Sliding-mode control of power converters feasible to be implemented with low-cost µControllers♦

Domingo Cortes, David Navarro, Analys Barajas, Ismael Araujo

National Polytechnic Institute, Av. Santa Ana No. 1000, Postal Code 04430, México, D.F.
Email: {domingo.cortes, david.navarro.d, analysbs}@gmail.com, iaraujo@ipn.mx.

Abstract: Traditionally, high-end Digital Signal Processors are used to implement nonlinear controllers for power converters. In this paper it is discussed why is necessary to do so and it is shown that if nonlinear controllers are designed to be simple to implement from the beginning then it is possible to implement it using low-cost micro-controllers. Instead of trying to overcome the problems that arise due to digital implementation, a nonlinear controller that achieve an acceptable performance even for slow sampling time is employed. Following this path, it is shown that very efficient, reliable, low-cost, low-power devices for massive applications can be produced.

Keywords: power converters, sliding-mode control, digital control, nonlinear control, energy conversion.

1. INTRODUCTION

Applications of low-cost power converters are ubiquitous. Low power regulators and inverters, small power supplies and power factor correctors, are everywhere, see (Gopinath *et al.*, 2004; Krein and Balog 2002; A. M. Tuckey, 2002; Fossas-Colet and Olm-Miras, 2002; Sira-Ramírez *et al.*, 2002; Chan, 2007). Regardless of its power capacity power converters are switched nonlinear systems with highly variable parameters. Switches are inherent to power converters and make them discontinuous systems. However since commutation of the switch(es) is high continuous model can be obtained by averaging.

When the desired output is constant and load and input voltage variations are not large, small signals models can be used for controller design. However, when the output changes (e.g. inverters), and robustness under significant load and input voltage variation is required, it is convenient and even necessary to use nonlinear (large signal) models. During the last two decades several nonlinear continuous controllers for the average model has been proposed. Most of these controllers are based on the nonlinear average model, see (Garofalo *et al.*, 1994; Escobar *et al.*, 1999; Escobar *et al.*, 2008; Alvarez-Ramírez *et al.*, 2001; Fossas-Colet and Olm-Miras, 2002), some of them are based on the switched model though see (Navarro-López *et al.*, 2009; Cortes *et al.*, 2009; Sira-Ramírez *et al.*, 2001; Sira-Ramírez *et al.*, 2002).

Nonlinear controllers can significantly improve the performance of these power converters see (Escobar *et al.*, 1999; Escobar *et al.*, 2008; Siew-Chong *et al.*, 2008; Siew-Chong *et al.*, 2007; Navarro-López *et al.*, 2009; Cortes *et al.*, 2009). If in addition the controller is digitally implemented some flexibility can be added to applications like warnings and protection signals, autodiagnose, communication with other devices, etc. (Gopinath *et al.*, 2004; Rao *et al.*, 2008; Peng *et al.*, 2007). Hence, it would be desirable to be able to implement nonlinear controllers using digital systems in low cost power converters. However, there are some obstacles that prevent this to happen regularly in the industry as it is shown bellow.

Although there has been efforts to obtain direct discrete models for power converters, they are small signal models, see (Maksimovic and Zane, 2007). As a matter of fact most of the digital control theory is for linear systems. The main cause of this situation is due to obtaining a discrete large signal models of nonlinear systems suitable for control design is very difficult and even impossible. In the rare case that a nonlinear continuous system can be discretized, usually the resulting model is too complicated as is the case for power converters, see (Cortes, 2002). Therefore, direct nonlinear digital control design is not considered as an option. Hence, the path that is usually employed for nonlinear digital control is to design an analog controller based on a continuous-time model, and then implement a digital approximation of such analog controller, hoping that the digital approximation performance to be similar to the analog one. Usually, the digital approximation performs well for small enough sampling time.

However, "small enough sampling time" can not be precisely defined for general nonlinear systems as an acceptable sampling time depends on how fast the system is, what kind of nonlinearities does it have, and the degree of approximation to the original continuous controller we want to achieve. For slow systems the sampling time does not pose any problem because nowadays relatively high frequency sampling time can be achieved even with modest digital systems. This is not the case for power converters whose nonlinear switched model, variable parameters, speed response and requirements such as robustness and reliability, usually needs involved control algorithms. To be digitally implemented with small enough sampling time these algorithms requires large computing power (Kazmierkowski *et al.*, 2011; Nussbaumer *et al.*, 2008; Corradini *et al.*, 2008; Gopinath *et al.*, 2004; Escobar *et al.*, 1999).

For high power, high cost converters, the price of having the required computing power is justified. However, for many

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massive applications like power regulators, small power factor correction, small inverters, etc. having a large computing power is forbidden. Nevertheless dynamics characteristics and requirements of these low power, low cost converters are almost the same than its high power counterparts. That means that complex control algorithms are required regardless the power level of the converters.

In view of previous arguments it does make sense to ask, if it is possible to implement nonlinear control algorithm with small cost micro-controllers (μC). In general, the answer to this question is no, because low cost μC does not have the computing power that most nonlinear controllers requires. This paper aims to show that there is a class of sliding-mode controllers that can be implemented with low cost μC. Note that since it is difficult to know theoretically whether the sampling time achieved with a certain low cost μC will be sufficiently low enough for a good performance, the study must be experimental.

Methods to overcome the problems that arise due to differences between the analog controller and its digital approximation had been proposed in (Nussbaumer *et al.*, 2008; Corradini *et al.*, 2008). These differences are attributed to time lags inherent to digital systems such as 1) the delay time due to the sampling of the control quantities; 2) the one due to the calculation time of the DSP; and 3) the one due to the sample-and-hold function of the pulse-width modulator. However, even when these time lags does not exist there are differences between analog and digital world particularly in nonlinear systems, see (M. and E., 2004; Kotta, 1995). Instead of trying to overcome time lags effects of digital implementation in this paper the attention is placed in the simplicity of the analog function to be approximated.

The kind of controllers used in this paper are Sliding-Mode Controllers (SMC). SMC are known to be robust and reliable. Furthermore, since SMC are discontinuous, they are particularly suitable for power converters see (Navarro-López *et al.*, 2009; Cortes *et al.*, 2009; Sabanovic, 2011 and the references therein). The idea behind the controllers employed in this paper, is the same that the long established current-mode control. That is, to make that the inductor(s) current(s) be a function of the output voltage(s). This idea can be used in several converter topologies. To precisely describe the controller, in Section 2 the boost converter model is presented to be used as an example. In the same Section the controller is also described.

In Section 3 the digital implementation of the controller is described. In Section 4 simulation results are presented and in Section 5 the experimental setup is described and its results are presented. Finally in Section 6 some conclusions are given.

2. CONVERTER MODELS AND ITS CONTROLLERS

2.1 The boost converter

Let us to consider the boost converter as the example to develop the ideas presented in this paper. Similar steps can be followed for other converters. The boost converter is depicted in Figure 1. The transistor and the diode synthesize a switch in such a way that the circuit of the Fig**. 1** is equivalent to the diagram 2. Note that when the transistor is on (Fig**. 1**) the switch of Figure 2 is on the position 0. On the contrary when the voltage at the transistor gate is zero, the switch position is 1. The naming of the switch position in Figure 2 is arbitrary and does not affect the converter behavior. Generally speaking controlling the converter means to measure the output voltage and/or the inductor current to change the switch position to keep a desired output voltage, regardless input voltage and load variations. As is explained in the following section there are two general ways of modeling the converter. Each way give rise to its own control scheme.

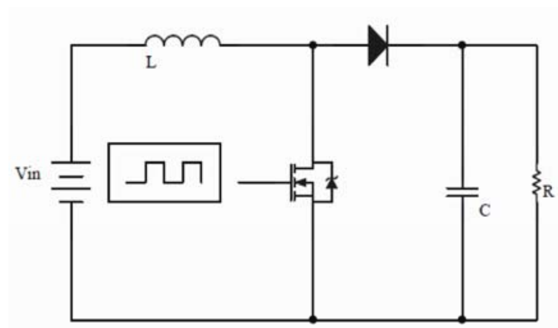

Fig. 1. The boost converter.

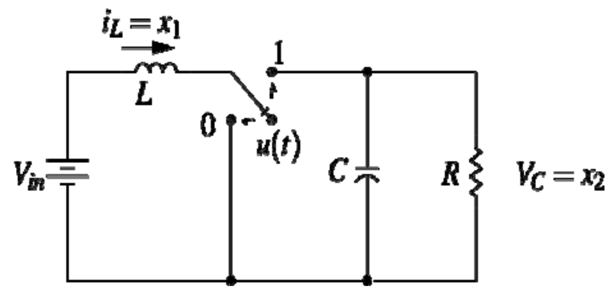

Fig. 2. Schematic diagram of the boost converter.

2.2 Switched model and sliding mode control

With reference to Figure 2, note that when $u = 0$, the converter can be described by

$$\dot{x}_1 = \frac{V_{in}}{L} \qquad (1a)$$

$$\dot{x}_2 = -\frac{x_2}{RC} \qquad (1b)$$

where x_1 and x_2 are the inductor current and the capacitor (output) voltage, respectively. The dot operator means the derivative with respect to time, that is $\dot{x} = \frac{d}{dt}x$. The inductor L, the capacitor C, and the input voltage V_{in} are supposed to be known constants. The resistance R is unknown, but is considered to be constant for analysis purposes. When $u = 1$, the system equations becomes

$$\dot{x}_1 = \frac{V_{in}}{L} - \frac{x_2}{L} \qquad (2a)$$

$$\dot{x}_2 = -\frac{x_2}{RC} + \frac{x_1}{C} \qquad (2b)$$

Combining (1) and (2) into a single equation it can be obtained

$$\dot{x}_1 = \frac{V_{in}}{L} - u\frac{x_2}{L} \quad (3a)$$

$$\dot{x}_2 = -\frac{x_2}{RC} + u\frac{x_1}{C} \quad (3b)$$

Note that u in (3) is a binary variable, $u \in \{0,1\}$. That means that the control is discontinuous.

Due to discontinuous characteristic of model (3) it is suitable to be controlled by Sliding Mode Control (SMC). The general expression of SMC for power converters is given by (see Utkin, 1991).

$$u = \begin{cases} 0 & \text{if } \sigma(\mathbf{x},t) < 0, \\ 1 & \text{if } \sigma(\mathbf{x},t) > 0, \end{cases} \quad (4)$$

where σ is a function of the states and possibly of time. The aim of the switching policy (4) is to force the system trajectories to evolve on the surface $\sigma = 0$. Since the switching policy (4) is always the same, to design a SMC for power converters means to design σ. The general diagram for implementing a SMC is shown in Figure 3.

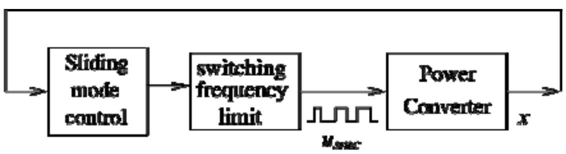

Fig. 3. General diagram to implement a SMC given by (4).

2.3 Average model and PWM control

Considering that position of the switch changes at high frequency it is possible to obtain an average model see (Filippov, 1988) of (3). An average model can also be obtained by heuristic approach by analyzing the circuit operation, see (Kassakian et al., 1991) in both cases results

$$\dot{\tilde{x}}_1 = \frac{V_{in}}{L} - d\frac{\tilde{x}_2}{L} \quad (5a)$$

$$\dot{\tilde{x}}_2 = -\frac{\tilde{x}_2}{RC} + d\frac{\tilde{x}_1}{C} \quad (5b)$$

where $\tilde{x}_1$ and $\tilde{x}_2$ are the average of the inductor currents and the average of output voltage respectively. In this case the control variable is d which is the average of the switch position u and is also called the duty cycle. Since $u \in \{0,1\}$ then d is a continuous bounded signal that is $d \in [0,1]$. Hence the model (5) is continuous. To design a controller for this model means to propose an expression for d. Being d a continuous signal and the control element a switch, a mechanism to translate a continuous signal to pulses is necessary. This is achieved by pulse with modulator (PWM). The general diagram for implementing a PWM based control is shown in Figure 4.

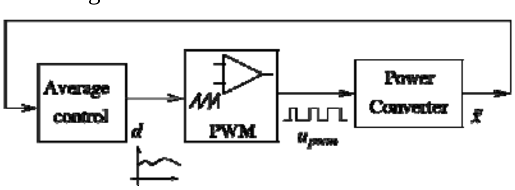

Fig. 4. General diagram to implement a PWM based control

2.4 Combining Sliding Mode Control and PWM control

SMC applied to power converters has the advantage that deals directly with the discontinuous nonlinear model. Hence, does not require averaging neither linearization. However, the switching frequency can be limited but can not be kept constant. In practical applications a constant switching frequency is desirable and even mandatory sometimes. On the other hand, in PWM based control the switching frequency is constant but the controller design and stability analysis frequently use linearization which is a disadvantage. In (Navarro-López *et al.*, 2009) it is proposed a kind of SMC controllers that can be implemented using a PWM. For the sake of clarity and completeness the main idea is described in what follows.

Let suppose the trajectories of a system controlled by SMC evolves on the surface $\sigma(x) = 0$, then in this case $\dot{\sigma}(x) = 0$. In general $\dot{\sigma}(x)$ depends on u. Hence in order to the system trajectories continue to evolve on the surface the expression for u must be the solution for u of the equation $\dot{\sigma} = 0$. Usually this result in a continuous function but from (4) can only have two values. To interpret this apparent contradiction it is said that the solution for u of equation $\dot{\sigma} = 0$ is a fictitious control called equivalent control denoted as u_{eq} which is the average of the real control u. The fact that u_{eq} is indeed the average of u when the system evolves on the surface $\sigma = 0$ is proved in (Utkin, 1991).

On the other hand, on a PWM based control the input of the PWM block d is the average of the pulse u. Hence it could be designed an SMC whose equivalent control $u_{eq}(x)$ be equal to $d(x)$. In other words a PWM control can be seen as an SMC $u_{eq} = d$. More precisely the schemes of Figures 3 and 6 has (approximately) the same performance. Effectively, from the Figure 3

$$\sigma = \int_0^t (\phi(x(\tau)) - u(\tau))\,\mathrm{d}\tau \quad (6)$$

Taking the derivative, results in

$$\dot{\sigma} = \phi(x(\tau)) - u(\tau) \quad (7)$$

hence the solution for u of $\dot{\sigma} = 0$ results in the equivalent control

$$u_{eq} = \phi(x(t)) \quad (8)$$

which can be implemented as the Figure 6

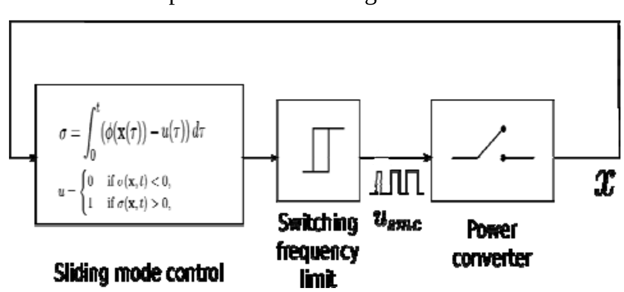

Fig. 5. General diagram to implement a SMC given by (4,6).

The function ϕ has to be determined for every converter. For the boost converter, it is proposed in (Navarro-López et al., 2009) that

$$\phi(t) = G\int_0^t (V_{in} - u(\tau)x_2(\tau))\,d\tau + Gk_p\left(x_2(t) - V_{ref}\right) + Gk_i\int_0^t \left(x_2(\tau) - V_{ref}\right)d\tau \tag{9}$$

where constants G, k_p and k_i are design parameters of the controller that have to be adjusted.

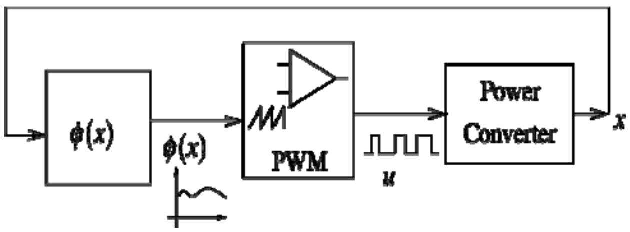

Fig. 6. PWM implementation of the SMC given by (4,6).

Is important to emphasize that the controller is given by (4, 6) and the Fig. 6 is just a way to implement it. It can be though that the controller depicted in Fig. 6 is a PWM controller, and indeed is, but to obtain ϕ a SMC reasoning is followed, which is different than the traditional average control that the PWM is based on. Furthermore, stability analysis can be done using the switched model (3) and controller (4, 6) without relying in the linear average model. This is done in Navarro-López et al. (2009).

2.5 The controller seen as a current-mode controller

Let us show, that current-mode control idea is behind the controller (4, 6, 9). Like in any SMC, the goal of (4) is to make the surface

$$\sigma = 0 \tag{10}$$

attractive. Note that when (10) accomplishes then

$$\phi(x) = u_{ss} \tag{11}$$

where u_{ss} is the steady state average of u, see (Navarro-López *et al.*, 2009) for details. Hence, from (9) and (11)

$$G\int_0^t (V_{in} - u(\tau)x_2(s))\,d\tau = -Gk_p\left(x_2(t) - V_{ref}\right)\ -Gk_i\int_0^t \left(x_2(\tau) - V_{ref}\right)d\tau + u_{ss} \tag{12}$$

Note that from (3a)

$$x_1 = \frac{1}{L}\int_0^t (V_{in} - u(\tau)x_2(\tau))\,\mathrm{d}\tau + x_1(0) \tag{13}$$

From (12) and (13) it can be seen that the controller given by (4, 6, 9) makes the inductor current a function of the output voltage, which is the main idea of current-mode control.

More precisely, if $G = 1/L$, controller (4, 6, 9) makes

$$x_1 = x_{1ref} - x_1(0) + u_{ss} \tag{14}$$

where

$$x_{1ref} = -Gk_p\left(x_2(t) - V_{ref}\right) - Gk_i\int_0^t \left(x_2(\tau) - V_{ref}\right)\mathrm{d}\tau \tag{15}$$

The terms $x_1(0)$, u_{ss} and parasitic elements not considered in the model are compensated by the integral parts of the controller.

2.6 Controller parameters selection

Since the underlying idea of the controller is the same that the current mode controller, the same design methods can be applied to set initial values of k_p and k_i. There are several of this methods reported in the bibliography. One of them is to take (3b), consider the system is close to the steady state, in this case u is almost a constant $\bar{u}$, then consider that the system is controlled by $x_1 = v = x_{1ref}$. Under these assumptions, (3b) becomes the first order linear system

$$\dot{x}_2 = -ax_2 + bv;\ \ a = \frac{1}{RC};\ \ b = \frac{\bar{u}}{C} \tag{16}$$

For this system is easy to design a PI controller v that achieve $x_2 \to V_{ref}$. This v is the x_{1ref} and set the goal for x_1. In this manner, initial k_p and k_i values are obtained. If $G = 1/L$ then the expression $G\int_0^t (V_{in} - u(\tau)x_2(\tau))\,\mathrm{d}\tau$ is an estimated for the current.

Note that if the three parameters are scaled by the same ratio the closed loop dynamics does not change because the surface $\sigma = 0$ remains the same. Hence, G need not be equal to $1/L$ as long as k_p and k_i change accordingly. Once having parameters initial values fine tuning of the controller can be done.

3. DIGITAL IMPLEMENTATION OF A CLASS OF SMC

Till now, controller (4, 6, 9) has been implemented with analog components. A digital implementation would add some flexibility to the intended application, for example: a) warning and diagnose signals can be added. b) communication with other devices becomes possible, c) the power converter could be embedded in a larger digital device and cooperate with the rest of the equipment. In what follows it will be shown that such controller can be digitally implemented using low cost μC with small computing power.

It is important to point out that in the controller depicted in Fig. 6, the construction of ϕ is separated from the PWM block. Hence, the program on the μC only have to approximate (9). Let call the time elapsed between two ϕ calculations the refreshing period of the controller. It is important to point out that this refreshing period does not have to be equal to the switching period.

In the ideal case the refreshing period of ϕ function (T_ϕ), should be faster than the switching period (T_s). However to achieve this ideal, a large computing power is needed.

By examining term by term of the ϕ expression (9) it will be observed that it could be possible to slow the refreshing time without significant performance reduction. This means that μC with low computing power could be used for implementing the controller (4, 6, 9) as consequence the overall converter cost is significantly reduced.

Focusing on the last two terms of (9)

$$Gk_p\left(x_2(t)-V_{ref}\right)+Gk_i\int_0^t\left(x_2(\tau)-V_{ref}\right)\mathrm{d}\tau \qquad (17)$$

it can be noticed that the parameters G, K_p, K_i and V_{ref} are well defined constants so do not present variations. Due to the switching the output voltage x_2 has a rapid variation (called ripple), however this variation has a small amplitude, hence its average changes slowly. Then the term $x_2(t)-V_{ref}$ can be slowly sampled without loosing valuable information.

Since term $(x_2(t)-V_{ref})$ changes slowly, the integral $\int_0^t\left(x_2(\tau)-V_{ref}\right)\mathrm{d}\tau$ can be approximated by the trapezoidal method as follows. Let us call

$$F(t)=\int_0^t f(s)\,ds \qquad (18)$$

then

$$F((k+1)T)=F(kT)+\frac{f((k+1)T)+f(kT)}{2}T \qquad (19)$$

In the previous discussion it has been shown that the terms in eq.(17) could be calculated with low refreshing period without loosing valuable information.

Consider now the first term of (9)

$$G\int_0^t\left(V_{in}-u(\tau)x_2(\tau)\right)\mathrm{d}\tau \qquad (20)$$

which is the estimated inductor current. Since u is rapidly changing between 0 and 1 the term within the integral $(V_{in}-u(t)x_2(t))$ changes between V_{in} and $V_{in}-x_2(t)$ as it is shown in Figure 3. Note that this variation is large. As it ca be observed in the figure if the refreshing period T_ϕ is slower than the switching period T_s very valuable information will be lost, as a consequence the current can no be adequately estimated. One possible solution is to faster the refreshing period T_ϕ however this prevent the use of low cost μC. Note that the term $(V_{in}-u(s)X_2(s))$ is indeed the inductor voltage, hence it can be measured very easily, with some operational amplifiers as it is shown in Figure 8. The circuit of Figure 8 can be though as a cheap current sensor. The output signal of this circuit change slow enough to be sampled slowly. With this solution all the terms of ϕ can be calculated using small computing power and its digital implementation using low cost components becomes possible.

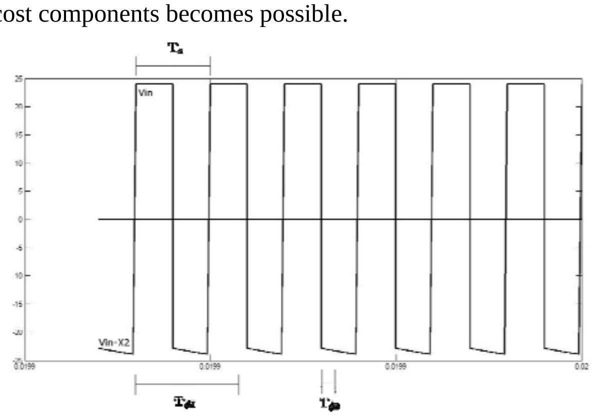

Fig. 7. Inductor voltage on Boost converter.

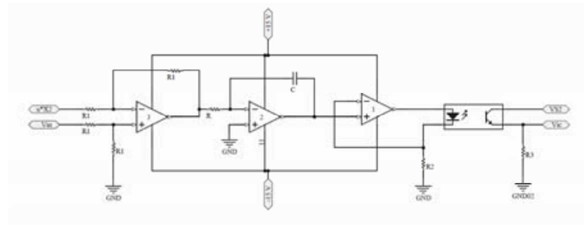

Fig. 8. Cheap current estimator.

By the facts discussed in the introduction, it can not be theoretically assured that a given T_ϕ is sufficiently fast enough to get an acceptable converter performance. Hence experimental tests is mandatory for every application.

For sake of simplicity the boost converter has been used to introduce the SMC proposed to be implemented. However, the same idea of making inductor(s) current(s) a function of output voltage(s) can be applied to other converters as well. In (Cortes *et al.*, 2009) it is applied to the boost inverter proposed in (Cáceres and Barbi, 1999).

4. SIMULATION RESULTS

To evaluate the digital control developed in the previous section some simulations are presented in this Section. The parameters used in this Section are shown in Table 1

Table 1. Design parameter.

f_{sw}	50 kHz
L	$250\mu H$
C	$40\mu F$
R	210Ω
V_{in}	24V
V_{ref}	60V

The PI control parameters are obtained as it is described in Section 2.6. The selected parameters are $k_p=0.0037$ an $k_i=10.125$. Through all the simulations of this Section the load varies between 200Ω and 100Ω to evaluate the performance under sudden load changes.

In Figure 9 the controller performance when the refreshing period is equal to the switching period. Note that the performance is practically the same than the obtained by analog controller. It can also be noticed that under sudden load changes the voltage is slightly perturbed for just some mS. while the inductor current change its stationary value.

It is interesting to observe the performance of Figures 10, 11 and 12. In these figures the controller refreshing period is slower than the switching period. Particularly in Figure 10, the refreshing period is 10 times slower than the switching period. It is must be pointed out that even when the refreshing time is significantly slower, the performance is almost the same than the previous case.

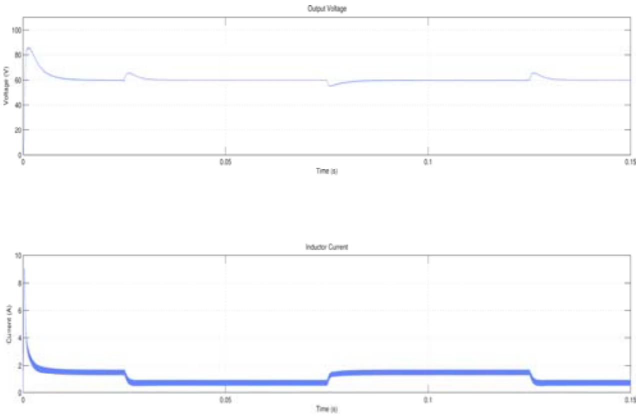

Fig. 9. Controller performance at $T_{\phi} = T_s$ with load variations Up: Output voltage Low: Inductor current.

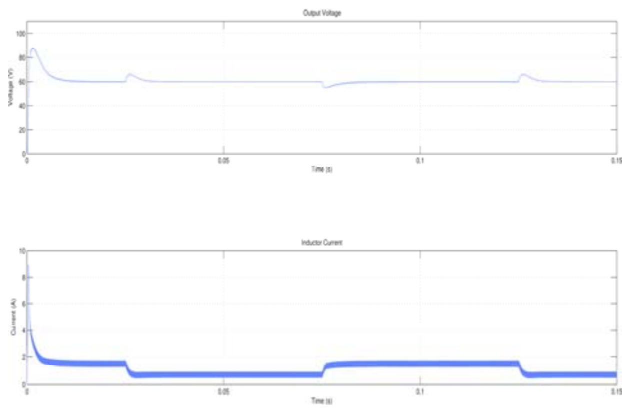

Fig. 10. Controller performance at $T_{\phi} = 10T_s$ with load variations Up: Output voltage Low: Inductor current.

In Figure 11 the refreshing period is 50 times slower than the switching period. The performance is slightly reduced. Nevertheless most changes are observed in the inductor current performance and not in the output voltage.

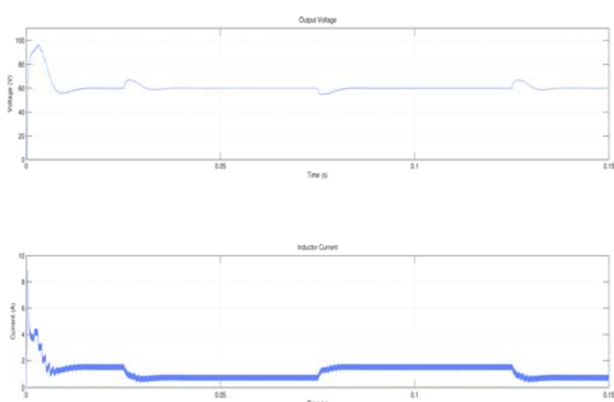

Fig. 11. Controller performance at $T_{\phi} = 50T_s$ with load variations Up: Output voltage Low: Inductor current.

The Figure 12 was obtained using a controller refreshing period 100 times slower than the switching period. As expected, the performance is reduced. However, note that refreshing time is really slow. Such a refreshing time could be achieved by very slow μC, nevertheless the performance obtained could be acceptable for some applications.

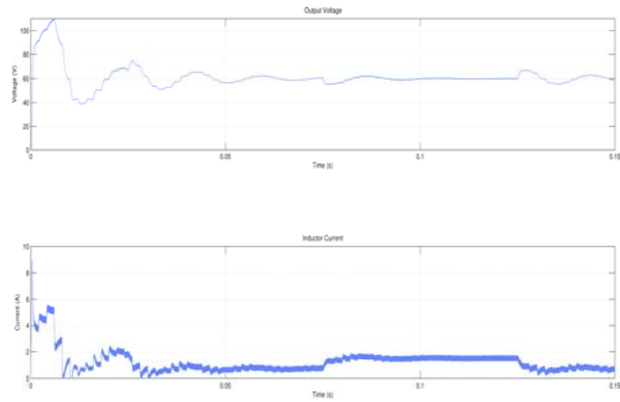

Fig. 12. Controllers performance at $T_{\phi} = 100T_s$ under load variation Up: Output voltage Low: Inductor current.

5. EXPERIMENTAL RESULTS

A modular prototype consisting in the power converter board, a coupling board and a controller board was constructed (see Figure 13). The power converter was constructed using the same parameters employed in Section 4. The coupling board is used to measure the voltage and to estimate the current. The controller proposed in (4, 6, 9) is implemented on dsPIC30F4013 according to the modifications suggested in previous sections. The sliding surface is programed in the μC with $k_p = 0.0037$ and $k_i = 10.125$, these values are different from those used in simulation because include the the sensors gains. As a precaution the the duty cycle is limited to 80% to avoid the inductor be connected to ground for a long time, because its current could become very large and create a short circuit. That is the reason the converter response is bounded at the startup.

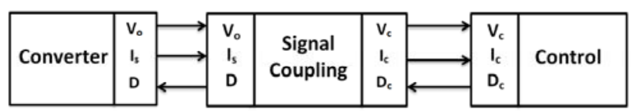

Fig. 13. Modules of constructed prototype.

Figures 14, 15, 16 and 17 present several experimental results obtained when the controller refreshing period is 10 times slower than the switching period. The startup performance is shown in Figure 14. With the exception of the bounded output the response is very similar to that obtained in simulation (see Figure 10). The steady state response is shown in Figure 15. Performance when the load suddenly changes between 200Ω and 100Ω is shown in Figure 16. In the upper part of Figure 16 two voltages are shown, one correspond to the output voltage and the other is proportional to load value. Note that every time that the load changes the inductor current has significant variation, however the change in the output voltage is small and it recover its steady state value after 10ms. To test the controller under more demanding situation the startup with load variation was evaluated. The results obtained are shown in Figure 16.

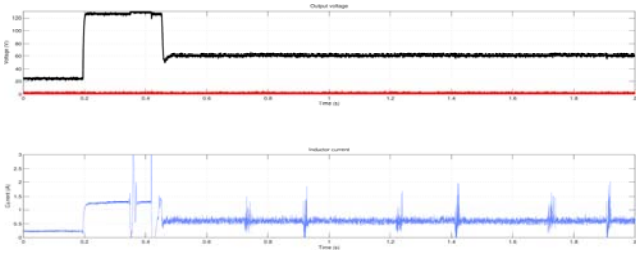

Fig. 14. Experimental startup performance. $T_\phi = 10T_s$.

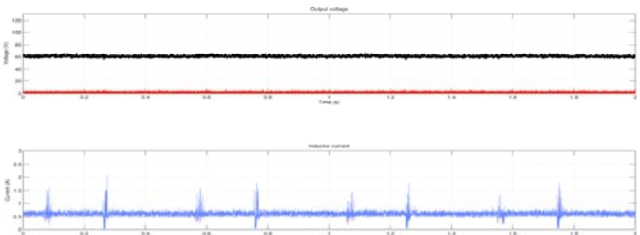

Fig. 15. Steady state response. $T_\phi = 10T_s$.

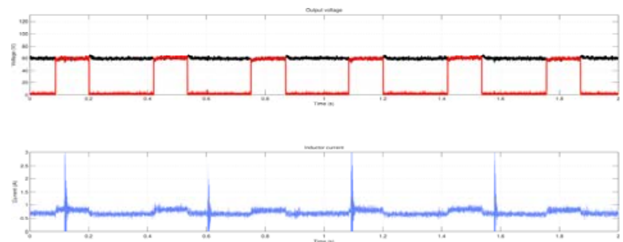

Fig. 16. Performance under load variations. $T_\phi = 10T_s$.

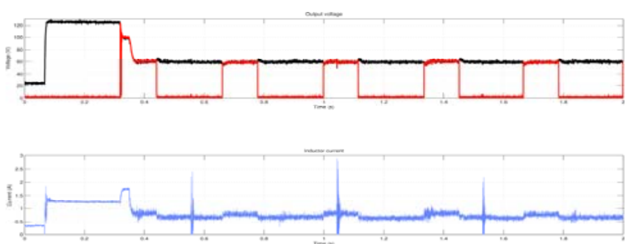

Fig. 17. Startup with load variations performance. $T_\phi = 10T_s$.

Figures 18, 19, 20 and 21 present the same experiments previously described but this time using a refreshing period of 2ms This is an extreme condition because it means that is 100 times slower than the switching period. As it was expected in all cases, there is a reduction in the performance. However for slow systems this is an acceptable performance and can be achieved by using a very low capable μC.

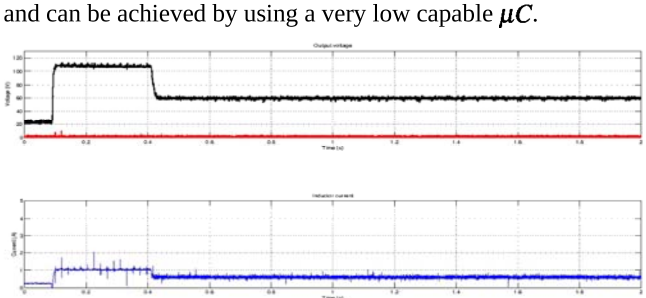

Fig. 18. Experimental startup performance. $T_\phi = 100T_s$.

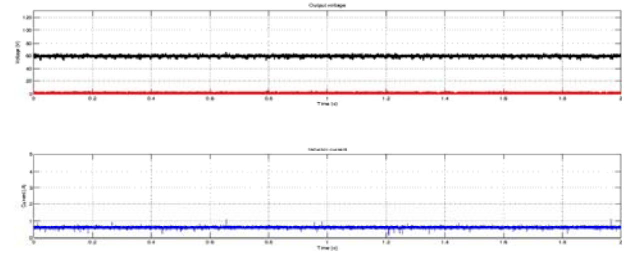

Fig. 19. Steady state response. $T_\phi = 100T_s$.

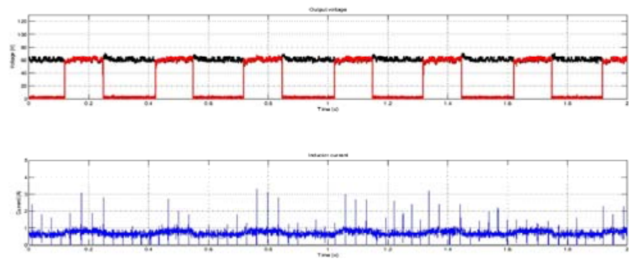

Fig. 20. Performance under load variations. $T_\phi = 100T_s$.

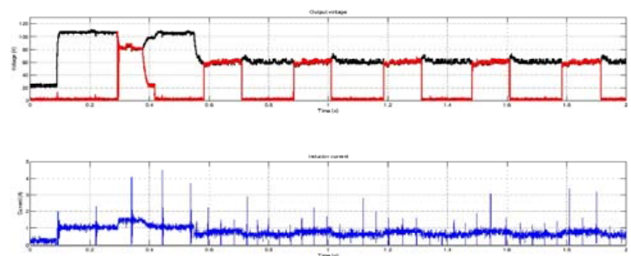

Fig. 21. Startup with load variation performance. $T_\phi = 100T_s$.

6. CONCLUSIONS

It has been shown that it is possible to implement highly nonlinear, robust controllers using low-cost μC if those are adequately designed. This has been done using a class of sliding-mode controllers that are robust to the effects of digital implementation in the sense that its performance decrease is not significant when they are implemented digitally. An alternate paradigm to that of trying to overcome the effects of digital implementation of analog controller can be observed here: that might be better to develop easy to implement controllers that are robust to digital implementation. If this kind of controllers are created then it would be possible to be implemented by low-cost μC. Consequently, highly efficient, digitally controlled power converters could be used in massive applications.

Contrary to the common sense that suggest the refreshing period must be faster than the switching period in digital control of power converter, in this paper has been shown that is possible to slow the refreshing period and yet have an acceptable response.

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