CONTROL-THEORY AND ADVANCED TECHNOLOGY
Vol. 9, No. 1, pp. 247-258, March, 1993

C92019

Contributed Paper

A DYNAMIC POLE ASSIGNMENT APPROACH TO STABILIZATION AND TRACKING IN DC-TO-DC FULL BRIDGE POWER CONVERTERS*

H. SIRA-RAMIREZ[1] AND M. T. PRADA-RIZZO[2]

Abstract. In this article, a new approach is proposed for the design of a Pulse Width Modulated (PWM) feedback control scheme, regulating a Full Bridge Buck Converter in both stabilization and AC signal tracking tasks. The approach is based on the specification of a dynamical feedback control law which accomplishes asymptotic stabilization to a pre-selected constant operating point, or, alternatively, indirect asymptotic sinusoidal output voltage tracking for the average PWM converter model. The approach emphasizes the use of Fliess's Generalized Observability Canonical Form of the average converter model and partial inversion techniques.

Key Words—D.C. power supplies, pulse-width-modulation, asymptotic tracking.

1. Introduction

The Full Bridge Buck Converter (FBBC) constitutes a popular means of generating automatically regulated constant output voltages, of either polarity, which represent an arbitrarily specified fraction of the constant input source voltage (see Boudjema et al., 1989). The FBBC may also be used in the generation of controlled sinusoidal voltages through the asymptotic tracking of a suitable AC reference signal referred to as the "template signal." A typical application of this capability lies in the area of low power emergency inversion. For a detailed account of the principles of operation of this versatile class of DC power converters, the reader is referred to the books by Csaki et al. (1983), Mohan et al. (1989) and Severns and Bloom (1985), as well as the current literature appearing in specialized journals.

Sliding mode and PWM control schemes were proposed in Boudjema et al. (1989) for the efficient handling of AC signal tracking problems in FBBC. For the PWM regulator case, the design was accomplished, using static pole assignment techniques through full state feedback defined on the basis of the average PWM converter model.

In this article, we propose the use of a dynamical PWM feedback regulator

* Received by the editors February 21, 1992.
This research was supported by the Consejo de Desarrollo Científico, Humanístico y Tecnológico of the Universidad de Los Andes under Research Grant I-362-91.
[1] Departamento Sistemas de Control, Escuela de Ingeniería de Sistemas, Universidad de Los Andes, Mérida, Venezuela.
[2] Departamento Circuitos y Medidas, Escuela de Ingeniería Eléctrica, Universidad de Los Andes, Mérida, Venezuela.

which accomplishes asymptotic stabilization of the output voltage to a pre-selected constant operating point, or, alternatively, induces indirect asymptotic sinusoidal output voltage tracking for the average PWM converter model. The approach emphasizes the use of a partial (state) inversion strategy in conjunction with Fliess's Generalized Observability Canonical Form (GOCF) (see Fliess, 1990 a) of the average PWM controlled converter.

2. A Dynamical PWM Feedback Controller for Stabilization and Tracking in the Full Bridge Buck Converter

2.1 The Full Bridge Buck Converter Consider the following system of ordinary differential equations describing an FBBC which includes an ideal high frequency isolation transformer (see Fig. 1):

$$\left.\begin{aligned} \dot{x}_1 &= -w_0 x_2 + ub \\ \dot{x}_2 &= w_0 x_1 - w_1 x_2 \end{aligned}\right\}, \tag{2.1}$$

where x_1 is the normalized input inductor current, defined as $x_1 = I_L\sqrt{L}$, x_2 is the normalized output capacitor voltage given by $x_2 = V_0 N\sqrt{C}$, with N being the transformer's winding turn ratio and V_0 is the primary winding voltage drop. The constant w_0 represents the natural oscillating frequency of the LC input circuit, $w_0 = 1/(N\sqrt{LC})$, with the capacitance being referred to the primary of the transformer, and w_1 is the inverse time constant associated to the RC output circuit; i.e., $w_1 = 1/RC$. The constant b is the normalized input source voltage, $b = V_s/\sqrt{L}$, assumed to be constant. The variable u acts as the control input taking values in the discrete set $\{-1, 0, 1\}$, according to the positions of the switches. This feature makes the FBBC truly a generalization of the standard two position switch-controlled Buck converter, also known as the "step-down" converter (see, for instance, Sira-Ramirez, 1987). A suitable operation of the switch arrangement which is capable of realizing such a discrete set of available control inputs is simply summarized in Table 1 (see also Boudjema et al., 1989).

The FBBC constitutes, therefore, a *Variable Structure System* which may be naturally controlled by means of an ON-OFF-ON PWM control policy (see the appendix in Sira-Ramirez, 1991 c) of the following type:

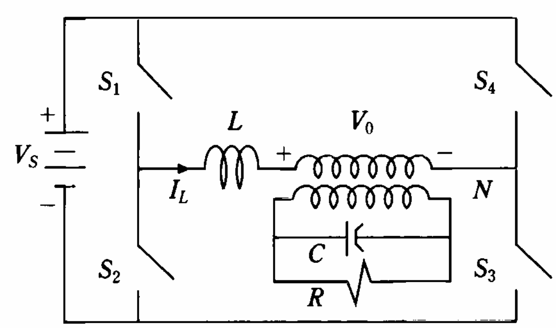


Fig. 1. Full Bridge Buck Converter.

Table 1.

	S_1	S_2	S_3	S_4
$u=1$	ON	OFF	ON	OFF
$u=0$	OFF	ON	ON	OFF
$u=-1$	OFF	ON	OFF	ON

$$u = \begin{cases} \operatorname{sign} \mu[x(t_k)] & \text{for} \quad t_k \le t < t_k + |\mu[x(t_k)]|T, \\ 0 & \text{for} \quad t_k + |\mu[x(t_k)]|T \le t < t_k + T, \end{cases} \tag{2.2}$$

where $\mu : R^2 \to R$ is the *duty ratio function* taking values in the closed interval $[-1, \ 1]$ of the real line, $x(t_k)$ is the sampled value of the state at time t_k and T represents the constant sampling period. The function "sign" stands for the "signum" function. The absolute value of the duty ratio function μ specifies, at each sampling instant t_k, the width of the sign modulated control pulse which will be enforced within the upcoming inter-sampling period of length T. The duty ratio function $\mu(x)$ must be regarded, effectively, as a feedback control policy.

The PWM controller design problem, carried out in this article, consists in specifying such a bounded feedback control policy $\mu(x)$ for stabilization and tracking purposes. The use of the following, infinite switching frequency, *average PWM controlled converter model*, for controller design purposes, has been thoroughly justified in several journal publications by Sira-Ramirez and his co-workers (see Sira-Ramirez, 1989; 1991a; Sira-Ramirez and Prada-Rizzo, 1992):

$$\left.\begin{aligned} \dot{z}_1 &= -w_0 z_2 + \mu b \\ \dot{z}_2 &= w_0 z_1 - w_1 z_2 \end{aligned}\right\}, \tag{2.3}$$

where μ is now regarded as a smooth function of the average state vector z, taking values in the closed interval $[-1, \ 1]$.

Given a constant value U of the duty ratio μ, such that $U \in [-1, \ 1]$, the corresponding average equilibrium state of the FBBC is simply obtained from (2.3) as

$$z_1 = Z_1(U) = \frac{b w_1}{w_0^2} U; \quad z_2 = Z_2(U) = \frac{b}{w_0} U. \tag{2.4}$$

A steady state relationship can be derived from Eq. (2.4) for the average equilibrium values of the normalized input circuit current z_1 and the normalized output capacitor voltage z_2, which is independent of the constant value U of the duty ratio function. Such relation is readily obtained as

$$Z_1 = \frac{w_1}{w_0} Z_2. \tag{2.5}$$

Hence, regulation of the average output capacitor voltage z_2 towards constant equilibrium values, specified by (2.4), can be indirectly accomplished through regulation of the average input inductor current z_1 toward its equilibrium value Z_1. This simple observation also allows for the possibility of specifying a dynamical feedback regulator instead of a static one (Boudjema et al., 1989). For tracking problems related to the automatic generation of regulated AC signals, such an indirect control policy will require partial inversion of the FBBC dynamical equations.

2.2 Stabilization of the FBBC through dynamical PWM control

Consider the average model (2.3) of the FBBC with output equation

$$\left.\begin{aligned} \dot z_1 &= -w_0 z_2 + \mu b \\ \dot z_2 &= w_0 z_1 - w_1 z_2 \\ y &= z_1 - Z_1 = z_1 - \frac{w_1}{w_0}\cdot Z_2 \end{aligned}\right\}. \tag{2.6}$$

The following invertible duty-ratio (i.e., control) dependent state coordinate transformation,

$$\left.\begin{aligned} \xi_1 &= z_1 - \frac{w_1}{w_0}\cdot Z_2, \quad \xi_2 = -w_0 z_2 + \mu b \\ z_1 &= \xi_1 + \frac{w_1}{w_0}\cdot Z_2, \quad z_2 = \frac{-\xi_2 + \mu b}{w_0} \end{aligned}\right\} \tag{2.7}$$

takes the average model (2.6) into Fliess's GOCF (Fliess, 1990 a)

$$\left.\begin{aligned} \dot\xi_1 &= \xi_2 \\ \dot\xi_2 &= -w_0^2 \xi_1 - w_0 w_1 Z_2 - w_1 \xi_2 + b w_1 \mu + \dot\mu b \\ y &= \xi_1 \end{aligned}\right\}. \tag{2.8}$$

Notice that if a feedback control is specified which asymptotically stabilizes the transformed coordinates ξ_1 and ξ_2 to zero (i.e., the system output y and its time derivative dy/dt are stabilized to zero), then, according to the definitions in (2.7), one obtains, respectively, asymptotic stabilization of the original average state coordinates z_1 and z_2 to their equilibrium values Z_1 and Z_2.

The *zero dynamics*, associated to the steady state equilibrium condition: $\xi_1 = 0$, $\xi_2 = 0$ is obtained by equating the last differential equation in (2.8) to zero (see Fliess, 1990 b)

$$\dot\mu = -\ w_1(\mu - \frac{w_0}{b}Z_2) = -\ w_1(\mu - U), \tag{2.9}$$

which evidently implies an asymptotically stable solution towards $\mu = U$; i.e., the achieved equilibrium corresponds to a minimum phase behavior.

An unrestricted dynamical feedback controller specifying the required (computed) stabilizing duty ratio function μ is immediately obtained from (2.8) by simply equating the last differential equation in (2.8) to a suitable linear combination of the transformed variables. This operation is to bestow on the closed loop transformed system convenient, stable, pole locations characterized by, say, a damping ratio ζ and a natural undamped frequency ω_n

$$\dot{\xi}_1 = \xi_2,$$

$$\dot{\xi}_2 = -w_0^2\xi_1 - w_0 w_1 Z_1 - w_1\xi_2 + bw_1\mu + \dot{\mu} b$$
$$= -2\zeta\omega_n\xi_2 - w_n^2\xi_1,$$

$$y = \xi_1,$$

i.e.,

$$\dot{\mu} = -w_1\mu + \left(\frac{w_1 - 2\zeta\omega_n}{b}\right)\xi_2 + \left(\frac{w_0^2 - \omega_n^2}{b}\right)\xi_1 + \frac{w_0 w_1}{b} Z_2. \quad (2.10)$$

We denote the solution of (2.10) as $\hat{\mu}$, and regard it as the *computed duty ratio function*.

It is quite straightforward to interpret the dynamical duty ratio synthesizer (2.10), expressed in transformed coordinates, as a classical proportional-derivative controller, followed by a low pass filter, with cut off frequency w_1 and a set-point feedforward term. Figure 2 depicts such an interpretation of the proposed controller (see also Sira-Ramirez, 1991 b; Sira-Ramirez and Lischinsky-Arenas, 1991).

In original average coordinates, one obtains the following dynamical feedback controller:

$$\frac{d}{dt}\hat{\mu} = -2\zeta\omega_n\hat{\mu} + \left(\frac{w_0^2 - \omega_n^2}{b}\right)z_1 + \frac{(2\zeta\omega_n - w_1)}{b} w_0 z_2 + \frac{\omega_n^2 w_1}{b w_0} Z_2. (2.11)$$

The *actual duty ratio function* μ is simply obtained by bounding the solutions of (2.11) within the physically meaningful interval $[-1, \ 1]$; i.e.,

$$\mu = \begin{cases} +1, & \text{if } \hat{\mu} > 1, \\ \hat{\mu}, & \text{if } -1 \leq \hat{\mu} \leq +1, \\ -1, & \text{if } \hat{\mu} < -1. \end{cases} \quad (2.12)$$

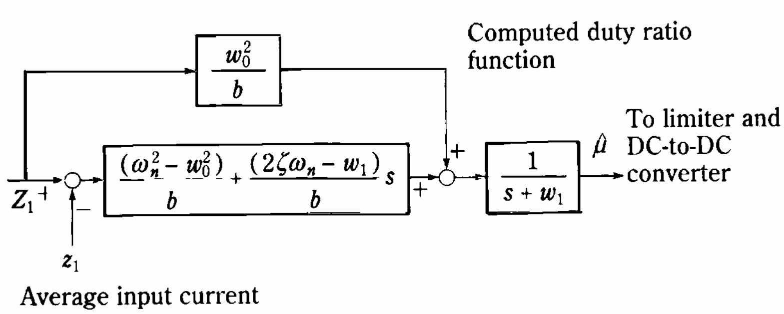


Fig. 2. Classical P-D compensator interpretation of average dynamical duty ratio synthesizer in a PWM controlled Full Bridge Buck Converter.

Figure 3 depicts the dynamical PWM controller scheme accomplishing asymptotic stabilization to a desirable average equilibrium point for the FBBC.

A simulation example. Simulations were carried out on an FBBC with parameter values: $R = 1.5[\Omega], C = 2700[\mu F]$, $L = 40[\mu H]$, $V_s = 30[V]$ and a transformer winding turn ratio $N = 10$, (i.e., $w_0 = 304.29$, $w_1 = 246.91$, $b = 4.743 \times 10^4$). The chosen damping ratio and natural oscillating frequency of the closed loop system stable complex poles were set, respectively, at $\zeta = 0.7$ and $w_n = 1000[\text{rad}/\text{s}]$. The required equilibrium value for the normalized output voltage was specified as 7.794, which corresponds to an actual output voltage of $15[V]$. The corresponding equilibrium value for the normalized input current set point is 6.324 (i.e., $I_L = 1[A]$). These equilibrium values correspond to a constant steady state value of the duty ratio function: $\mu = U = 0.5$. A sampling frequency of $2[KHz]$ was used for the proposed PWM controller. Figures 4(a) and 4(b) depict, respectively, the transient response of the average and the discontinuously PWM controlled input current and output voltage of the FBBC. Figure 4(c) represents the evolution of the dynamically synthesized duty ratio function μ and the corresponding discontinuous PWM control actions u.

2.3 Sinusoidal signal generation of the FBBC through dynamical feedback PWM control In this section, we propose the use of partial (state) inversion and dynamical PWM feedback control for the adequate tracking of sinusoidal template signals by the output capacitor voltage in an FBBC. Partial inversion is referred to the indirect tracking policy by which the average input inductor current is made to follow a periodical reference signal, which, in turn, corresponds to a perfect sinusoidal voltage at the output capacitor terminals in the FBBC. The natural limitations imposed on the synthesized duty ratio function imply some template amplitude-frequency tradeoffs, which can be explicitly computed in terms of the system parameters.

Suppose the average normalized output capacitor voltage z_2 is to track a (template) sinusoidal signal of the form

$$z_{2d}(t) = A \sin(\omega t). \tag{2.13}$$

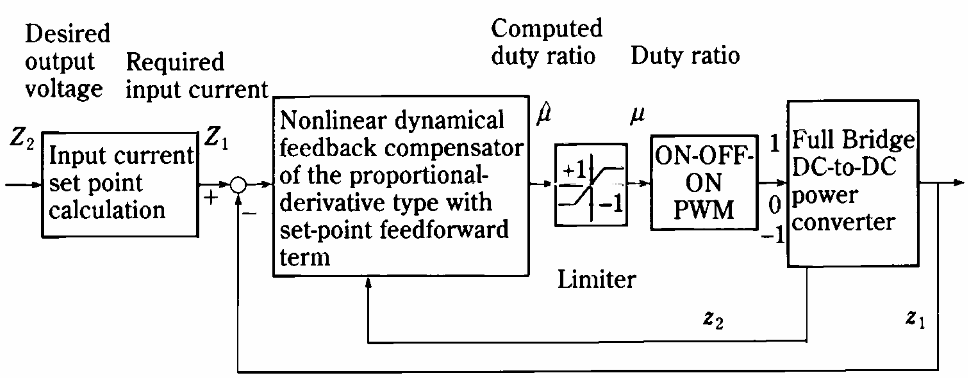
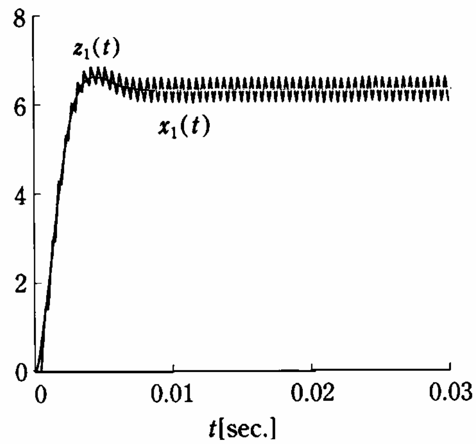
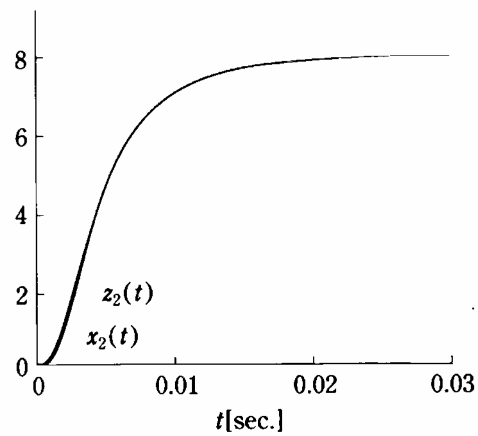


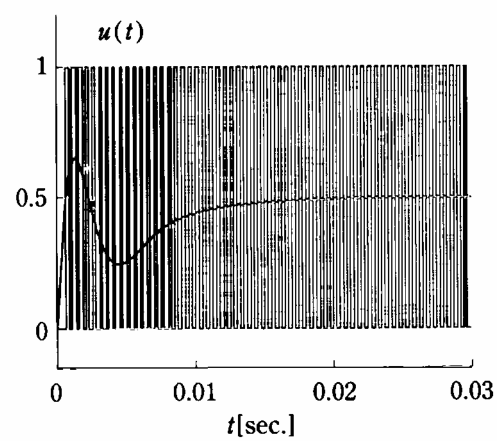
Fig. 3. Dynamical PWM feedback compensator structure for indirect output stabilization and tracking tasks in a Full Bridge Buck Converter.



(a) Average and actual input inductor current response for a dynamical PWM stabilization in a Full Bridge Buck Converter.



(b) Average and actual output capacitor voltage response for a dynamical PWM stabilization in a Full Bridge Buck Converter.



(c) Duty ratio function and control input responses in dynamical PWM stabilization in a Full Bridge Buck Converter.

Fig. 4.

Inversion of the average system dynamics (2.6) leads to the corresponding required reference signal for the average input inductor current z_1.

$$z_{1d}(t) = \frac{A}{w_0}\sqrt{\omega^2 + w_1^2}\,\sin\left(\omega t + \tan^{-1}\frac{\omega}{w_1}\right). \tag{2.14}$$

The following invertible time-varying input-dependent state coordinate transformation:

$$\left.\begin{aligned} \xi_1(t) &= z_1 - \frac{A}{w_0}\sqrt{\omega^2 + w_1^2}\,\sin\left(\omega t + \tan^{-1}\frac{\omega}{w_1}\right) \\ \xi_2(t) &= -w_0 z_2 + \mu b - \frac{A\omega}{w_0}\sqrt{\omega^2 + w_1^2}\,\cos\left(\omega t + \tan^{-1}\frac{\omega}{w_1}\right) \end{aligned}\right\} \tag{2.15}$$

places the average FBBC model into Fliess's GOCF

$$\left.\begin{aligned} \dot{\xi}_1 &= \xi_2 \\ \dot{\xi}_2 &= -w_0^2\xi_1 - w_1\xi_2 + \mu b w_1 + \dot{\mu} b \\ &\quad - \frac{A}{w_0}\sqrt{(\omega^2 + w_1^2)[(\omega w_1)^2 + (\omega^2 - w_0^2)^2]} \\ &\quad \times \sin\left[\omega t + \tan^{-1}\left(\frac{\omega}{w_1}\right) + \tan^{-1}\left(\frac{\omega w_1}{w_0^2 - \omega^2}\right)\right] \\ y &= \xi_1 \end{aligned}\right\}. \tag{2.16}$$

Asymptotic stabilization of the output ξ_1 to zero implies perfect tracking of the input current reference signal z_{1d} by the average input inductor current z_1 and, hence, indirect perfect tracking of the desired AC signal: $z_{2d} = A\sin\omega t$, by the average output capacitor voltage z_2.

The *zero dynamics* associated to system (2.16) is simply given by

$$\begin{aligned} \dot{\mu} &= -w_1\mu + \frac{A}{w_0 b}\sqrt{(\omega^2 + w_1^2)[(\omega w_1)^2 + (\omega^2 - w_0^2)^2]} \\ &\quad \times \sin\left[\omega t + \tan^{-1}\left(\frac{\omega}{w_1}\right) + \tan^{-1}\left(\frac{\omega w_1}{w_0^2 - \omega^2}\right)\right]. \end{aligned} \tag{2.17}$$

After some tedious but straightforward manipulations, one finds the steady state solution of the forced linear differential equation (2.17) as

$$\begin{aligned} \tilde{\mu}(t) &= \frac{A}{w_0 b}\sqrt{(\omega w_1)^2 + (\omega^2 - w_0^2)^2} \\ &\quad \times \sin\left[\omega t + \tan^{-1}\left(\frac{\omega w_1}{w_0^2 - \omega^2}\right)\right]. \end{aligned} \tag{2.18}$$

A dynamical duty ratio synthesizer is readily obtained by equating the last differential equation in (2.16) to an appropriate linear expression in the transformed coordinates, which ensures proper stable pole location in the complex plane. Proceeding as in the stabilization case of Sec. 2.2, one obtains the following time-varying differential equation for the computed duty ratio function $\hat{\mu}$ in original average state coordinates:

$$\frac{d}{dt}\hat{\mu} = -2\zeta\omega_n\hat{\mu} + \left(\frac{w_0^2-\omega_n^2}{b}\right)z_1 + \frac{2\zeta\omega_n - w_1}{b}w_0 z_2$$
$$+ \frac{A}{w_0 b}\sqrt{(\omega^2+w_1^2)[(\omega_n^2-w_0^2)^2+(2\zeta\omega_n\omega)^2]}$$
$$\times \sin\left[\omega t + \tan^{-1}\left(\frac{\omega}{w_1}\right) + \tan^{-1}\left(\frac{2\zeta\omega_n\omega}{\omega_n^2-w_0^2}\right)\right]. \quad (2.19)$$

The actual duty ratio function μ is obtained via appropriate bounding of the solution of (2.19) as it was expressed before in (2.12).

It should be clear that perfect tracking of the desired output voltage is achieved whenever the computed duty ratio function $\hat{\mu}$ does not saturate beyond the bounding values prescribed by the interval $[-1,\ 1]$; i.e., when the actual and the computed duty ratio functions coincide. In order to guarantee that, under steady state operating conditions, the amplitude of the synthesized duty ratio function μ does not violate the bounding limits of the interval $[-1,\ 1]$, one imposes, on the amplitude of the steady state sinusoidal wave given in (2.18), the following frequency-dependent magnitude restriction:

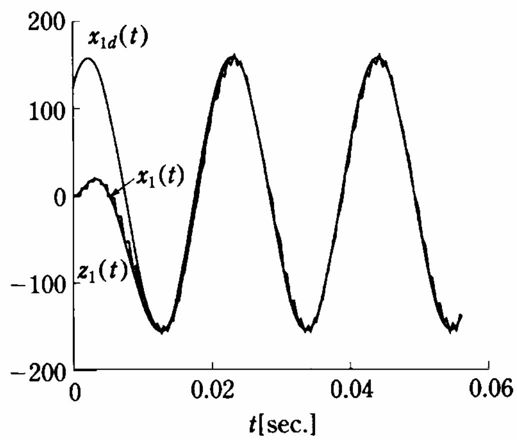
$$\frac{A}{w_0 b}\sqrt{(\omega w_1)^2+(\omega^2-w_0^2)^2} < 1 \Leftrightarrow A < \frac{w_0 b}{\sqrt{(\omega w_1)^2+(\omega^2-w_0^2)^2}}. \quad (2.20)$$

The above "tradeoff" relation only involves the FBBC parameters w_0, w_1, b and the defining parameters of the desired (template) reference sinusoidal output voltage amplitude A and frequency ω.

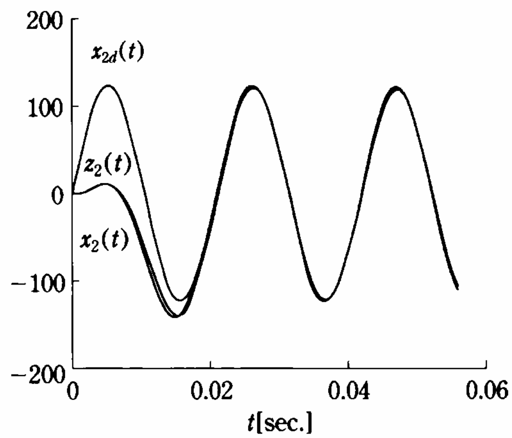
A simulation example. Simulations were carried out on an FBBC with the same parameter values as in the previous simulation example, except for the fact that here, V_s was taken as $300[V]$. The chosen damping ratio and natural oscillating frequency of the closed loop system stable complex poles were set, respectively, at $\zeta = 0.7$ and $\omega_n = 300[\text{rad/s}]$. The required normalized sinusoidal output voltage was specified as $z_{2d} = 120\sin(314t)$. Figures 5 (a) and 5 (b) depict, respectively, the transient response of the average and the discontinuously PWM controlled input current and output voltage of the FBBC in comparison with the corresponding AC reference signals. Figure 5 (c) represents the evolution of the synthesized duty ratio feedback function μ, the steady state duty ratio signal and the corresponding discontinuous PWM control actions u.

3. Conclusions

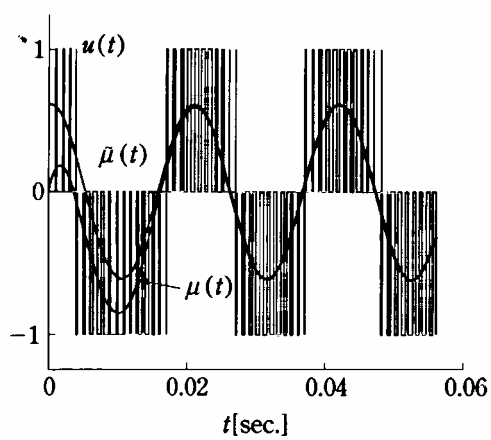
In this article, a dynamical PWM control scheme has been presented for both stabilization and tracking problems in DC-to-DC power converters of the Full Bridge Buck type. The approach was based on synthesizing a dynamical, unrestricted, smooth feedback control law specifying the required duty ratio function for the regulation of the infinite frequency average model of the PWM controlled converter. Such a dynamical scheme corresponds to a pole placement approach which imposes a desirable stable dynamics on Fliess's Generalized Observability Canonical form of the error dynamics associated with the average stabilization, or tracking, problem defined on the bridge converter. Aside from the high response quality of the controlled variables, the approach also allows for an explicit determination of the stabilization and tracking limitations



(a) Average and actual input inductor current response for dynamical PWM AC tracking task in a Full Bridge Buck Converter.



(b) Average and actual output capacitor voltage response for dynamical PWM AC tracking task in a Full Bridge Buck Converter.



(c) Duty ratio function and control input responses for dynamical PWM AC tracking task in a Full Bridge Buck Converter.

Fig. 5.

exhibited by the treated class of "step down" converters. In the tracking task, where sinusoidal reference signal tracking is demanded from the output capacitor voltage of the converter, the amplitude-frequency tradeoffs, inherent in the physically meaningful limitations of the duty ratio function, were explicitly computed.

References

Boudjema, F., M. Boscardin, P. Bidan, J. C. Marpinard, M. Valentin and J. L. Abatut (1989). VSS approach to a Full Bridge Buck Converter used for AC sine voltage generation. *Proc. of the 15th IEEE Annual Conference of the Industrial Electronics Society (IECON-89)*, Philadelphia, 82–89.

Csaki, F., K. Gansky, I. Ipsitz and S. Marti (1983). *Power Electronics*. Akademia Kiado, Budapest, Hungary.

Fliess, M. (1990 a). Nonlinear control theory and differential algebra. Ch.I. Byrnes and A. Kurzhanski (ed.), *Modeling and Adaptive Control*, Lecture Notes in Control and Information Sciences, **105**, Springer-Verlag, N.Y.

Fliess, M. (1990 b). What the Kalman state variable representation is good for. *Proc. of the 29th IEEE Conference on Decision and Control*, Honolulu, Hawaii, **3**, 1282–1287.

Mohan, N., T. M. Undeland and W. P. Robbins (1989). *Power Electronics, Converters, Applications and Design*. John Wiley and Sons, N. Y.

Severns, R. P. and G. Bloom (1985). *Modern DC-to-DC Switch Mode Power Converter Circuits*. Van Nostrand-Reinhold, N. Y.

Sira-Ramirez, H. (1987). Sliding motions in bilinear switch networks. *IEEE Trans. Circuits and Systems*, **CAS-34**, 8, 919–933.

Sira-Ramirez, H. (1989). A geometric approach to pulse-width-modulated control in nonlinear dynamical systems. *IEEE Trans. Automatic Control*, **AC-34**, 2, 184–187.

Sira-Ramirez, H. (1991 a). Nonlinear PI controller design for switchmode DC-to-DC power converters. *IEEE Trans. Circuits and Systems*, **CAS-38**, 4, 410–417.

Sira-Ramirez, H. (1991 b). Dynamical feedback regulation in DC-to-DC power converters. *Proc. of the IEEE Workshop on Variable Structure Control of Power Conversion Systems*, University of Nevada, Reno.

Sira-Ramirez, H. (1991 c). Nonlinear dynamical discontinuous feedback controlled descent on a non atmosphere-free planet: A differential algebraic approach. *Control–Theory and Advanced Technology* (*C-TAT*), **7**, 2, 301–320.

Sira-Ramirez, H. and M.T. Prada-Rizzo (1992). Nonlinear feedback regulator design for the Cúk converter. *IEEE Trans. Automatic Control*, **AC-37**, 8, 1173–1180.

Sira-Ramirez, H. and P. Lischinsky-Arenas (1991). The differential algebraic approach in nonlinear dynamical compensator design for DC-to-DC power converters. *Int. J. Control*, **54**, 1, 111–134.

Hebertt Sira-Ramírez obtained the degree of Ingeniero Electricista from the Universidad de Los Andes in Mérida-Venezuela in 1970. He pursued graduate studies from 1972 to 1977 at the Massachusetts Institute of Technology (MIT), Cambridge, MA, where he obtained the degree of Master of Science in Electrical Engineering (MSEE) and the degree of Electrical Engineer (EE) in 1974. He later obtained the degree of Doctor of Philosophy (Ph.D.) in Electrical Engineering in 1977. He is currently a Full Professor in the Control Systems Department of the Systems Engineering School of the Universidad de Los Andes, where he has also held the positions of Head of the Control Systems Department from 1978 to 1980 and was elected Vice-president of the University for the term 1980-1984. He was appointed Head of the Graduate School in Control Engineering in 1992. Dr. Sira-Ramírez is a Senior Member of the Institute of Electrical and Electronics Engineers (IEEE), where he serves as a member of the IEEE International Committee. He is also a member of the International Federation of Automatic Control (IFAC), Sigma Xi, The Planetary Society and of the Venezuelan College of Engineers (CIV).

He has held visiting positions in the Coordinated Science Laboratory of the University of Illinois at Urbana-Champaign, in the Department of Aeronautical and Astronautical Engineering and the Department of Electrical Engineering of the same University in 1986. He has also held brief visiting positions in the School of Electrical Engineering of Purdue University in 1989, 1990 and 1991. He was appointed Associate Research Director at the Laboratoire de Signaux et Systems of the CNRS in Plateau du Moulon, (Paris) France in 1992. Dr. Sira-Ramírez received the Venezuelan College of Engineers Award for Scientific Research in 1987 and the Senior Researcher Scholarship Award from the Venezuelan National Council for Scientific and Technological Research (CONICIT) in 1990. Dr. Sira-Ramírez is interested in the theory and applications of discontinuous feedback control strategies for nonlinear dynamic systems.

Marco Tulio Prada-Rizzo was born in Tabay, Mérida (Venezuela) in 1951. He obtained the degree of Ingeniero Electricista from the Universidad de Los Andes in Mérida in 1977. Mr. Prada-Rizzo was appointed as Chief Executive Officer of CADAFE, the Utilitiy Company serving the Merida Region, for the period 1979-1984. He pursued graduate studies and obtained the degree of Master of Science in Automatic Control Engineering from the University of the Andes in 1991. Mr. Prada-Rizzo is an Aggregate professor in the Department of Circuits and Systems of the School of Electrical Engineering of the Universidad de Los Andes. He is interested in the applications of Automatic Control Theory to Power Electronics and other Electrical Engineering problems.