power system, it thus should meet the existing standards related to power factor and harmonic distortion.

Several authors worldwide have presented many works related to the use of control strategies [1], [2] in order to meet the international standards. Some of them employ analog control [3], and others employ digital control [4].

Manuscript received March 10, 2003; revised July 5, 2004. Abstract published on the Internet September 10, 2004.

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Digital Object Identifier 10.1109/TIE.2004.837909

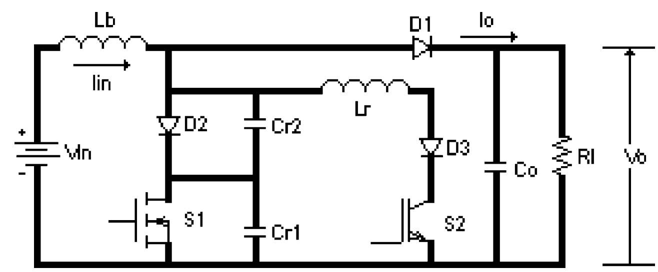

Fig. 1. Boost converter associated to a nondissipative snubber.

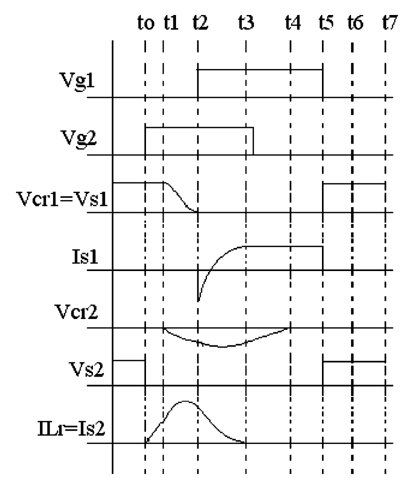

Fig. 2. Theoretical waveforms for the boost converter associated to a nondissipative snubber.

Another important factor to be noticed is the reduction of electromagnetic interference levels (EMIs). A simple way of solving this problem is the use of switching techniques that employ null current and/or null voltage [5], [6]. These techniques increase the converters' efficiency and switches utility life.

This paper presents a pulsewidth-modulation (PWM) boost converter with nondissipative commutation [6] using both analog and digital control types and applying the control strategy by current imposition called hysteresis "bang-bang" [2]. It also seeks a simple comparison between the analog and digital control types, contributing to the development of works related to the use of microprocessors in converter control.

II. Preregulator Boost Converter

Fig. 1 shows the simplified schematic circuit of the proposed nondissipative snubber associated to a boost converter. This converter operates without commutation losses.

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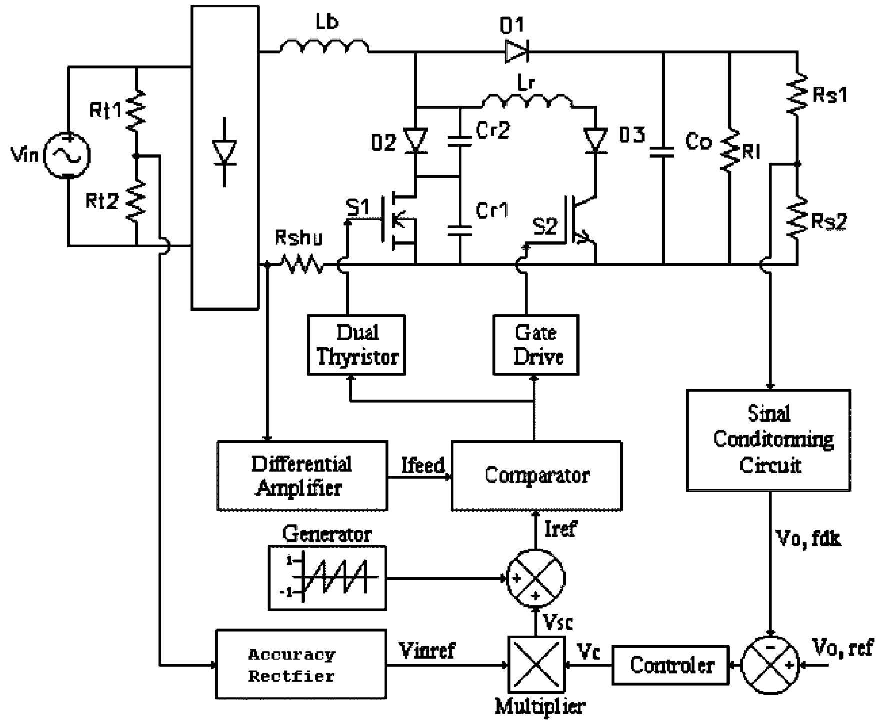


Fig. 3. Boost converter associated to a nondissipative snubber and the analog control circuit.

Switches S_1 and S_2 commutate softly. Switch S_1 commutates in a zero-voltage-switching (ZVS) way and switch S_2 commutates in a zero-current-switching (ZCS) way.

III. Principle of Operation

A complete theoretical analysis for the approach shown in Fig. 1 is presented as follows. Since both converters present the same principle of operation, a single boost converter will be analyzed.

The analysis begins with the description of seven operational stages that describe a complete switching cycle.

- *First stage* (t_0–t_1)*:* This stage begins when switch S_2 is turned on in a ZCS way. During this stage, the resonant current I_{Lr} increases linearly. This stage finishes when I_{Lr} is equal to output current I_{in}.
- *Second stage* (t_1–t_2)*:* When resonant current is equal to input current, this stage begins. This is the first resonant stage, where both resonant capacitors (C_{r1} and C_{r2}) are in resonance with L_r. During this stage, capacitor C_{r1} is discharged and capacitor Cr2 is charged, which finishes when C_{r1} is fully discharged, allowing switch S_1 to be turned on in a ZVS way.
- *Third stage* (t_2–t_3)*:* This is the second resonant stage. During this stage, capacitor C_{r2} is in resonance with inductor L_r. This stage finishes when the resonant current I_{Lr} reaches zero. In this stage, switch S_1 is turned on in a ZVS way.
- *Fourth stage* (t_3–t_4)*:* This stage begins when current I_{Lr} is equal to zero. In this stage, resonant capacitor C_{r2} is linearly fully discharged by input current I_{in}. During this stage, switch S_2 can be turned off in a ZCS way.
- *Fifth stage* (t_4–t_5)*:* When capacitor C_{r2} is fully discharged, this stage begins. During this stage, input voltage transfers its energy to boost inductor L_b. This stage finishes when switch S_1 is turned off in a ZVS way.
- *Sixth stage* (t_5–t_6)*:* When switch S_1 is turned off, capacitor C_{r1} is linearly charged by input current I_{in} up to output voltage V_o, which represents the end of this stage.
- *Seventh stage* (t_6–t_7)*:* During this stage, the stored energy in the boost inductor L_b is transferred to the load. This stage finishes when a new switching cycle begins.

From the operating stages described above, one can obtain the waveforms shown in Fig. 2. The transfer function between V_o and V_{in} is given by (1) (see Appendix A)

$$G = \frac{V_o}{V_{\text{in}}} = \frac{1}{1 - \left\{D + \frac{K_1}{2\pi}\left[\arccos\left(-\frac{1}{X}\right) + \frac{X+1}{X\alpha} + \frac{\alpha}{2}\right]\right\}} \tag{1}$$

f_s switching frequency ;
f_0 resonant frequency;
D duty cycle;

and

$$\alpha = \frac{I_{\text{in}}}{V_o}\sqrt{\frac{L_r}{C_r}} \tag{2}$$

$$K_1 = \frac{f_s}{f_0}. \tag{3}$$

IV. Control Strategy

The block diagram of the analog control circuit block diagram is shown in Fig. 3, and the digital control circuit is shown in Fig. 4. This converter operates with constant switching frequency and high power factor, using a bang-bang current control strategy.

The input current and line voltage samples are obtained from Rshu and Rt1/Rt2 sensors. The voltage sample is rectified in the Accuracy Rectifier block.where

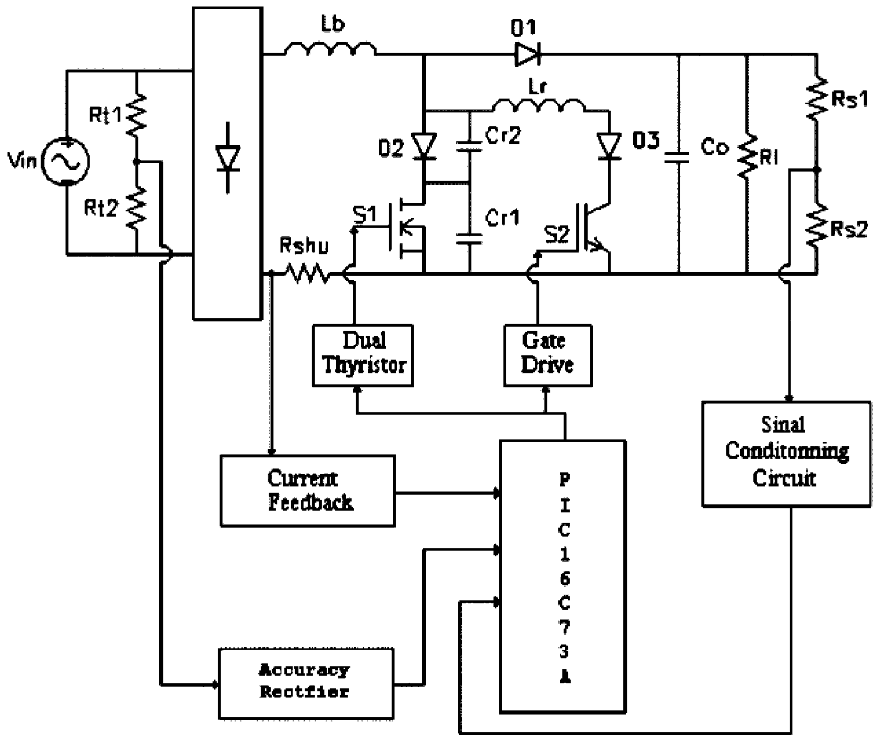


Fig. 4. Boost converter associated to a nondissipative snubber and the digital control circuit.

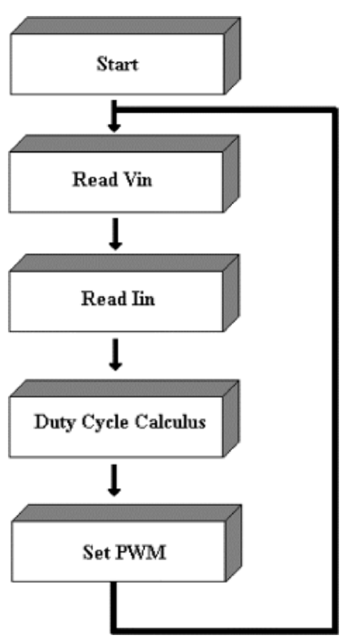


Fig. 5. Flow chart representation of the employed algorithm.

The Signal Conditioning Circuit is a filter. The controller is implemented to provide the control signal (Vc), which is multiplied by Vinref. Then this signal is added to the sawtooth signal (produced by generator block) generating the reference current signal (Iref). The comparator block will produce the pulses to gate switches. The drive signals are obtained by comparing the current feedback signal, generated in the sensor Rshu, with the reference current signal.

The signal obtained from the comparator block output drives auxiliary switch S2 directly. The same signal will drive switch S1, but only when the zero voltage transition on the switch is satisfied.

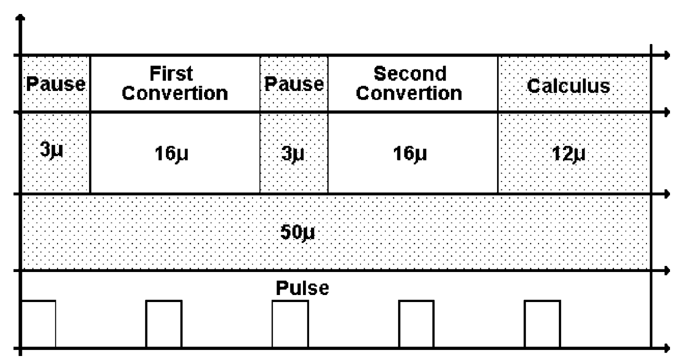


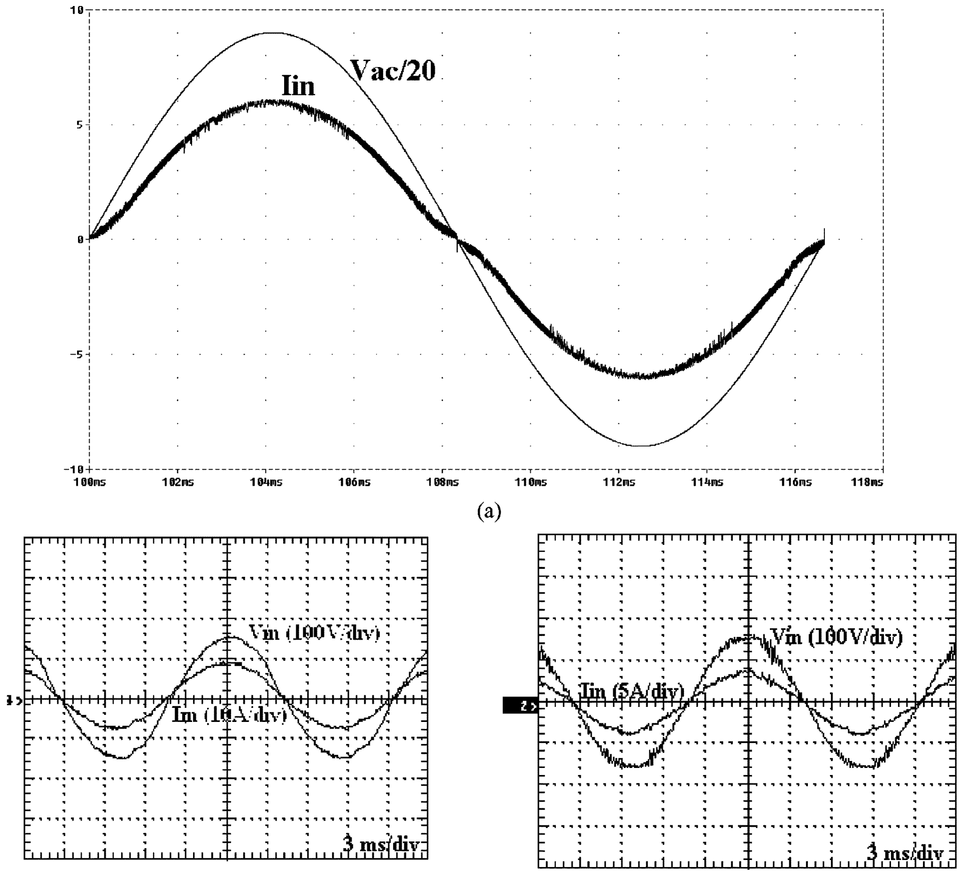
Fig. 6. Total time diagram.

In order to establish a more accurate comparison between the analog and digital control types, an algorithm with the same features as those presented above was implemented and then applied to microcontroller PIC16c73a.

The algorithm that applies the hysteresis "bang-bang" control strategy follows the flow chart shown in Fig. 5. The "Start" stage is characterized by the declaration of variables. In the stage called "Read Vin," the reading of the analog-to-digital (A/D) channel happens corresponding to the input voltage. In the stage called "Read Iin," the reading of the A/D channel happens corresponding to the input current. However, the conversion time is very sluggish (16 μs each); then the reload of the output voltage was not used.

In the stage called "Duty Cycle calculus," a mathematical algorithm that corresponds to the analog control is used. In the last stage the PIC's PWM channels are set.

Fig. 6 shows the total time diagram. It can be seen that a new pulse update is set after a time interval equal to five times the switching period in order to reach 100 kHz.



(b) (c)

Fig. 7. Input voltage and current under rated load condition: (a) simulated results; (b) experimental results using analog control; and (c) experimental results using digital control.

V. SIMULATION AND EXPERIMENTAL RESULTS

The boost converter associated to a nondissipative snubber and the control circuit was analyzed by simulation using PSpice software with the following parameter set.

$$
\begin{aligned}
S_1 &= \text{IRFP460 (Mosfet)}; & S_2 &= \text{Ideal}\\
L_r &= 2.5\ \mu\text{H}; & C_o &= 680\ \mu\text{F}\\
D_1 \text{ and } D_2 &= \text{MUR1560}; & D_3 &= \text{Ideal}\\
L_b &= 1.5\ \text{mH}; & V_{\text{in}} &= 115\ \text{Vac}\\
C_{r1} &= 10\ \text{nF}; & C_{r2} &= 27\ \text{nF}\\
X &= 2.7; & f_s &= 100\ \text{kHz}\\
P_o &= 600\ \text{W}.
\end{aligned}
$$

A boost converter prototype associated to a nondissipative snubber, using the analog and digital control circuits, was built using the following parameter set:

$$
\begin{aligned}
S_1 &= \text{IRFP460 (Mosfet)}; & C_{r1} &= 10\ \text{nF}\\
C_o &= 680\ \mu\text{F}; & L_b &= 1.5\ \text{mH}\\
S_2 &= \text{IRGBC20F (IGBT)}; & C_{r2} &= 27\ \text{nF}\\
V_{\text{in}} &= 115\ \text{Vac}; & f_S &= 100\ \text{kHz}\\
D_1, D_2, \text{ and } D_3 &= \text{MUR1560}; & X &= 2.7\\
L_r &= 2.5\ \mu\text{H}; & P_o &= 600\ \text{W}.
\end{aligned}
$$

Figs. 7–12 show the simulation and experimental results. As can be seen, the switches' commutations occur without losses

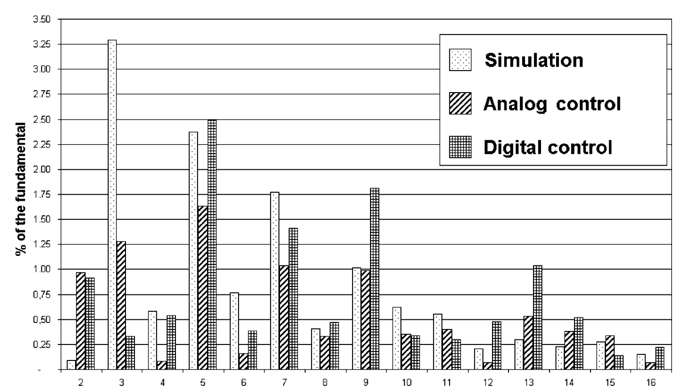


Fig. 8. Input current harmonic distortion.

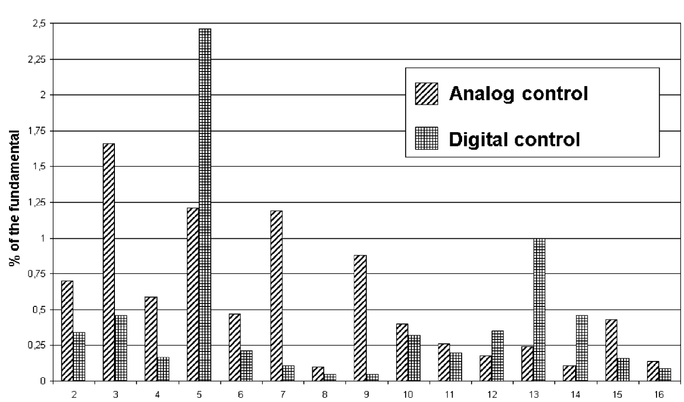


Fig. 9. Input voltage harmonic distortion.

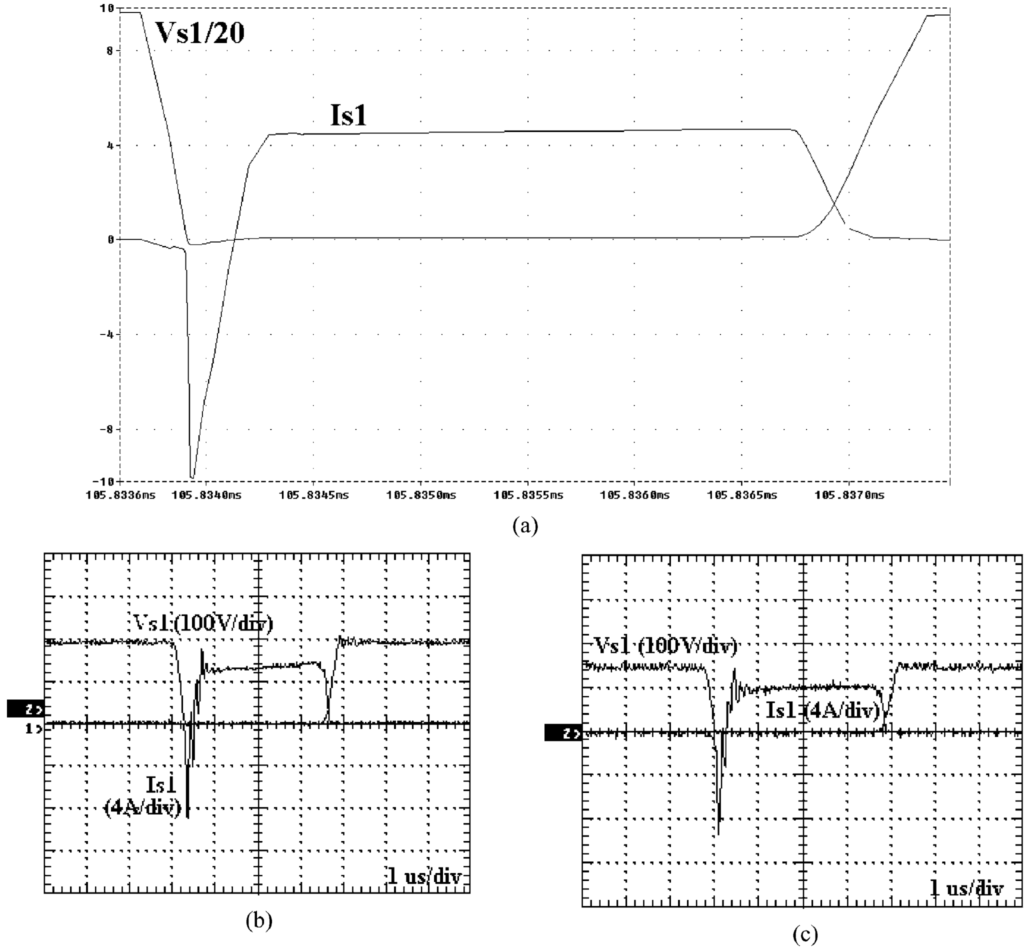


Fig. 10. Switch S_1 waveforms: (a) simulation results under rated load condition; (b) experimental results using analog control; and (c) experimental results using digital control.

and the power factor is almost unity. Fig. 7 shows the power factor correction, under rated load condition. The value obtained for the analog control was 0.998 and that obtained for the digital control was 0.990.

The obtained current and voltage total harmonic distortion rates (THDs) are relatively low, as can be seen in Figs. 8 and 9

In Fig. 9, the current total harmonic distortion rates (THDIs) were calculated in both simulation and experimental analysis, while the voltage total harmonic distortion rate (THDV) was calculated just for the prototypes.

The THDI value obtained in the simulation was 4.85%. The current and voltage THD values obtained for the prototype employing the analog control were 2.84% and 2.83%, respectively. The current and THDV values obtained for the prototype employing the digital control were 5.09% and 2.84%, respectively.

Fig. 10 shows the commutation in the main switch S_1. One can notice that it does not present current and/or voltage stresses, as well as that the commutations are lossless. Fig. 11 shows the commutation in the auxiliary switch S_2. One can see that it does not present current and/or voltage stresses, as well as that the commutations are lossless. Fig. 12 shows the efficiency of the boost converter with both control techniques, which present practically the same efficiency, 92.87% to analog control and 92.46% to digital control.

VI. CONCLUSION

The analog control applied to the boost converter with nondissipative commutation presented a high power factor (0.998), high efficiency (92.87%), and low harmonic distortion (THDI $=$ 2.84% and THDV $=$ 2.83%). Some drawbacks of the analog control are the control circuit complexity, the problems with the gain adjustment, and the increased number of components implying larger cost and weight. Its main advantage is related to the response speed of the control circuit.

The digital control applied to the boost converter with nondissipative commutation presented a high power factor (0.990), high efficiency (92.46%), and low harmonic distortion (THDI $=$ 5.09% and THDV $=$ 2.84%). A digital control disadvantage consists in the need for A/D converters, which reduces its dynamic response; its main advantage lies in the reduced number of components and the system flexibility.

As the digital control brings a reduction in the number of components, the complexity factor is concerned directly with the microcontroller programming. It is also a more versatile control and can be updated and modified in order to assist new demands that may occur. However, the control response speed is limited as a function of the number of A/D converters existent in PICs.

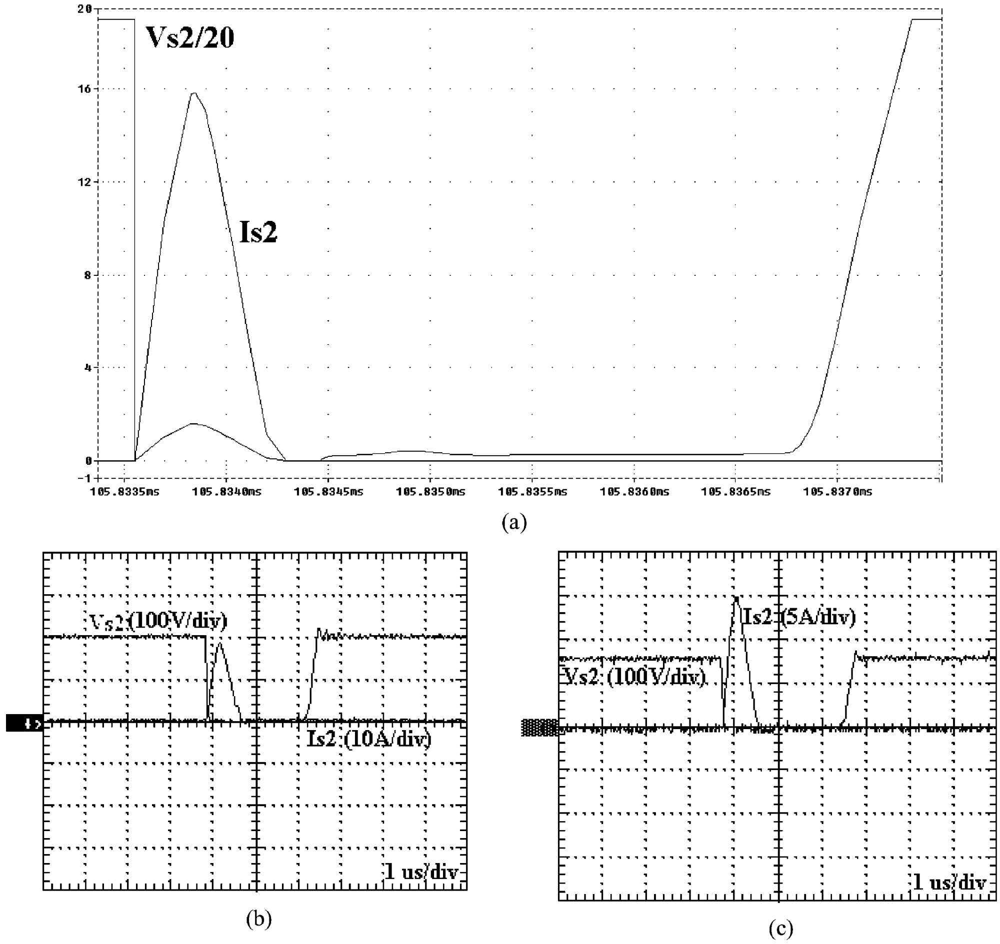


Fig. 11. Switch S_2 waveforms: (a) simulation results under rated load condition; (b) experimental results using analog control; and (c) for experimental results using digital control.

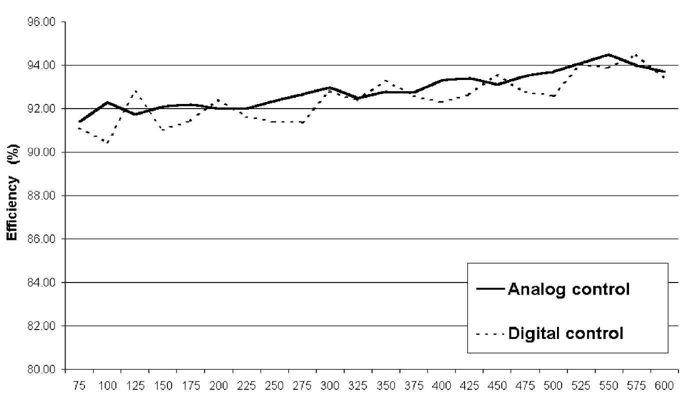


Fig. 12. Efficiency.

APPENDIX

The transfer function between V_o and $V_{\rm in}$ is given by energy conservation

$$P_o = P_{\rm in} \tag{4}$$

$$V_o \cdot i_{\rm omed} = V_{\rm in} \cdot I_{\rm in} \tag{5}$$

$$G = \frac{V_o}{V_{\rm in}} = \frac{I_{\rm in}}{i_{\rm omed}} \tag{6}$$

where

P_o output power;
$P_{\rm in}$ input power;
V_o outpour voltage;
$i_{\rm omed}$ output average current;
$V_{\rm in}$ input voltage;
$I_{\rm in}$ input current.

To reach the output average current ($i_{\rm omed}$), only the first and seventh operational stages were analyzed because in others stages this average current was null

$$i_{\rm omed1} = \frac{I_{\rm in}}{T_s}\Delta T_1 - \frac{\alpha I_{\rm in}}{2\omega_o T_s} \tag{7}$$

$$i_{\rm omed} = \frac{I_{\rm in}}{T_s}\left(T_s - D \cdot T_s - \Delta T_1 - \Delta T_2 - \Delta T_6\right) \tag{8}$$

where

ΔT_y duration time of operational stage y;
T_s switch period;

and

$$\Delta T_1 = \frac{I_{\rm in}}{V_o} L_r \tag{9}$$

$$\Delta T_2 = \frac{1}{\omega_o} \arccos\left(-\frac{1}{X}\right) \tag{10}$$

$$\Delta T_6 = \frac{(X+1)}{X\alpha\omega_o} \tag{11}$$

$$\omega_o = \frac{1}{\sqrt{L_r C_r}} \tag{12}$$

$$\alpha = \frac{I_{in}}{V_o}\sqrt{\frac{L_r}{C_r}}. \tag{13}$$

Therefore, i_{omed} was defined by

$$i_{\text{omed}} = I_{\text{in}}\left\{1 - D - \frac{K_1}{2\pi}\left[\arccos\left(-\frac{1}{X}\right) + \frac{X+1}{X\alpha} + \frac{\alpha}{2}\right]\right\} \tag{14}$$

and the static gain by

$$G = \frac{V_o}{V_{\text{in}}} = \frac{1}{1 - \left\{D + \frac{K_1}{2\pi}\left[\arccos\left(-\frac{1}{X}\right) + \frac{X+1}{X\alpha} + \frac{\alpha}{2}\right]\right\}}. \tag{15}$$

ACKNOWLEDGMENT

The authors thank T. Inpec, Siemens, Texas Instruments, CAPES, CNPQ, and FAPEMIG for their support.

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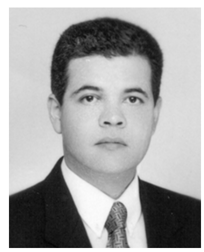

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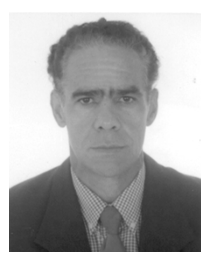

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