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Passivity-Based Controllers for the Stabilization of DC-to-DC Power Converters *

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Abstract

Average models of PWM regulated DC-to-DC Power supplies are shown to be Euler-Lagrange Systems. As such, passivity-based dynamical feedback controllers can be derived for the indirect stabilization of the average output voltage. The derived controllers are based on a suitable stabilizing "damping injection" scheme. The approach is applied to regulate dc-to-dc power converters of the "boost" and "buck-boost" types. The effectivity and robustness of the proposed duty ratio synthesis policies are tested, via computer simulations, on a stochastically perturbed model of a switched "boost" converter.

Keywords : DC-to-DC Power Converters, Passivity based regulation, Euler-Lagrange systems.

1 Introduction

Regulation of DC-to-DC Power supplies has been extensively studied from a Sliding Mode control viewpoint (see [1], [2]) and also from a Pulse-Width-Modulation (PWM) feedback control alternative (see [3]). Both approaches entirely overlook the physical properties of the system and blindly insist on average closed-loop linearization in order to solve stabilization and tracking tasks.

Recently, a feedback control methodology has been developed which is aimed at modifying the closed loop energy dissipation and potential energy properties of nonlinear systems. The approach, known as *passivity-based* controller design, has been successfully used in the regulation of Euler-Lagrange (EL) systems, such as robotic manipulators and electromechanical energy conversion devices (see [4], [5] and [6]).

In this article a passivity-based approach is proposed for the average dynamical stabilization of PWM controlled dc-to-dc power converters of the "boost" and "buck-boost" types. The physically motivated duty ratio synthesizers, or regulators, are derived to achieve *indirect* average output voltage stabilization for the treated power converters. The controllers are based exclusively on a *damping injection* scheme. Due to the non-minimum phase character of the underlying input-output systems, direct output voltage stabilization is shown to be unfeasible. The performance of one of the derived indirect controllers was successfully tested, via computer simulations, on a stochastically perturbed version of the switch-regulated model of a typical "boost" converter circuit.

Section 2 presents an EL based derivation of the average PWM models of the "boost" and "buck-boost" converters. Section 3 develops the passivity-based feedback controllers and demonstrates, for the "boost" converter case the non-minimum phase character of the direct regulation option. Section 4 presents simulation results testing the validity and robustness of the proposed dynamical control schemes. Section 5 contains the conclusions and suggestions for further research.

2 Average Models of DC-to-DC Converters as EL Systems

2.1 The "Boost" Converter

Consider the switch-regulated "Boost" converter circuit of Figure 1. The differential equations describing the system are given by

$$\begin{aligned} \dot{x}_1 &= -(1-u)\,\frac{1}{L}x_2 + \frac{E}{L} \\ \dot{x}_2 &= (1-u)\,\frac{1}{C}x_1 - \frac{1}{RC}x_2 \end{aligned} \tag{2.1}$$

where x_1 and x_2 represent, respectively, the input inductor current and the output capacitor voltage variables. The positive quantity E represents the constant voltage value of the external source. The variable u denotes the switch position function, acting as

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a control input. Such a control input takes values on the discrete set $\{0,1\}$.

A PWM based regulation policy for the switch position function may be specified as follows,

$$u = \begin{cases} 1 \text{ for} & t_k \leq t < t_k + \mu(t_k)T \\ 0 \text{ for} & t_k + \mu(t_k)T \leq t < t_k + T \end{cases}$$
$$t_{k+1} = t_k + T \quad ; \quad k = 0,1,\ldots \qquad (2.2)$$

where t_k represents a sampling instant; the parameter T is the fixed sampling period, also called the *duty cycle*. The function, $\mu(\cdot)$, is the *duty ratio function* acting as a truly feedback policy. The value of the duty ratio function, $\mu(t_k)$, determines, at every sampling instant, t_k, the width of the upcoming "pulse" (switch at the position $u = 1$) as $\mu(t_k)T$. The actual duty ratio function, $\mu(\cdot)$, is evidently a function limited to the closed interval $[0,1]$ on the real line.

We consider separately the Lagrangian dynamics formulation of the two circuits associated with each one of the two possible positions of the regulating switch. We express the input current x_1 in terms of the derivative of the circulating charge q_L, denoted by $\dot{q}_L$. Also, the capacitor voltage x_2 will be rewritten as q_C/C where q_C is the electrical charge stored in the output capacitor.

Consider first the switch position value $u = 1$. In this case, the corresponding lagrangian dynamics formulation of the circuit is as follows.

Define $\mathcal{T}_1(\dot{q}_L)$ and $\mathcal{V}_1(q_C)$ as the kynetic and potential energies of the circuit, respectively. Let $\mathcal{D}_1(\dot{q}_C)$ denote the Rayleigh dissipation function. The EL parameters of the circuit are readily found to be,

$$\begin{aligned} \mathcal{T}_1(\dot{q}_L) &= \frac{1}{2}L(\dot{q}_L)^2 \quad ; \quad \mathcal{V}_1(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_1(\dot{q}_C) &= \frac{1}{2}R(-\dot{q}_C)^2 \quad ; \quad \mathcal{F}_{q_L} = E \; ; \; \mathcal{F}_{q_C} = 0 \end{aligned} \qquad (2.3)$$

where $\mathcal{F}_{q_L}$ and $\mathcal{F}_{q_C}$ are the *generalized forcing* functions associated with the coordinates q_L and q_C, respectively.

Consider now the case $u = 0$. These corresponding EL parameters of the resulting circuit are, in this case

$$\begin{aligned} \mathcal{T}_0(\dot{q}_L) = \frac{1}{2}L(\dot{q}_L)^2 \quad &; \quad \mathcal{V}_0(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_0(\dot{q}_L,\dot{q}_C) &= \frac{1}{2}R(\dot{q}_L - \dot{q}_C)^2 \\ \mathcal{F}_{q_L} = E \quad &; \quad \mathcal{F}_{q_C} = 0 \end{aligned} \qquad (2.4)$$

The EL parameters of the two circuits, generated by the two switch position values, result in identical kynetic and potential energies as well as identical forcing functions. The switching action only changes the Rayleigh dissipation function between the values $\mathcal{D}_0(\dot{q}_C)$ and $\mathcal{D}_1(\dot{q}_L,\dot{q}_C)$.

A possible set of *average* EL parameters is obtained by retaining, for the average system, all of the switch–invariant EL parameters while "modulating", by means of the appropriate fraction of time, those EL parameters which are modified by the change of switch position. The following set of EL parameters are then proposed as the average EL parameters corresponding to a certain *average* PWM circuit,

$$\begin{aligned} \mathcal{T}_\mu(\dot{q}_L) = \frac{1}{2}L(\dot{q}_L)^2 \quad &; \quad \mathcal{V}_\mu(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_\mu(\dot{q}_L,\dot{q}_C) &= \frac{1}{2}R[(1-\mu)\dot{q}_L - \dot{q}_C]^2 \\ \mathcal{F}^\mu_{q_L} = E \quad &; \quad \mathcal{F}^\mu_{q_C} = 0 \end{aligned} \qquad (2.5)$$

Note that when μ takes the extreme values $\mu = 1$ and $\mu = 0$, one recovers, from the proposed dissipation function $\mathcal{D}_\mu(\dot{q}_L,\dot{q}_C)$, of equation (2.5), the dissipation functions $\mathcal{D}_1(\dot{q}_C)$ in (2.3) and $\mathcal{D}_0(\dot{q}_L,\dot{q}_C)$ in (2.4), respectively.

The lagrangian function associated with the above defined average EL parameters is,

$$\mathcal{L}_\mu = \mathcal{T}_\mu(\dot{q}_L) - \mathcal{V}_\mu(q_C) = \frac{1}{2}L(\dot{q}_L)^2 - \frac{1}{2C}q_C^2 \qquad (2.6)$$

One obtains the set of differential equations corresponding to the proposed average EL parameters (2.5) by using the classical EL equations,

$$\begin{aligned} \frac{\mathrm{d}}{\mathrm{d}t}\left(\frac{\partial \mathcal{L}_\mu}{\partial \dot{q}_L}\right) - \frac{\partial \mathcal{L}_\mu}{\partial q_L} &= -\frac{\partial \mathcal{D}_\mu}{\partial \dot{q}_L} + \mathcal{F}^\mu_{q_L} \\ \frac{\mathrm{d}}{\mathrm{d}t}\left(\frac{\partial \mathcal{L}_\mu}{\partial \dot{q}_C}\right) - \frac{\partial \mathcal{L}_\mu}{\partial q_C} &= -\frac{\partial \mathcal{D}_\mu}{\partial \dot{q}_C} + \mathcal{F}^\mu_{q_C} \end{aligned} \qquad (2.7)$$

One obtains the following average system description,

$$\begin{aligned} L\ddot{q}_L &= -(1-\mu)R[(1-\mu)\dot{q}_L - \dot{q}_C] + E \\ \frac{q_C}{C} &= R[(1-\mu)\dot{q}_L - \dot{q}_C] \end{aligned} \qquad (2.8)$$

which can be rewritten, using $z_1 = \dot{q}_L$ and $z_2 = q_C/C$, and substituting the second equation of (2.8) into the first of (2.8), as

$$\mathcal{D}\dot{z} + (1-\mu)\mathcal{J}z + \mathcal{R}z = \mathcal{E} \qquad (2.9)$$

where

$$\begin{aligned} \mathcal{D} &= \begin{bmatrix} L & 0 \\ 0 & C \end{bmatrix} \quad ; \quad \mathcal{J} = \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix} \\ \mathcal{R} &= \begin{bmatrix} 0 & 0 \\ 0 & 1/R \end{bmatrix} \quad ; \quad \mathcal{E} = \begin{bmatrix} E \\ 0 \end{bmatrix} \end{aligned} \qquad (2.10)$$

where z_1 and z_2 denote the *average input current* and the *average output capacitor voltage*, respectively, of the PWM regulated "Boost" converter. Note that we distinguish the average state variables z_1, z_2, from the actual PWM regulated state variables, denoted by x_1 and x_2.

The proposed average dynamics coincides with the state average model developed in [7] and with the infinite switching frequency model, or Filippov average model, found in [8].

2.2 The "Buck–Boost" Converter

The "buck–boost" converter model (see Figure 2) is given by

$$\begin{aligned} \dot{x}_1 &= (1-u)\,\frac{1}{L}x_2 + u\frac{E}{L} \\ \dot{x}_2 &= -(1-u)\,\frac{1}{C}x_1 - \frac{1}{RC}x_2 \end{aligned} \quad (2.11)$$

where x_1 and x_2 are, respectively, the input inductor current and the output capacitor voltage. E is the external source voltage and u is switch position function.

The EL parameters corresponding to the switch position value $u = 1$ are.

$$\begin{aligned} \mathcal{T}_1(\dot{q}_L) &= \frac{1}{2}L\,(\dot{q}_L)^2 \quad ; \quad \mathcal{V}_1(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_1(\dot{q}_C) &= \frac{1}{2}R\,(-\dot{q}_C)^2 \quad ; \quad \mathcal{F}_{q_L} = E \; ; \; \mathcal{F}_{q_C} = 0 \end{aligned} \quad (2.12)$$

The EL parameters corresponding to $u = 0$ are

$$\begin{aligned} \mathcal{T}_0(\dot{q}_L) = \frac{1}{2}L\,(\dot{q}_L)^2 \quad &; \quad \mathcal{V}_0(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_0(\dot{q}_L,\dot{q}_C) &= \frac{1}{2}R\,(-\dot{q}_L - \dot{q}_C)^2 \\ \mathcal{F}_{q_L} = 0 \quad &; \quad \mathcal{F}_{q_C} = 0 \end{aligned} \quad (2.13)$$

A possible set of *average* EL parameters is obtained as

$$\begin{aligned} \mathcal{T}_\mu(\dot{q}_L) = \frac{1}{2}L\,(\dot{q}_L)^2 \quad &; \quad \mathcal{V}_\mu(q_C) = \frac{1}{2C}q_C^2 \\ \mathcal{D}_\mu(\dot{q}_L,\dot{q}_C) &= \frac{1}{2}R[-(1-\mu)\dot{q}_L - \dot{q}_C]^2 \\ \mathcal{F}^\mu_{q_L} = \mu E \quad &; \quad \mathcal{F}^\mu_{q_C} = 0 \end{aligned} \quad (2.14)$$

The lagrangian function associated with the above defined average EL parameters is the same as in (2.6)

The following set of differential equations is obtained from (2.7), in terms of the average state variables $z_1 = \dot{q}_L$ and $z_2 = q_C/C$,

$$\mathcal{D}\dot{z} + (1-\mu)\mathcal{J}z + \mathcal{R}z = \mu\mathcal{E} \quad (2.15)$$

where

$$\begin{aligned} \mathcal{D} &= \begin{bmatrix} L & 0 \\ 0 & C \end{bmatrix} \quad ; \quad \mathcal{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \\ \mathcal{R} &= \begin{bmatrix} 0 & 0 \\ 0 & 1/R \end{bmatrix} \quad ; \quad \mathcal{E} = \begin{bmatrix} E \\ 0 \end{bmatrix} \end{aligned} \quad (2.16)$$

3 Passivity Based PWM Controllers for DC–to–DC Power Converters

3.1 Controller Design Based on Damping Injection : The "boost" converter

In this section we derive two controllers, the first one attempts "direct" output voltage regulation while the second is based on input current regulation for "indirect" stabilization of the output voltage. It turns out that if a desired output voltage is required, a partial inversion of the system must be performed in order to calculate the corresponding input current trajectory, from which the necessary control actions can be immediately derived. Since the average system is non-minimum phase with respect to the output capacitor voltage, the resulting controller is unstable. On the other hand, when a particular input current is desired, the partial inversion that computes the required output voltage trajectory leads to a locally stable controller.

3.1.1 Direct output voltage regulation

Suppose it is desired to directly regulate the output capacitor voltage to a constant value, Y. A partial inversion of the systems dynamics reveals that the average input inductor current must converge to a value $z_{1d}(t)$ that satisfies,

$$z_{1d}(t) = \frac{Y}{R(1-\mu(t))} \quad (3.1)$$

Consider then the error variables $\tilde{z}_1(t) = z_1(t) - z_{1d}(t)$ and $\tilde{z}_2(t) = z_2(t) - Y$. The error dynamics is given by

$$\mathcal{D}\dot{\tilde{z}} + (1-\mu)\mathcal{J}\tilde{z} + \mathcal{R}\tilde{z} = \mathcal{E} - (\mathcal{D}\dot{z}_d + (1-\mu)\mathcal{J}z_d + \mathcal{R}z_d) \quad (3.2)$$

One may perform a *damping injection* on (3.2) by considering the following desired dissipation term,

$$\mathcal{R}_d\tilde{z} = (\mathcal{R} + \mathcal{R}_1)\,\tilde{z} \quad (3.3)$$

where

$$\mathcal{R}_1 = \begin{bmatrix} R_1 & 0 \\ 0 & 0 \end{bmatrix} \; ; \; R_1 > 0 \quad (3.4)$$

Adding to both sides of equation (3.2) the necessary expressions we obtain,

$$\begin{aligned} &\mathcal{D}\dot{\tilde{z}} + (1-\mu)\mathcal{J}\tilde{z} + \mathcal{R}_d\tilde{z} \\ &= \mathcal{E} - (\mathcal{D}\dot{z}_d + (1-\mu)\mathcal{J}z_d + \mathcal{R}z_d - \mathcal{R}_1\tilde{z}) \end{aligned} \quad (3.5)$$

Suppose for a moment that the right hand side of equation (3.5) is identically zero. Under these circumstances the stabilization error dynamics would satisfy,

$$\mathcal{D}\dot{\tilde{z}} + (1-\mu)\mathcal{J}\tilde{z} + \mathcal{R}_d\tilde{z} = 0 \quad (3.6)$$

Take as a Lyapunov function candidate the total energy associated with the stabilization error, namely

$$H_d = \frac{1}{2}\tilde{z}^T\mathcal{D}\tilde{z} \; > \; 0 \;\; \forall\, \tilde{z} \neq 0 \quad (3.7)$$

The time derivative of H_d along the solutions of (3.6) results, for some strictly positive constant α, in

$$\dot{H}_d = -\tilde{z}\mathcal{R}_d\tilde{z} \; \leq -\alpha\, H_d \; < \; 0 \quad \forall\, \tilde{z} \neq 0 \quad (3.8)$$

We conclude that the if the error dynamics coincides with (3.6), the stabilization error behavior is asymptotically stable to zero independently of the value of μ. Thus, in order to have (3.8) satisfied one must demand, from (3.5) that

$$\mathcal{D}\dot{z}_d + (1-\mu)\mathcal{J}z_d + \mathcal{R}z_d - \mathcal{R}_1\tilde{z} = \mathcal{E} \qquad (3.9)$$

These conditions are explicitly written as

$$\begin{aligned} L\dot{z}_{1d} + (1-\mu)z_{2d} - (z_1 - z_{1d})R_1 &= E \\ C\dot{z}_{2d} - (1-\mu)z_{1d} + \frac{1}{R}z_{2d} &= 0 \quad (3.10) \end{aligned}$$

The problem, thus, consists in, given a desired constant output voltage value $z_{2d} = Y$, find a bounded function $z_{1d}(t)$, and a suitable duty ratio function μ, such that (3.10) is satisfied. Substituting expression (3.1), which, incidentally, can also be obtained from the second of (3.10), into the first equation of (3.10), one obtains, after some algebraic manipulations, an expression for the dynamical feedback duty ratio synthesizer of the form,

$$\dot{\mu} = -\frac{R(1-\mu)^2}{LY}\left[E - (1-\mu)Y + R_1\left(z_1 - \frac{Y}{R(1-\mu)}\right)\right] \qquad (3.11)$$

This controller stabilizes z_1 and z_2 towards their desired values z_{1d} and z_{2d}, respectively. However controller (3.11) is, unfortunately, unfeasible due to its lack of stability. Indeed the "remaining", or zero, dynamics associated with the above controller results in

$$\dot{\mu} = -\frac{R(1-\mu)^2}{LY}\left[E - (1-\mu)Y\right] \qquad (3.12)$$

which has two equilibrium points, namely,

$$\mu = 1 \quad ; \quad \mu = 1 - \frac{E}{Y} \qquad (3.13)$$

A phase diagram of (3.12) readily reveals that the second equilibrium point is unstable while the first one being marginally stable represents a fixed switch position that renders the input current of the converter unstable.

3.1.2 Indirect output voltage regulation

The previous section has shown that the only left alternative is to *indirectly* try to control the output capacitor voltage through regulation of the input current.

Suppose it is desired to regulate z_1 towards a constant value $z_{1d} = I$. Corresponding to such an equilibrium value for z_1 one finds, by partial system inversion, or simply by using the second equation of (3.10), that $z_2(t)$ should adopt the value given by,

$$z_{2d}(t) = \frac{E + (z_1 - I)R_1}{(1-\mu(t))} \qquad (3.14)$$

Substituting (3.14) into the second equation of (3.10), one obtains, after some algebraic manipulations,

$$\begin{aligned} \dot{\mu} &= \frac{(1-\mu)}{C\left[E + (z_1 - I)R_1\right]}\Big\{(1-\mu)^2 I \\ &\quad - \frac{E + (z_1 - I)R_1}{R} - \frac{R_1 C}{L}\left[E - (1-\mu)z_2\right]\Big\} \end{aligned} \qquad (3.15)$$

The "remaining" dynamics associated with controller (3.15), is obtained by lettting z_1 and z_2 coincide with their respective desired values. Such dynamics is given by

$$\dot{\mu} = \frac{1-\mu}{RCE}\left[(1-\mu)^2 RI - E\right] \qquad (3.16)$$

The equilibrium points associated with the zero dynamics (3.16) are, in this case,

$$\mu = 1 \ ; \ \mu = 1 - \sqrt{\frac{E}{IR}} \ ; \ \mu = 1 + \sqrt{\frac{E}{IR}} \qquad (3.17)$$

The only physically significant equilibrium point is, in this case, the second one, provided $E < IR$, thus revealing the amplifying features of the converter. A phase diagram of (3.16) reveals that the second equilibrium point is locally asymptotically stable, while the other two are unstable.

3.2 Controller Design Based on Damping Injection : The "buck–boost" converter

Following exactly the same procedure as in the previous case one concludes that direct regulation of output voltage capacitor, for the "buck–boost" converter, is unfeasible due to non–minimum phase phenomena. We write down the expression for the dynamical feedback controller achieving indirect output capacitor voltage regulation through input current stabilization.

$$\begin{aligned} \dot{\mu} &= \frac{1-\mu}{C\left[E + (z_1 - I)R_1\right]}\Big\{(1-\mu)^2 I \\ &\quad - \frac{\mu E + (z_1 - I)R_1}{R} - \frac{R_1 C}{L}\left[\mu E + z_2(1-\mu)\right]\Big\} \end{aligned} \qquad (3.18)$$

The zero dynamics associated with the controller (3.18) is given by

$$\dot{\mu} = \frac{(1-\mu)}{RCE}\left\{(1-\mu)^2 RI - \mu E\right\} \qquad (3.19)$$

which has three equilibrium points,

$$\mu = 1 \ ; \ \mu = 1 + \frac{E}{2RI} \pm \sqrt{\left(\frac{E}{2RI}\right)^2 + \frac{E}{RI}} \qquad (3.20)$$

Two of the equilibrium points ($\mu = 1$, and the one corresponding to the plus sign of the square root) are unstable while the remaining one, which is the only physically significant one, is locally asymptotically stable.

4 Simulation Results

Simulations were performed for a typical PWM regulated "boost" converter circuit with circuit parameter values given by

$$C = 20\ \mu\mathrm{F}\ \ ;\ R = 30\ \Omega\ \ ;\ L = 20\ \mathrm{mH}\ \ ;\ E = 15\ \mathrm{V}$$

and a sampling frequency of 3 KHz for the PWM policy. The duty ratio function is obtained from a sampling process carried out on the output $\mu(t)$ of the dynamical controller derived above. To avoid the construction of low pass filters, instead of using the averaged state variables z_1, z_2 for feedback on the duty ratio synthesizers, one may go ahead and use the actual PWM controlled states x_1, x_2 on the controllers expressions. The desired input inductor current was set to be $I = 3.125$ Amp., with a steady state duty ratio of $U = 0.6$. This corresponds with a desired output voltage $x_2 = Y = 37.5$ Volts. Figure 3 shows the closed loop state trajectories corresponding to the feasible duty ratio synthesizer derived for the "boost" converter. Figure 4 depicts the realization of the computer generated stochastic perturbation signal directly added to the external source voltage E in (2.1). As it can be seen from the simulations, the proposed controller achieves the desired stabilization while exhibiting a high degree of robustness with respect to the stochastic perturbation input.

5 Conclusions

Physically motivated dynamical feedback duty ratio synthesizers were derived for the indirect output voltage stabilization of dc–to–dc power converters of the "boost" and "buck-boost" types. The controllers are based on the preservation of the passivity of the average converter circuit model and the possibilities of damping injection through dynamical feedback. The study revealed that direct output regulation leads to an unstable contoller in both types of converters, thus limiting the passivity-based controllers to minimum-phase cases. The obtained simulation results for the "boost" converter case were highly encouraging so as to attempt actual implementation in the future.

Some other useful connections of pasivity-based controllers with the ***differential flatness*** associated with the average PWM models of dc–to–dc power converters remain to be explored. Similarly, Sliding Mode controllers as well as various kinds of adaptive feedback control schemes, based on passivity considerations, are to be developed for dc power supplies.

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FIGURES

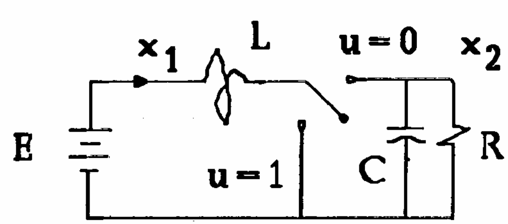

Figure 1: "Boost" Converter Circuit.

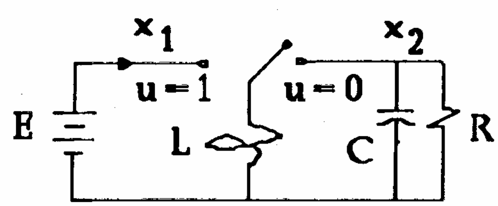

Figure 2: "Buck-Boost" Converter Circuit.

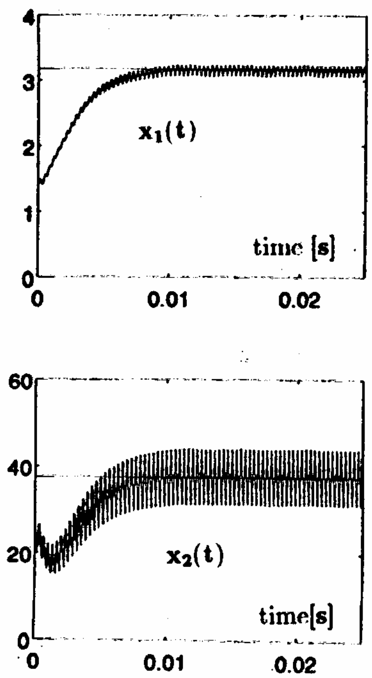

Figure 3: Boost Converter PWM Controlled State Trajectories: Damping Injection-based Controller

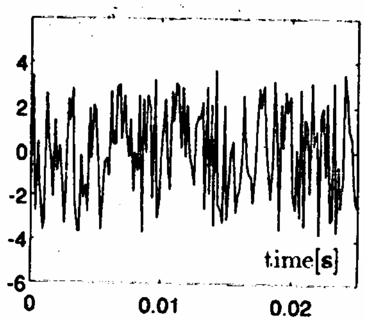

Figure 4: Stochastic Perturbation Input affecting the Voltage Source