

Robust Step-up DC/AC conversion with a Full-Bridge Non-Inverting Buck-Boost

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Abstract—The substitution of the original switches by a full-bridge in a Non-Inverting Buck-Boost converter results in an inverter capable of performing step-down and step-up tasks under sliding mode control. Furthermore, semi-infinite programming techniques are used to minimize power losses while preventing control action saturation. The performance of the inverter is shown to be robust in front of load perturbations. The procedure assumes known bounds for the disturbances, as well as full state knowledge. Realistic simulations validate the proposed scheme.

I. INTRODUCTION

The possibility of using nonlinear DC-DC switching power converters as source inverters has been thoroughly studied during the last twenty years. The main reason for this is that the conventional Full-Bridge Buck converter must incorporate a transformer to adequately perform step-up tasks, thus resulting in a significative increase of the weight and size of the power supply equipment. When using nonlinear converters, efforts are handicapped by the non-minimum phase character shown by these devices when direct control over the output voltage is exerted [1].

First attempts to solve the problem by means of a current-based indirect control in boost and buck-boost DC-AC inverters [2],[3] have lead to systems with sensitivity to external perturbations and parameter uncertainties. Moreover, direct control strategies of the output voltage that include passivity-based schemes [4] and PID-type sliding mode controllers [5] have been used for regulation purposes. PI controllers also offer interesting performance in full-bridge nonlinear inverters [6],[7],[8],[9]. However, it is well known that PI control designs are based on a small signal model; this leads to output waveforms being sensitive to power stage parameter variations, such as the output load.

The Full-bridge Non-Inverting Buck-Boost inverter, which is essentially achieved by a full-bridge inverter in series with a Boost converter, has two control inputs. Hence, it is possible to design a sliding mode control strategy which is able to yield robust tracking of periodic signals by the output voltage and, at the same time, maintain the input current regulated at a prescribed level. The proposal uses a full-state reference profile-based switching surface that does not depend on the plant parameters. The inverter is assumed to

undergo possible load variations in a set with known bounds. Restrictions for candidate signals to be tracked are derived by demanding non-saturation of the control action, which has fixed control gains, in the steady state.

Existing literature dealing with the Non-Inverting Buck-Boost converter contains results for regulation tasks [10], [11]. In [12], sliding modes are used to achieve output voltage tracking of sinusoidal signals with offset: the performance is shown to be robust under resistive load variation in a set with known bounds. The theoretical development presented in that article is applied in this paper to the Full-Bridge Non-Inverting Buck-Boost converter, this leading to a robust step-up/down DC/AC inversion.

It is already known that proper energy transfer constitutes the main goal of power converters, this meaning good efficiency and high output signal quality. Maximizing power efficiency requires minimization of the Root Mean Square (RMS) of the current flowing in the switching converter, this leading to two different effects, namely, optimization of the losses due to the power switching and minimization of the resistive losses in the inductors. Hence, the design incorporates a procedure to reduce power losses based in semi-infinite programming theory [13] and assuming known upper and lower bounds for the load variation; this technique has been successfully applied to a Full-Bridge Boost converter in [14] and also to a Full-Bridge Non-Inverting Buck-Boost in [12]. Finally, the theoretical development is validated by carrying out illustrative simulations with the realistic power electronics software package PSIM.

The article is structured as follows. The mathematical model of the Full-bridge Non-Inverting Buck-Boost is established in Section II. A sliding mode control strategy to achieve the output voltage tracking target is developed in Section III. The selection of a current reference profile that minimizes power losses and guarantees non-saturation of the controller is studied in Section IV. Simulation results are presented in Section V, while Conclusions and some suggestions for further research are in Section VI.

II. THE FULL-BRIDGE NON-INVERTING BUCK-BOOST CONVERTER

The Non-Inverting Buck-Boost converter, depicted in Figure 1, can be modelled as a two-dimensional, bilinear system with the inductor current i_L and the capacitor voltage v_C as

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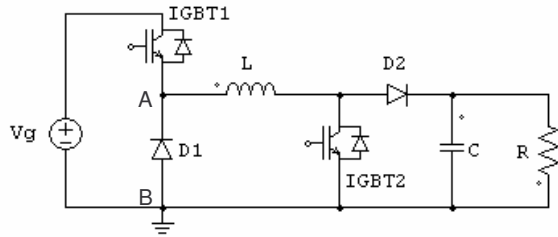


Fig. 1. Non-Inverting Buck-Boost converter.

state variables:

$$L \frac{di_L}{d\tau} = V_{AB} - (1-q)v_C \quad (1)$$

$$C \frac{dv_C}{d\tau} = -\frac{v_C}{R} + (1-q)i_L. \quad (2)$$

The control actions envisage the possibility of forcing $V_{AB} = V_g$ or $V_{AB} = 0$ at will, and $q = 0$, $q = 1$ as well. Introducing the control gains u_1 and $u_2 = 1 - q$, both of them taking values in the discrete set $\{0, 1\}$, system (1),(2) results in

$$L \frac{di_L}{d\tau} = V_g u_1 - v_C u_2 \quad (3)$$

$$C \frac{dv_C}{d\tau} = -\frac{v_C}{R} + i_L u_2. \quad (4)$$

For a systematic analysis it is advisable to minimize the number of parameters of the system. This purpose may be achieved with the change of variables:

$$x_1 = \frac{1}{V_g} \sqrt{\frac{L}{C}} i_L, \quad x_2 = \frac{1}{V_g} v_C, \quad t = \frac{1}{\sqrt{LC}} \tau,$$

and the introduction of

$$\lambda = \frac{1}{R} \sqrt{\frac{L}{C}} > 0,$$

which make the system dimensionless:

$$\dot{x}_1 = u_1 - x_2 u_2 \quad (5)$$

$$\dot{x}_2 = -\lambda x_2 + x_1 u_2. \quad (6)$$

The converter is assumed to undergo possible load variations in the bounded set $[R_{min}, R_{max}]$, this yielding $\lambda \in \Lambda = [\lambda_{min}, \lambda_{max}]$.

A detailed study of the dynamics of the Non-Inverting Buck-Boost converter reveals the impossibility of performing inversion tasks [12]. This situation may be overcome with the replacement of each of the original switches by a full bridge of switches that allows bi-directional current flows. The resulting converter is shown in Figure 2. However, the new system dynamics are still modelled by (5),(6), the change being in the fact that now the control gains u_1, u_2 take values in the set $\{-1, 1\}$.

Assume that the control goal is the tracking of certain T -periodic reference signals $x_{1d}(t), x_{2d}(t)$ by the state

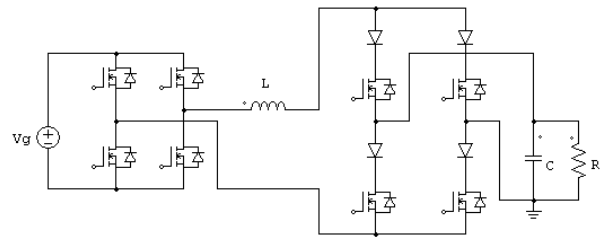


Fig. 2. Full-bridge Non-Inverting Buck-Boost converter.

variables x_1, x_2 , respectively. Hence, in the steady state it must be

$$\dot{x}_{1d} = u_{1N} - x_{2d} u_{2N} \quad (7)$$

$$\dot{x}_{2d} = -\lambda x_{2d} + x_{1d} u_{2N}, \quad (8)$$

where u_{1N}, u_{2N} are the average tracking controls that yield $x_1 = x_{1d}(t), x_2 = x_{2d}(t)$.

Remark 1: Note that the control saturation avoidance condition, i.e. $u_{1N}, u_{2N} \in [-1, 1]$, entails restrictions on admissible target functions: on the one hand it must be $x_{1d}(t) \neq 0, \forall t \geq 0$, and, on the other hand, $\forall t \in [0, T]$ and $\forall \lambda \in \Lambda$,

$$-1 < u_{1N} = \frac{x_{1d} \dot{x}_{1d} + x_{2d} (\dot{x}_{2d} + \lambda x_{2d})}{x_{1d}} < 1 \quad (9)$$

$$-1 < u_{2N} = \frac{\dot{x}_{2d} + \lambda x_{2d}}{x_{1d}} < 1. \quad (10)$$

The fulfillment of (9),(10) guarantees unsaturation of the control action in the steady state.

Finally, using error variables $e_i = x_i - x_{id}(t)$, and $e_{u_i} = u_i - u_{iN}, i = 1, 2$, equations (5),(6) result in

$$\dot{e}_1 = e_{u_1} - x_{2d} e_{u_2} - e_2 u_2 \quad (11)$$

$$\dot{e}_2 = -\lambda e_2 + x_{1d} e_{u_2} + e_1 u_2. \quad (12)$$

III. SLIDING CONTROL OF THE FULL-BRIDGE NON-INVERTING BUCK-BOOST

First of all, consider that the following assumption is fulfilled from now on:

Assumption A.

(i) The state vector reference profiles $x_{1d}(t), x_{2d}(t)$, are C^2 and T -periodic.

(ii) $x_{1d}(t) \neq 0, \forall t \geq 0$.

(iii) The nominal tracking controls u_{1N}, u_{2N} , defined in (9), (10), respectively, lie inside the \mathbb{R}^2 region $[-1, 1] \times [-1, 1], \forall (t, \lambda) \in [0, T] \times \Lambda$.

Let now $\sigma := \sigma(e, t) \in \mathbb{R}^2$, with

$$\begin{aligned} \sigma_1 &= -e_1 \\ \sigma_2 &= x_{2d}(t)e_1 - x_{1d}(t)e_2, \end{aligned} \quad (13)$$

be a switching surface for system (11),(12).

By Assumption A.(ii), $x_{1d}(t) \neq 0$. Hence,

$$\sigma = \begin{pmatrix} -1 & 0 \\ x_{2d} & -x_{1d} \end{pmatrix} \begin{pmatrix} e_1 \\ e_2 \end{pmatrix} = 0 \iff \begin{cases} e_1 = 0 \\ e_2 = 0; \end{cases}$$

additionally, it is straightforwardly verifiable that the equivalent controls coincide with the nominal controls given in (9),(10), i.e.

$$u_{1eq} = u_{1N}, \quad u_{2eq} = u_{2N}.$$

The switching logic that provides the tracking target is given in next Proposition:

Proposition 1: Let Assumption A hold. Let also σ defined in (13) be a switching surface for the system modelled by equations (11),(12). Then, the control law

$$u_1 = \begin{cases} 1 & \text{if } \sigma_1 > 0 \\ -1 & \text{if } \sigma_1 < 0 \end{cases}, \quad u_2 = \begin{cases} 1 & \text{if } \sigma_2 > 0 \\ -1 & \text{if } \sigma_2 < 0 \end{cases},$$

yields sliding modes on $\sigma(e, t) = 0$ and the ideal sliding dynamics results in $x_1 = x_{1d}(t)$, $x_2 = x_{2d}(t)$.

Proof: Let $W(t)$ be the time dependent, real, positive definite, symmetric matrix

$$W(t) = \frac{1}{x_{1d}^2(t)} \begin{pmatrix} x_{1d}^2(t) + x_{2d}^2(t) & x_{2d}(t) \\ x_{2d}(t) & 1 \end{pmatrix}.$$

Then,

$$V(\sigma, t) = \frac{1}{2} \sigma^\top W(t) \sigma$$

is a smooth, positive definite, quadratic function. Furthermore, as the eigenvalues of $W(t)$ are positive real functions of t , it results that (see [15], for example),

$$0 \leq \kappa_{min}(t) \|\sigma\|^2 \leq 2V(\sigma, t) \leq \kappa_{max}(t) \|\sigma\|^2,$$

where $\kappa_{min}(t)$, $\kappa_{max}(t)$ are the smallest and largest eigenvalues of $W(t)$, respectively. The continuity and T -periodicity of such eigenvalues allow us to conclude that they achieve a maximum and a minimum value in $[0, T]$, i.e. there exist real, positive constants ρ_m , ρ_M fulfilling

$$\begin{aligned} 2\rho_m &\leq \min_{t \in [0, T]} \{\kappa_{min}(t)\} \\ 2\rho_M &\geq \max_{t \in [0, T]} \{\kappa_{max}(t)\}. \end{aligned}$$

Therefore, $V(\sigma, t)$ is lower and upper bounded in each sphere $\|\sigma\| = R$ inside a neighborhood of $\sigma = 0$ by positive quantities depending only on R , and these lower and upper bounds $h_R = \rho_m R^2$, $H_R = \rho_M R^2$, respectively, are such that

$$\lim_{R \rightarrow 0} H_R = 0, \quad \lim_{R \rightarrow \infty} h_R = \infty.$$

In order to evaluate the derivative of $V(\sigma, t)$ along the trajectories of (11),(12), note that

$$\dot{V}(\sigma, t) = \frac{1}{2} e^\top \dot{e}.$$

Thus,

$$\begin{aligned} \dot{V} &= e_1 [e_{u_1} - x_{2d} e_{u_2} - e_2 u_2] + \\ &+ e_2 [-\lambda e_2 + x_{1d} e_{u_2} + e_1 u_2] = \\ &= -\lambda e_2^2 + e_1 e_{u_1} + (x_{1d} e_2 - x_{2d} e_1) e_{u_2} \leq \\ &\leq -\sigma_1 (u_1 - u_{1eq}) - \sigma_2 (u_2 - u_{2eq}). \end{aligned}$$

Note that, by Assumption A, u_{1eq} and u_{2eq} are continuous, T -periodic and both of them lay inside $(-1, 1)$; therefore, they reach maximum and minimum values u_{ieq}^+ and u_{ieq}^- , $i = 1, 2$, therein, respectively. Let ϵ_1, ϵ_2 be

$$\epsilon_i = \inf \{ |1 - u_{ieq}^+|, | -1 - u_{ieq}^- | \}, \quad i = 1, 2.$$

Then, the proposed switching logic yields

$$\dot{V}(\sigma, t) \leq -\epsilon_1 |\sigma_1| - \epsilon_2 |\sigma_2| \leq -\alpha (|\sigma_1| + |\sigma_2|),$$

with $\alpha = \inf \{ \epsilon_1, \epsilon_2 \}$. By norm equivalence in \mathbb{R}^n , there exists $\hat{\alpha} > 0$ such that

$$\dot{V}(\sigma, t) \leq -\hat{\alpha} \sqrt{\sigma_1^2 + \sigma_2^2} = -\hat{\alpha} \|\sigma\|.$$

Then, a stable sliding mode along the intersection of the discontinuity surfaces $\{\sigma_1 = 0\} \cap \{\sigma_2 = 0\}$ occurs [16]. ■

Remark 2: Notice from (11), (12) that, if $x_{1d} \neq 0$, every perturbation of the system satisfies the matching condition [17]. Hence, the induced sliding regimes satisfy a so-called strong invariance property [17], which results in the ideal sliding dynamics being independent of the perturbation signal, thus guaranteeing robustness. However, the disturbance may affect the fulfillment of (9),(10): therefore, Assumption A.(iii) has to be preserved in order to prevent the loss of sliding motion on $\sigma(e, t) = 0$.

IV. POWER LOSS MINIMIZATION

Due to both technical and economical reasons, it is extremely convenient to reduce as much as possible power losses in the converters. In this sense, in [12] the authors have developed a procedure to reduce power losses in a class of switched converters through the minimization of the RMS current reference profile. The technique is applied below to the Full-Bridge Non-Inverting Buck-Boost converter.

The power dissipated in a resistive circuit element is directly proportional to the square of the Root Mean Square (RMS) of the current that flows through the element. Moreover, as the target is the tracking of T -periodic output voltage references $x_{2d}(t)$, it is reasonable to search for inductor current reference profiles $x_{1d}(t)$ which are also T -periodic. Hence, assume a truncated Fourier series development for x_{1d} :

$$x_{1d}(t) = a_0 + \sum_{k=1}^r a_k \cos k\omega t + b_k \sin k\omega t, \quad (14)$$

with $\omega = 2\pi T^{-1}$. We may minimize its RMS

$$\begin{aligned} F(a_0, a_1, \dots, b_r) &= \sqrt{\frac{1}{T} \int_0^T x_{1d}^2(t) dt} = \\ &= \sqrt{a_0^2 + \sum_{k=1}^r \frac{a_k^2 + b_k^2}{2}}, \quad (15) \end{aligned}$$

subjected to the restrictions imposed by (9) and (10). Then, the problem to be solved consists of choosing $a_0, a_1, b_1, \dots, a_r, b_r$ so that function F defined in (15) reaches a minimum on the domain defined by inequalities (9) and (10).

The minimization procedure is based on the following result:

Theorem 1: Let the scalar functions $\phi_i : \mathbb{R}^p \times \mathbb{R}^q \rightarrow \mathbb{R}$, $i = 1, \dots, m$, be continuously differentiable, and let the set

$$Z^P = \{z \in \mathbb{R}^p, \phi_i(z, y) \leq 0, i = 1, \dots, m\} \quad (16)$$

be nonempty. Then, there exists $z \in \mathbb{R}^p$ that minimizes the real valued function

$$F(z) = \|z\|$$

on the domain defined by Z^P .

Proof: See [12]. ■

In order to simplify notation, let us denote

$$x_{1d}(z, t) = z_1 + \sqrt{2} \sum_{k=1}^r z_{2k} \cos k\omega t + z_{2k+1} \sin k\omega t, \quad (17)$$

with $z \in \mathbb{R}^{2r+1}$, $t \in \mathbb{R}$, which has an RMS given by

$$F(z) = \|z\|.$$

Moreover, set $y = (t, \lambda) \in \mathbb{R}^2$ and let us define

$$\begin{aligned} f(y) &= f(t, \lambda) = \dot{x}_{2d}(t) + \lambda x_{2d}(t) \\ g(y) &= g(t, \lambda) = x_{2d}(t) [\dot{x}_{2d}(t) + \lambda x_{2d}(t)]. \end{aligned} \quad (18)$$

Straightforward calculation allows one to prove the following result:

Proposition 2: Let $Y = [0, T] \times \Lambda$ be a compact subset of \mathbb{R}^2 , and consider the scalar functions

$$\begin{aligned} \phi_1(z, y) &= \phi_1(z, t, \lambda) = g(t, \lambda) + \\ &+ x_{1d}(z, t) \left[\frac{\partial x_{1d}(z, t)}{\partial t} - 1 \right] \end{aligned} \quad (19)$$

$$\begin{aligned} \phi_2(z, y) &= \phi_2(z, t, \lambda) = -g(t, \lambda) + \\ &- x_{1d}(z, t) \left[\frac{\partial x_{1d}(z, t)}{\partial t} + 1 \right] \end{aligned} \quad (20)$$

$$\phi_3(z, y) = \phi_3(z, t, \lambda) = -[x_{1d}(z, t) - f(t, \lambda)] \quad (21)$$

$$\phi_4(z, y) = \phi_4(z, t, \lambda) = -[x_{1d}(z, t) + f(t, \lambda)], \quad (22)$$

where x_{1d} , $f(t, \lambda)$ and $g(t, \lambda)$ are defined in (17) and (18), respectively. If $\phi_i(z, y) < 0$, $\forall y \in Y$, $i = 1, 2, 3, 4$, then Assumption A.(iii) holds.

Proposition 3: Let Assumption A hold, and let also $Y = [0, T] \times \Lambda$ be a compact subset of \mathbb{R}^2 . Then,

$$F(z) = \|z\|$$

reaches a minimum on the domain

$$Z^P = \{z \in \mathbb{R}^{2r+1}, \phi_i(z, y) \leq 0, \forall y \in Y, \forall i = 1, 2, 3, 4\}.$$

Proof: The continuous differentiability of $\phi_i(z, y)$, $i = 1, 2, 3, 4$, follows immediately from its own definition and from Assumption A. Furthermore, Y being compact, the continuous, real valued functions $|f(y)|$ and $|g(y)|$ reach maximum values in Y , i.e. there exist $f_M, g_M \geq 0$ such that:

$$f_M = \max_{y \in Y} \{|f(y)|\}, \quad g_M = \max_{y \in Y} \{|g(y)|\}.$$

Setting $z_2 = \dots = z_{2r+1} = 0$ in (19),(20),(21),(22) one gets:

$$\begin{aligned} \phi_1(z_1, 0, \dots, 0; y) &= g(y) - z_1, \\ \phi_2(z_1, 0, \dots, 0; y) &= -g(y) - z_1, \\ \phi_3(z_1, 0, \dots, 0; y) &= f(y) - z_1, \\ \phi_4(z_1, 0, \dots, 0; y) &= -f(y) - z_1. \end{aligned}$$

Then, the \mathbb{R}^{2r+1} subset

$$\hat{Z}_1^P = \{(z_1, 0, \dots, 0) \in \mathbb{R}^{2r+1}; z_1 \geq \max\{f_M, g_M\}\}$$

is, trivially, a non-empty subset of Z^P . Finally, the result follows from direct application of Theorem 1. ■

V. SIMULATION RESULTS

The power electronics software PSIM is used to carry out the simulations. The parameters of the Full-Bridge Non-Inverting Buck-Boost converter are: a DC voltage source of $V_g = 50V$, a nominal output resistance of $R = 5\Omega$, an inductance of $L = 1mH$ with an internal resistance of 0.01Ω and a capacitor of $C = 60\mu F$ with an internal resistance of 0.01Ω . Each switch is implemented by means of an IGBT with a saturation voltage of $2V$ and a power diode with a voltage drop of $0.5V$.

The sliding mode controllers require the use of multipliers, sums and comparators which can be implemented by means of analog techniques. A detailed description of the procedure may be found in [18]. In order to provide variable switching frequencies in the standard range, the sliding mode controllers use hysteresis cycle comparators [19], [20], with hysteresis cycles of $0.02V$ for σ_1 and $0.4V$ for σ_2 , as well as zero order holders. Hence, the maximum switching frequency is limited to 120 KHz. Expressions of σ_1 and σ_2 may be found in (13).

The output voltage reference for tracking is

$$v_{Cd}(\tau) = 100 \sin 2\pi\nu\tau,$$

with $\nu = 50Hz$. The values of the corresponding normalized variables are:

$$x_{2d}(t) = 2 \sin \omega t,$$

with $\omega = 0.0770$, corresponding to a normalized period of $T = 81.65$. The converter is assumed to undergo load variations up to 100% of the nominal value, which results in λ varying in the set $[\lambda_{min}, \lambda_{max}] = [0.4082, 0.8165]$. The input current profile is chosen to be

$$x_{1d}(t) = a_0 + a_1 \cos \omega t + b_1 \sin \omega t + a_2 \cos 2\omega t + b_1 \sin 2\omega t.$$

The optimization problem is solved by means of the function `fseminf`, available in the Optimization Toolbox of MATLAB. Essentially, the routine uses cubic and quadratic interpolation techniques to estimate peak values in the semi-infinite constraints. The peak values are used to form a set of constraints that are supplied to a sequential quadratic programming (SQP) method.

Table I contains the optimum values for the reference parameters provided by MATLAB, as well as the corresponding RMS values, in normalized variables, for constant and periodic references. The latter is defined by Fourier series

truncated at the second harmonic. Note that the use of a periodic reference for the inductor current yields a RMS reduction of 34.60% with respect to a constant reference. In terms of power consumption, this amount grows to 57.23%.

TABLE I
OPTIMIZED PARAMETERS FOR THE INDUCTOR CURRENT REFERENCE
AND ASSOCIATED RMS IN NORMALIZED VARIABLES

$x_{1d}(t)$	a_0	a_1	b_1	a_2	b_2	RMS
Constant	3.2731	0	0	0	0	3.2731
Periodic	1.9416	0	0	-1.1725	0.5	2.1406

When carrying out realistic simulations of the converter performance, the DC terms of the inductor current references depicted in Table I have to be slightly increased for sliding motion to be induced. This is a foreseeable effect of unmodelled dynamics and parasitic resistances. Table II provides the ideal and actual values of the DC components of the inductor current references, as well as the RSM values, in Amperes. The percentage reduction of RSM and power loss are similar to the above reported for the ideal case. Furthermore, for the particular case we are dealing the coefficients a_1 and b_1 of the first harmonic of the current reference profile are very low. Due to this fact, the simulations have been carried out setting $a_1 = b_1 = 0$.

TABLE II
IDEAL AND REAL VALUES FOR THE DC TERM OF THE INDUCTOR
CURRENT REFERENCE AND ASSOCIATED RMS

$i_{Ld}(\tau)$	i_{Ld0} (id.)	i_{Ld0} (real)	RMS (id.)	RMS (real)
Constant	40.09	64.00	40.09	64.00
Periodic	23.78	44.00	26.22	45.30

Figure 3 depicts the output voltage tracking the command profile for a constant inductor current reference, which is also captured. Furthermore, the plot includes the load current: its jumps indicate the presence of a resistive load disturbance. Figure 4 is analogous to Figure 3, except in the fact that now a variable inductor current reference is being used. The THD values for constant and variable inductor current references are both 0.02.

Typical applications of DC/AC converters deal not only with resistive loads but also with nonlinear loads. Figure 5 contains a simulation of output voltage tracking with a variable inductor current reference in the presence of a nonlinear load consisting of a Full-Wave Rectifier with a capacitor of $8mF$ and a resistive load of 24Ω . The THD is 0.025.

VI. CONCLUSIONS AND SUGGESTIONS FOR FURTHER RESEARCH

This article presents an inverter obtained from the replacement of the original switches by a full-bridge in a Non-Inverting Buck-Boost converter. Operating under sliding mode control, the converter performance involves robust step-down as well as step-up tasks in presence of resistive loads under the assumption of known bounds for the load

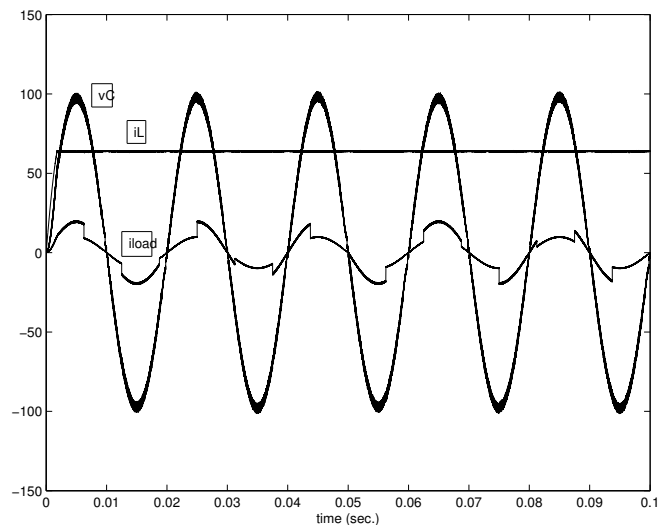


Fig. 3. Output voltage tracking with constant inductor current reference

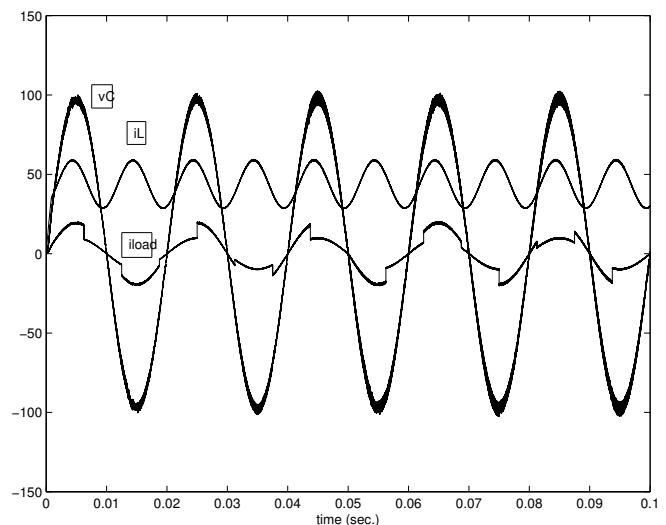


Fig. 4. Output voltage tracking with variable inductor current reference

variation. Using semi-infinite programming techniques, the input current reference is adequately chosen in order to minimize power losses and, at the same time, prevent control saturation. Realistic simulations carried out with PSIM validate the proposal.

Further research may address the performance of the inverter in front of nonlinear loads, for which promising preliminary simulation results have been obtained.

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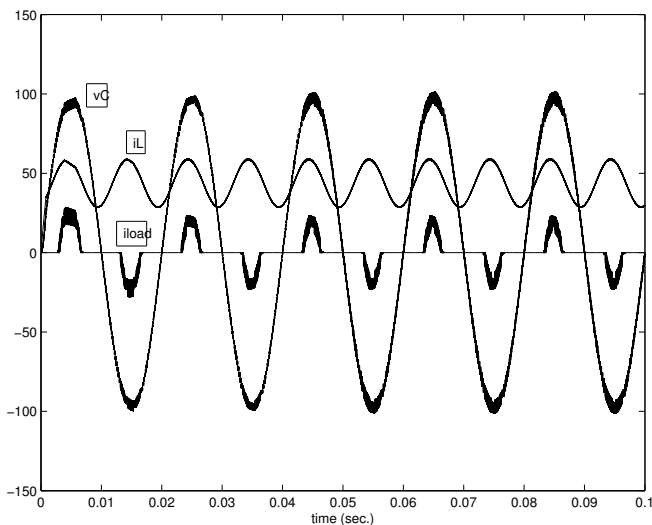


Fig. 5. Output voltage tracking with variable inductor current reference and nonlinear load

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