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Research Article

Energy Considerations for Tracking in DC to DC Power Converters

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A design of an adaptive controller applied to the boost and buck-boost converters to deal with the problem of tracking a sinusoidal voltage is proposed. The main contribution is to provide conditions on the design procedure in order to obtain a reduction in the DC voltage (offset of the sinusoidal signal) of the reference signal; in this way the AC energy is maximized. A nonlinear stable system is designed in order to produce the necessary inputs to exactly track the inductor reference current, which is a necessary condition to achieve the tracking behavior of the voltage reference signal $f(t) = A + B\sin(wt)$. A numerical example is provided to corroborate the result.

1. Introduction

With the advances in the power semiconductor technology, the use and control of switched converters and systems are exponentially increasing due to their potential applications [1–4]. The switched converters have been widely analysed; these converters can increase or decrease the magnitude of the DC voltage or invert its polarity in the load. The boost and buck-boost converters are highly applied in power systems, such as power factor correction [5, 6], in wind power turbines [7, 8]. In general for renewable energy systems [9], moreover, a double loop control of buck-boost converter has been reported, where the control loop is designed for wide range of resistance and reference voltage; however, the output reference voltage is constant [10].

In recent years some works which deal with the tracking problem in power converters have been reported; for instance, in [11, 12] the tracking problem of a sinusoidal reference signal in DC-DC converters is studied and the reported algorithm is locally stable and is based on sliding mode control and the Galerkin method, which can be viewed as a generalization of the harmonic balance method. The sliding mode control has been used to compensate disturbances on the output voltage in DC-DC converters [13]. In [14] an adaptive control topology is used in order to obtain sinusoidal tracking for the inductor current; such procedure design does not guarantee a small DC energy. Authors in [15] present an algebraic approach based on online identification of uncertain parameters which are used to solve the tracking problem in an arrangement of power converters. However, the DC voltage level is not considered as a parameter design. In [16] a sliding mode scheme was reported to perform a DC to AC conversion, and the parameters of the output voltage are chosen such that certain conditions are satisfied. There is a common issue in the previous reported results; that is, the DC voltage level is a parameter of the output voltage signal. Nevertheless, this voltage represents energy consumption since on this voltage the sinusoidal signal is mounted; therefore it is desirable to reduce this value in order to obtain the same sinusoidal signal but with a reduced DC voltage level and thus reduce the energy consumption.

Taking into account these published results we improve the procedure reported in [17] in order to reduce the DC level in the reference signal, which represents a reduction in the consumption of DC energy. A basic problem in the controller design consists in finding the zero dynamics of the system;

such dynamics is required to be stable in order for the system output to track the given reference. In order to overcome this difficulty, the indirect control method is proposed, where an alternative system output is considered and for such a system the internal dynamics is stable. However, other problems arise, for instance, the determination of the reference for the new system output and the control synthesis, which are commonly studied within the field of power converters. One of them is the current mode regulation which is justified in [11], which provides the manner to find an approach to the unstable periodic solution required for the tracking problem. This solution is obtained using the harmonic balance method [12]. On the other hand, a boost topology is used in [11] in order to obtain a system that provides an output voltage greater than the input. Nevertheless, it has been shown that to control the boost and buck-boost converters, the indirect control of the output voltage is required [17].

In this note we propose an adaptive method applied to a boost and buck-boost topologies to achieve tracking of a sinusoidal voltage, where the reference signal is given by $f(t) = A + B\sin(wt)$. The procedure consists in designing an adaptive controller. Besides, it is required to propose a stable nonlinear system in order to obtain the control inputs required to practically track the reference signal. The main contribution consists in finding the conditions under which the controller is implementable and with the capability of reducing the necessary DC energy imposed by the conditions given in [11, 14]. In this sense, the aim of the proposal is to reduce the use of the DC component in the tracking signals as established in [11, 14]; this represents a reduction in the energy required to generate or to track a sinusoidal signal; besides, the error in the tracking signals is practically zero and the DC component is almost the minimum required, which is a great advantage in the tracking of sinusoidal signals using DC-DC converters.

The paper is organized as follows: Section 2 presents the model of the converters and some considerations on electrical conditions. In Section 3 an analysis is performed in order to obtain conditions under which the design loses less DC energy than that given in [6, 9]. Section 4 presents the development of the adaptive scheme to track practically the voltage reference. In Section 5 the proposed result is corroborated via numerical simulations. Finally, the contribution is closed with some conclusions on the tracking voltage problem in DC-DC power converters.

2. Model Description

The mathematical equations which describe the boost and buck-boost converters are given by

$$\begin{aligned} L\frac{dI_L}{d\tau} &= (S_w - 1)(V_C + kV_{CC}) + V_{CC} \\ C\frac{dV_C}{d\tau} &= (1 - S_w) I_L - \frac{V_C}{R} \end{aligned} \tag{1}$$

where if k=0, we have the boost converter (Figure 1(a)); otherwise if k=1, we have the buck-boost converter (Figure 1(b)); V_C, I_L, and V_{CC} are the capacitor voltage, inductor current, and input voltage, respectively. The switching signal S_w is the control signal, which takes values from the set $S = \{0, 1\}$; τ is the nonscaled time. The next change of variable is proposed in order to simplify the analysis [12].

$$\begin{aligned} x &= \frac{I_L}{V_{CC}}\sqrt{\frac{L}{C}}; \\ y &= \frac{V_C}{V_{CC}}; \\ t &= \frac{\tau}{\sqrt[2]{LC}} \\ a &= \frac{1}{R}\sqrt{\frac{L}{C}}, \\ \overline{u} &= 1 - S_w \end{aligned} \tag{2}$$

Then the dynamical model takes the form bellows.

$$\begin{aligned} \dot{x} &= 1 - (k + y)\overline{u} \\ \dot{y} &= -ay + x\overline{u} \end{aligned} \tag{3}$$

Therefore, the problem consists in finding a control signal input such that the solution $y(t) \longrightarrow f(t) > 0$, where the reference $f(t)$ is a positive signal.

Multiplying (3) by x and $(k + y)$, the following differential equation is obtained.

$$(k + y)(\dot{y} + ay) = x(1 - \dot{x}) \tag{4}$$

When $y(t)$ reaches the steady state $y(t) = f(t)$ given by a periodic orbit, the differential equation $(k + f)(\dot{f} + af) = x(1 - \dot{x})$ has a unique solution given by an unstable limit cycle $ø(t)$ when $g(t) = (k + f)(\dot{f} + af) > 0$ [11]. Thus, to obtain convergence of the solution $x \longrightarrow ø_1(t)$, an approximation of $ø_1(t) > 0$ is proposed and is considered as the 1st harmonic of the Fourier series of $ø(t)$. From this, it is clear that these converters can track a sinusoidal signal.

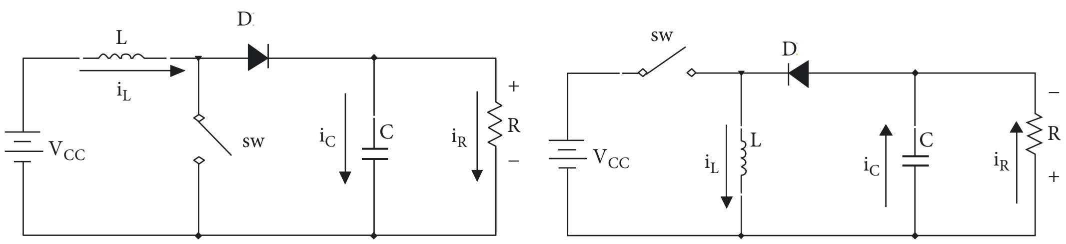
As it was mentioned previously, the boost and buck-boost converters require indirect control of the output voltage [11] in order to avoid instabilities in the internal dynamics. Considering the signal $ø_1(t)$ as the reference for x(t), the main tracking problem is defined, to control the system in order to obtain a zero error tracking between the reference and the output using a smooth control. From (4) and considering $x(t) = ø_1(t)$, we obtain

$$\dot{y} = -ay + \frac{ø_1(1 - \dot{ø}_1)}{k + y} \tag{5}$$

and now with the change $z_y = (k + y)^{-1}$ it can be seen that system (5) has a stable limit cycle $\gamma(t)$, as the dynamics of $z_y(t)$ is

$$\dot{z}_y = az_y - akz_y{}^2 - z_y{}^3 ø_1(1 - \dot{ø}_1) \tag{6}$$

where we can see that if $ø_1(t) > 0$ and $(1 - \dot{ø}_1) > 0$, then for any initial condition $z_y(0) > 0$, the solution z_y of system



(a) (b)

Figure 1: (a) Boost converter circuit. (b) Buck-boost converter circuit.

(6) tends to the limit cycle given by $\eta(t) = (k + \gamma(t))^{-1}$. Now, considering the steady state continuous control input $u_0(t)$ and $u(t)$ as the continuous control input calculated as the mean value of $\overline{u}$, $u_0(t) = (1 - \dot{\varnothing}_1)\eta(t)$. Thus, taking (3) and implementing (6) we can write

$$\begin{aligned} \dot{x} &= 1 - (k + y)\, u \\ \dot{y} &= -ay + xu \\ \dot{z} &= az - akz^2 - z^3 \varnothing_1 (1 - \dot{\varnothing}_1) \\ u &= (1 - \dot{\varnothing}_1)\, z \end{aligned} \tag{7}$$

and we have that $u(t) \longrightarrow u_0(t); x(t) \longrightarrow \varnothing_1(t); y(t) \longrightarrow \gamma(t)$ and the switched input $\overline{u}$ is replaced by its equivalent in continuous time $u(t)$. In order to show this assertion, consider the error signals $e_y = y - \gamma$, $e_x = x - \varnothing_1$ and note that $z \longrightarrow (k + \gamma)^{-1}$, since by design we consider $\varnothing_1 > 0$ and $(1 - \dot{\varnothing}_1) > 0$; then we write the dynamics for the error system on the steady state of z as

$$\begin{aligned} \dot{e}_y &= -ae_y + e_x \frac{(1 - \dot{\varnothing}_1)}{k + \gamma} \\ \dot{e}_x &= -\frac{(1 - \dot{\varnothing}_1)}{k + \gamma} e_y \end{aligned} \tag{8}$$

where $\dot{\gamma} = -a\gamma + \varnothing_1(1 - \dot{\varnothing}_1)/(k + \gamma)$. Taking the Lyapunov function $V(e) = e_x^2/2 + e_y^2/2$ and the derivative given as $\dot{V}(e) = -ae_y^2$, then, by LaSalle theorem, $e_y \longrightarrow 0$, and from the first equation $e_x \longrightarrow 0$. Solving (6) the input for $x \longrightarrow \varnothing_1$ is found. Nevertheless, the parameter a is uncertain but it can be estimated by means of an adaptive scheme. Now, if the input function $\varnothing_1$ is an affine function of the parameter a; thus it is possible to determine a globally stable controller.

Remark 1. The reference for the current $\varnothing_1$ could be any function of time and if $\varnothing_1 > 0$ and $(1 - \dot{\varnothing}_1) > 0$, then implementing (6) we have that $x \longrightarrow \varnothing_1$.

Let the reference voltage $y(t)$ given as $f(t) = A + B\sin(wt)$, and as it was described in [12], $\gamma(t)$ is the first harmonic of $f(t)$ if $\varnothing_1$ is properly chosen. Using the harmonic balance method, the function $\varnothing_1$ must satisfy (see [12])

$$\varnothing_1 = aA_0 + B_1 \sin(wt) + C_1 \cos(wt) \tag{9}$$

$$\begin{aligned} B_1 &= M \sin(\theta) = \frac{aB\left[(2A + k) - w^2 A_0 (k + A)\right]}{a^2 A_0^2 w^2 + 1} \\ C_1 &= M \cos(\theta) = \frac{wB\left[a^2 A_0 (2A + k) + (k + A)\right]}{a^2 A_0^2 w^2 + 1} \\ A_0 &= A(A + k) + \frac{B^2}{2} \end{aligned} \tag{10}$$

and these give the parameters to find $\varnothing_1$ which depends on the values of the constants A and B. Now consider that if M and θ are independent of a, then $\varnothing_1$ is an affine function of a.

In the following, we give some conditions for the amplitude of $\varnothing_1$ to find an implementable controller, such that it satisfies $0 \leq u_0 \leq 1$, and $\varnothing_1$ is an affine function of the parameter a.

In order to show that it is possible to choose $\varnothing_1$ as an affine function of a, consider the solution of $\varnothing_1$ in (9), and consider a fixed frequency given as $w = 2\pi f_r \sqrt[2]{LC}$ where f_r is the real frequency; therefore, the frequency w is determined by the circuit parameters L and C. It is clear that if the reference signal is modified, then parameters designs L and C must be readjusted.

From B_1 and C_1 in (9), we have that the value

$$w^2 = \frac{2A + k}{A_0 (k + A)} \tag{11}$$

leads to $B_1 = 0$, and $C_1 = Bw(k + A) = B\sqrt{2}\sqrt{(2A + k)(k + A)/(2A(A + k) + B^2)}$ which are independent of a. Then $\varnothing_1$ is affine to a. Thus, $M = C_1$, $\theta = 0$, and

$$\begin{aligned} \varnothing_1(t) &= aA_0 + M\cos(wt) \\ 1 - \dot{\varnothing}_1(t) &= 1 + Mw\sin(wt) \\ A_0 &= A(A + k) + \frac{B^2}{2}, \\ M &= Bw(k + A). \end{aligned} \tag{12}$$

Now we study the conditions on the parameters A and B in order to obtain $\varnothing_1(t) > 0$ and $(1 - \dot{\varnothing}_1(t)) > 0$. Defining

$D = B(2A+k)/A_0$ and considering the value of frequency (11), then

$$(1-\dot{\o}_1(t)) > 0 \Longleftrightarrow 1 > Mw \Longleftrightarrow 1 > D. \tag{13}$$

Solving the last equation we obtain the minimum value that must be taken by A to satisfy $(1-\dot{\o}_1(t)) > 0$, for a given value of B. We arrive to

$$A > A_m = \frac{-k}{2} + B + \frac{1}{2}\sqrt{k^2+2B^2} \tag{14}$$

and then if $A > A_m$, we have $(1-\dot{\o}_1(t)) > 0$. On the other hand, defining $F = \sqrt{BD(A+k)/A_0}$, we obtain that

$$\o_1(t) > 0 \Longleftrightarrow aA_0 > M \Longleftrightarrow a > F \tag{15}$$

and then taking $A = A_m + \delta$ for some $\delta > 0$, then $A_0(\delta)$, $F(\delta)$, and $D(\delta)$ are functions of δ.

Observing that, by design, it is possible to consider that $a > a_{min}$ for some known constant a_{min}, which corresponds to some given $R_{max}(a_{min} = (1/R_{max})\sqrt{L/C})$, and observing that a_{min} also can be considered as a design parameter if we take

$$C = \frac{w}{2\pi f_r R_{max} a_{min}}, \qquad L = \frac{w R_{max} a_{min}}{2\pi f_r} \tag{16}$$

then we have the following.

$$\o_1(t) > 0 \Longleftrightarrow a_{min} > F(\delta), \quad \delta > 0 \tag{17}$$

Considering $D(\delta)$ we have that $D(0) = 1$ and $\lim_{\delta\longrightarrow\infty} D(\delta) = 0$, and

$$\frac{dD(\delta)}{d\delta} = \frac{-2B}{A_0(\delta)^2}\left[\left(A_m+\delta+\frac{k}{2}\right)^2 + \frac{k^2}{4} - \frac{B^2}{2}\right] < 0. \tag{18}$$

By definition, $A_m > B$, and then $D(\delta)$ is a decreasing function of δ. For $F(\delta)$ we have the following.

$$F(0) = \frac{\sqrt{B(A_m+k)}}{A_m(A_m+k)+B^2/2} = \frac{\sqrt{B(A_m+k)}}{B(2A_m+k)}, \qquad \lim_{\delta\longrightarrow\infty} F(\delta) = 0 \tag{19}$$

where (19) is a decreasing function of δ as well, because $\sqrt{BD(\delta)}$ and $\sqrt{(A(\delta)+k)/A_0(\delta)}$ are also decreasing, finally taking $a_{min} = F(0)$, then (13) and (15) hold for any $\delta > 0$.

3. Main Result on Energy Considerations

In this section some conditions are given in order to obtain an implementable controller; that is, $0 \le u \le 1$ considering that the DC energy must be low. The condition for implementable controller in [9] is as follows.

$$\inf\left\{(k+f)\left(\dot{f}+af\right)\right\} \ge \sup\left\{\left(\dot{f}+af\right)\right\} \tag{20}$$

And a less restrictive condition given in [11] is as follows.

$$\min\{\o_1\}\min\{(1-\dot{\o}_1)\} \ge (a\max\{1-\dot{\o}_1\})(\max\{1-\dot{\o}_1\}-k) \tag{21}$$

The problem, in both cases, is that the maximum of one signal is compared with the minimum of other signal, which gives a restrictive conditions on the DC component of the reference signal $f(t)$, losing more DC energy than necessary. Considering this problem, taking into account that in steady state the input is given by

$$u_0(t) = \frac{1-\dot{\o}_1}{k+\gamma}, \tag{22}$$

and also considering that to obtain physical implementation we need that $u_0(t) > 0$ and $u_0(t) \le 1, \forall t$, which implies that $(1-\dot{\o}_1) \le (k+\gamma)$, then, the first harmonic of γ(t) is given as $f(t) = A + B\sin(wt)$ and $(1-\dot{\o}_1) = 1 + Mw\sin(wt)$ has the same phase; then it is possible to relax the above conditions ((20) and (21)).

Now in order to consider the conditions for the control input, that is, $0 \le u(t) \le 1$, we will study their dynamics, which can be taken from $u = (1-\dot{\o}_1)z$ and written as follows.

$$\begin{aligned}\dot{u} &= -\ddot{\o}_1 z + (1-\dot{\o}_1)\dot{z}\\ \dot{u} &= au - \frac{\ddot{\o}_1}{(1-\dot{\o}_1)}u - \frac{ak}{(1-\dot{\o}_1)}u^2 - \frac{\o_1}{(1-\dot{\o}_1)}u^3\end{aligned} \tag{23}$$

It follows that if $(1-\dot{\o}_1) > 0, \forall t$, and $u(0) > 0$, then $u(t) > 0, \forall t$, because the solution of the linear system is

$$\dot{u}_L = au_L - \frac{\ddot{\o}_1}{(1-\dot{\o}_1)}u_L \tag{24}$$

and the linear part of (23) is

$$u_L = \left(\frac{1-\dot{\o}_1(t)}{1-\dot{\o}_1(0)}\right)e^{at}u_L(0) \tag{25}$$

and is an unstable solution; that is, the equilibrium point $u = 0$ of (23) is unstable. Also this implies, in steady state, that if $\o_1 > 0$, then $u_0\o_1 = \dot{\gamma} + a\gamma > 0$ (see [12]).

At this point, we make the change of variable $v(t) = 1 - u(t)$, in order to find the conditions under which $u(t) \le 1$.With this change of variable, it is clear that $v(t) < 1$, because $u(t) > 0$. Then if we can show that $v(t) \ge 0$, this implies that $u(t) \le 1$. Writing the dynamics of $v(t)$, we found

$$\dot{v} = (1-v)\left(\Phi_t - \frac{ak}{1-\dot{\phi}_1}v - \frac{\o_1}{(1-\dot{\phi}_1)}v(2-v)\right) \tag{26}$$

with $\Phi_t = \ddot{\o}_1/(1-\dot{\o}_1) - a + ak/(1-\dot{\o}_1) + \o_1/(1-\dot{\o}_1)$.

In the dynamics, it can be seen that if $\Phi_t \geq 0$, then $v \geq 0$; to see this fact we only need to observe that for any $v < 0$ we have that $\dot{v} > 0$ and then there does not exist a minimum for $v < 0$; this implies that the minimum exists only for $v \geq 0$. Now considering the condition $\Phi_t \geq 0$ and $(1 - \dot{\varnothing}_1) > 0$, this is equivalent to

$$\left(1 - \dot{\varnothing}_1\right) \Phi_t = \ddot{\varnothing}_1 - a\left(1 - \dot{\varnothing}_1\right) + ak + \varnothing_1 \geq 0 \quad (27)$$

and, taking the definition for $\varnothing_1$, we can write

$$\begin{aligned} \varnothing_1 &= aA_0 + M\cos(wt) \\ 1 - \dot{\varnothing}_1 &= 1 + wM\sin(wt) \\ \ddot{\varnothing}_1 &= -w^2\cos(wt) = -w^2\left(\varnothing_1 - aA_0\right) \end{aligned} \quad (28)$$

and the condition was transformed on the following.

$$aw^2A_0 + ak - a\left(1 - \dot{\varnothing}_1\right) + \left(1 - w^2\right)\varnothing_1 \geq 0 \quad (29)$$

At this point if $w \leq 1$, then $(1-w^2)\varnothing_1 \geq 0$; to obtain $w \leq 1$, we use (11), and using the fact that w is a decreasing function of δ, we obtain the values of B that must be chosen to obtain $w \leq 1$ for any $\delta \geq 0$. Then for $k = 0$, we must choose $B \geq (\sqrt{1 + 1/\sqrt{2}})^{-1} = 0.7653669$, and for $k = 1$ the value is given for one of the roots of the equation $x^4 + 2x^3 - 4x^2 - 2x + 2 = 0$ and the value of B must satisfy $B \geq 0.5794246$. Now, it is only needed to fulfill the following.

$$\left(1 - \dot{\varnothing}_1\right) - w^2A_0 - k \leq 0 \quad (30)$$

Using $(1 - \dot{\varnothing}_1) \leq 1 + Mw = 1 + D$, it is possible to write

$$1 + Mw - w^2A_0 - k \leq 0 \Longleftrightarrow k + \frac{2A + k}{(k + A)} - 1 \geq D \quad (31)$$

which is satisfied for any $\delta \geq 0$, because

$$D \leq 1 \leq k + \frac{A}{(k + A)} \quad \text{for } k = 0 \text{ or } k = 1 \quad (32)$$

and then, with these values of $B, a_{min}, A = A_m + \delta$, and $\delta > 0$, it is ensured that $0 < u(t) \leq 1$. It is important to note that the parameter A is the bias in the reference signal and has a smaller value than that reported in [12, 14]. In this sense this is the reason for why we state that the energy consumption is reduced.

4. The Adaptive Controller

Now to accomplish the control objective which consists in the state $x \longrightarrow \varnothing_1(t)$ and $y \longrightarrow f(t) = A + B\sin(wt)$ in an approximated way, an adaptive controller is proposed. For a given $B > B_{min}$ and $a_{min} = F(0)$, with $a \geq a_{min}$, we determine the amplitude A and the frequency w for $\delta > 0$. Consider a fixed R_{max}, which corresponds to the a_{min}; then L and C are given by

$$\begin{aligned} C &= \frac{w}{2\pi f_r R_{max} a_{min}}, \\ L &= \frac{w R_{max} a_{min}}{2\pi f_r}, \end{aligned} \quad (33)$$

Table 1: Comparison of the conditions for the value A of the reference signal $f(t)$.

Reference	Condition
[11]	$A \geq 3.69537$
[14]	$A \geq 2.6476$
New condition	$A = 1.466$

as an example, for given $k = 1, B = 1, f_r = 60Hz, \delta = 0.1$, when $R_{max} = 500\Omega$, we obtain $A = 1.466, w = 0.6224613, a_{min} = 0.412156, L = 340.26mH$ and $C = 8.012\mu F$. In Table 1 we compare the conditions on the value of the parameter A of the reference signal $f(t)$, with the same value of $B = 1$.

From the parameters A, B, and w, the following can be calculated: $\varnothing_1 = aA_0 + M\cos(wt)$, $1 - \dot{\varnothing}_1 = 1 + Mw\sin(wt)$. Considering $a = a_{min} + a_p, \varnothing_{1m} = a_{min}A_0 + M\cos(wt), \varnothing_1 = a_pA_0 + \varnothing_{1m}$, (6) is written as

$$\begin{aligned} \dot{z} &= F_{1z}(z,t) + a_pF_{2z}(z,t) \\ F_{1z}(z,t) &= a_{min}z(1 - kz) - z^3\varnothing_{1m}\left(1 - \dot{\varnothing}_1\right) \\ F_{2z}(z,t) &= z(1 - kz) - A_0z^3\left(1 - \dot{\varnothing}_1\right) \end{aligned} \quad (34)$$

where $a_p > 0$ implies that the system is stable. Now consider that the voltage and the current are available for feedback; then the main result is given in the following theorem.

Theorem 2. *Consider $a = a_{min} + a_p$, with $a_p \geq 0, \varnothing_1 = aA_0 + M\cos(w\tau)$, where $A_0 = A(A+k) + B^2/2, M = Bw(k+A), w^2 = (2A + k)/A_0(k + A)$. Assume that the pair A, B satisfy $A = A_m + \delta, B > B_{min} = (0.7653669)(1 - k) + k(0.5794246)$, and $\delta > 0$; if the tracking adaptive controller is given by*

$$\begin{aligned} \dot{\hat{x}} &= 1 - \left(\hat{y} + k\right)u + g_1\left(x - \hat{x}\right) \\ \dot{\hat{y}} &= -a_{min}y - \hat{a}_py + \hat{x}u + g_2\left(y - \hat{y}\right) \\ \dot{\hat{a}}_p &= -g_3y\left(y - \hat{y}\right) \\ \dot{\hat{z}} &= F_{1z}(\hat{z},t) + \left|\hat{a}_p\right|F_{2z}(\hat{z},t) \\ u &= \left(1 - \dot{\varnothing}_1\right)\hat{z} \end{aligned} \quad (35)$$

and $\hat{z}(0) > 0$, then $x \longrightarrow \varnothing_1(t), y \longrightarrow \gamma(t), \hat{a}_p \longrightarrow a_p, \hat{x} \longrightarrow x, \hat{y} \longrightarrow y$, and $\hat{z} \longrightarrow z$, globally.

Proof. Note that $\hat{z}$ is bounded; taking $V_z(\hat{z}) = \hat{z}^2/2$, we get

$$\begin{aligned} \dot{V}_z(\hat{z}) &= \left(a_{min} + \left|\hat{a}_p\right|\right)\hat{z}^2(1 - k\hat{z}) \\ &\quad - \hat{z}^4\left(\left|\hat{a}_p\right|A_0 + \varnothing_{1m}\right)\left(1 - \dot{\varnothing}_1\right) \end{aligned} \quad (36)$$

which is negative for $\hat{z}^2 > (a_{min} + |\hat{a}_p|)(1 - k\hat{z})/(|\hat{a}_p|A_0 + \varnothing_{1m})(1-\dot{\varnothing}_1)$, and a finite value always exists because if $|\hat{a}_p| = 0$, then it is negative for $\hat{z}^2 > a_{min}(1 - k\hat{z})/\varnothing_{1m}(1 - \dot{\varnothing}_1)$, and if $|\hat{a}_p| \longrightarrow \infty$, then it is negative for $\hat{z}^2 > (1 - k\hat{z})/A_0(1 - \dot{\varnothing}_1)$;

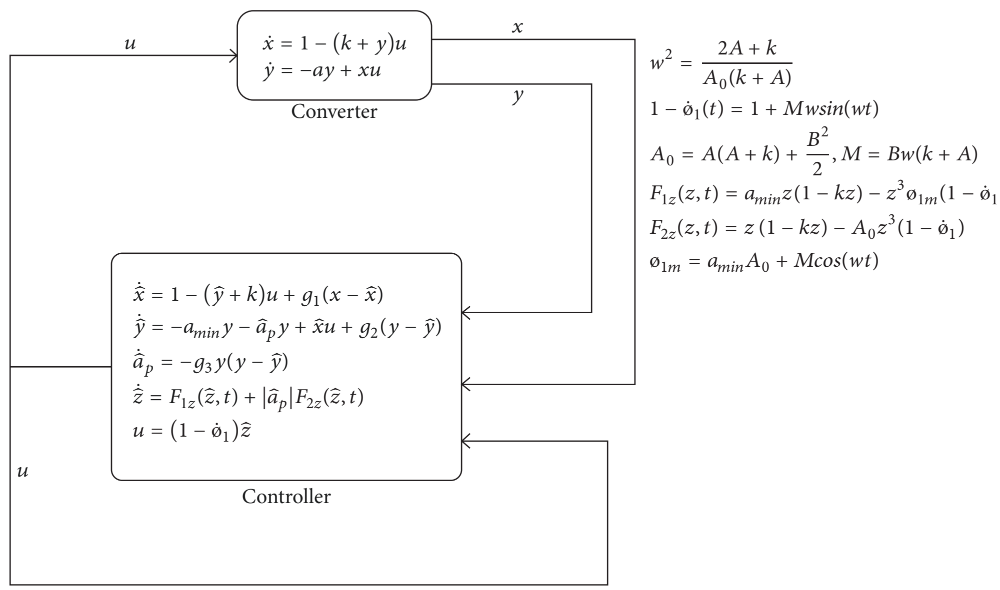


Figure 2: Block diagram of the closed-loop system.

also $\hat{z}$ cannot take the zero value because $(a_{min} + |\hat{a}_p|)\hat{z}^2 > 0$; therefore, $\hat{z}$ is bounded. Finally, u and y are bounded, since for any bounded input the time response of the system also is bounded.

Now consider the dynamical error system from (3) and (35), with $e_y = y - \hat{y}, e_x = x - \hat{x}, e_a = a_p - \hat{a}_p$; then

$$\begin{aligned} \dot{e}_x &= -ue_y - g_1 e_x \\ \dot{e}_y &= -ye_a + e_x u - g_2 e_y \\ \dot{e}_a &= g_3 y e_y \end{aligned} \tag{37}$$

and taking $V(e) = {e_x}^2/2 + {e_y}^2/2 + {e_a}^2/(2g_3)$, the time derivative is given by $\dot{V}(e) = -g_1{e_x}^2 - g_2{e_y}^2$, and by LaSalle theorem [18], $\dot{V}(e)$ is a decreasing function of t and $e_x \longrightarrow 0$, $e_y \longrightarrow 0$; then $ye_a = 0$, which implies that $e_a = 0$, since $y \neq 0$. Now, when $\hat{a}_p = a_p$, the dynamics of $\hat{z}$ produces the necessary input to obtain $x \longrightarrow \varnothing_1(t)$ and $y \longrightarrow \gamma(t)$ whose first harmonic is equal to $f(t) = A + B\sin(wt)$, and since the function $V(e)$ is radially unbounded, the error system is globally stable and the controller globally stabilizes system (3), which completes the *proof*. □

The scheme that describes the closed-loop system is illustrated by means of a block diagram in Figure 2, where the controller and the system plant (converter) are interconnected.

5. Numerical Results

A numerical simulation of the proposed scheme is presented, which illustrates that the converter buck-boost (given with $k = 1$) tracks a sinusoidal reference given by $f(t) = A + B\sin(wt)$ where the departing parameter is the amplitude of the reference signal given by $B = 1$ (known design parameter). Then the values w and A must be calculated. Now consider the following parameter values:

$$\begin{aligned} f(t) &= A + B\sin(wt), \\ \varnothing_{1m}(t) &= A_0 a_{min} + M\cos(wt) \\ 1 - \dot{\varnothing}_1(t) &= 1 + wM\sin(wt), \\ a_{min} &= 0.412156, \\ A &= 1.466, \\ B &= 1, \\ k &= 1, \\ w &= 0.6224613, \\ A_0 &= 4.1152, \\ Mw &= 0.95548, \\ A_0 a_{min} &= 1.6961, \\ M &= 1.535, \\ C &= 8.012\,\mu F, \\ L &= 340.26\,mH, \\ V_{CC} &= 12, \\ f_r &= 60Hz, \\ R_{max} &= 500\,\Omega \end{aligned} \tag{38}$$

and the controller parameters are chosen as $g_1 = g_2 = g_3 = 1$. Solving (35), Figure 3 shows the voltage response and the

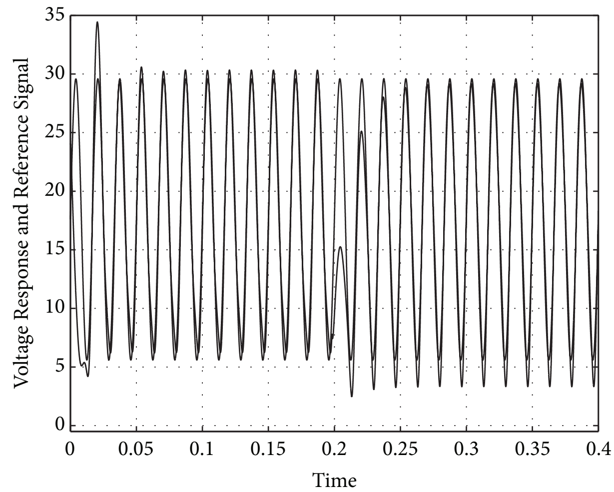


Figure 3: Voltage response and reference voltage signal.

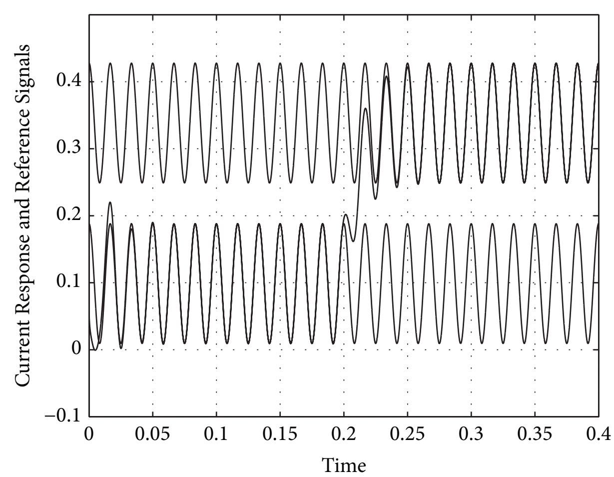


Figure 4: Current response and reference current signal.

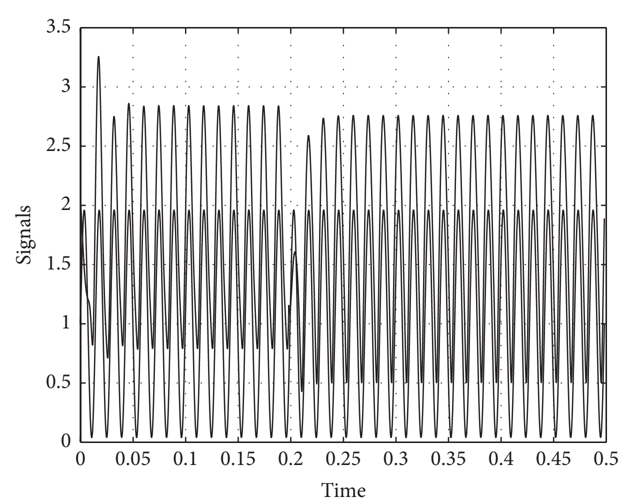
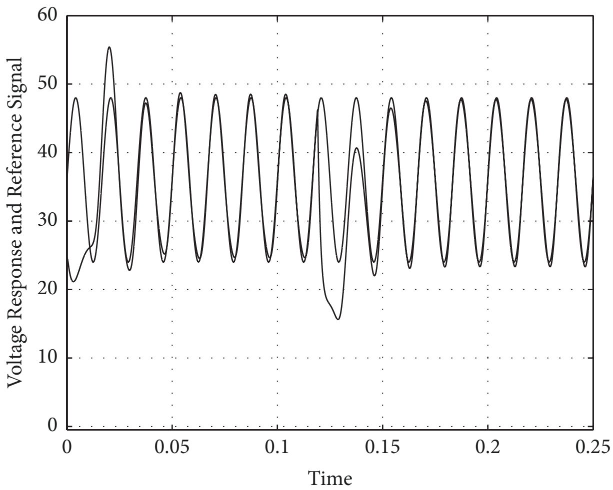
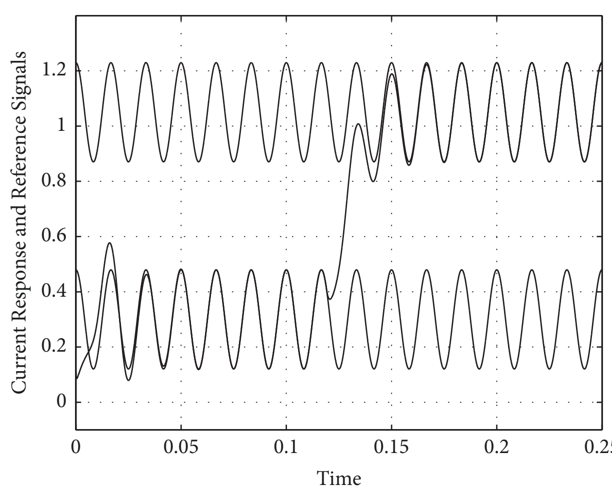


Figure 5: Comparison between $(1 - \dot{\varnothing}_1)$ and $(k + \gamma)$.



(a)



(b)

FIGURE 6: Voltage and current tracking of the sinusoidal signal, as given in [14].

reference signal. A variation in the parameter R is applied as follows: $\tau < 0.1321$ $(t < 80), a_p = 0$ $(R = 500\Omega)$ and for $\tau \geq 0.1321, a_p = 1$ $(R = 145.93\Omega)$, and such a variation in the parameter is compensated by the controller.

Figure 4 shows the response of the current and the reference signals under the same parameter change, once again the current tracks the reference with zero error.

Figure 5 shows the signals $(k+\gamma)$ and $(1-\dot{\varnothing}_1)$ for $k = 0$; it can be seen that it is not necessary that $\max(1-\dot{\varnothing}_1) \leq \min(k+\gamma)$.

In order to indicate the improvement in the design of the system, we present the performance of the design used as a base of the present one [14]. In Figure 6 the voltage and current as given in [14] are presented; note that in the improved design (Figures 3 and 4) the DC component is less than the design presented in Figure 6 as defined in [14]; in this sense the new algorithm requires a reduced value for the DC component, representing a reduction in the consumption of DC energy.

6. Conclusions

In this contribution an adaptive control for tracking of sinusoidal signals in DC-DC converters was described. The scheme reduces the DC energy consumption since a bias is calculated to be small and preserving the tracking of the reference signal. Less restrictive conditions on the current parameters were provided compared to those reported in [12, 14]. Also, an adaptive control scheme to track the voltage output of the power converters was proposed, and it was demonstrated that it can lead the voltage output signal to a biased sinusoidal reference. In this way such a controller is based on a nonlinear model which serves as a mean to obtain the exact necessary input signal to obtain an exact tracking of the current and its reference. Also it was shown that the adaptive controller is robust against changes in the load resistance.

Data Availability

The data used to support the findings of this study are included within the article.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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