Sliding Mode $\Sigma$-$\Delta$ Modulation Control of the Boost Converter

Hebertt Sira-Ramírez and Ramón Silva-Ortigoza

ABSTRACT

A switched implementation of average dynamic output feedback laws through a $\Sigma$-$\Delta$-modulator, widely known in the classic communications and analog signal encoding literature, not only frees the sliding mode control approach from state measurements and the corresponding synthesis of sliding surfaces in the plant’s state space, but it also allows to effectively transfer all desired closed loop features of an uniformly bounded, continuous, average output feedback controller design into the more restrictive discrete-valued (ON-OFF) control framework of a switched system. The proposed approach is here used for the input-output sliding mode stabilization of the “boost” DC-to-DC converter. This is achieved by means of a well known passivity based controller but any other output feedback design would have served our purposes. This emphasizes the flexibility of the proposed sliding mode control design implementation through $\Sigma$-$\Delta$-modulators.

KeyWords: Sliding mode control, delta-modulation, sigma-delta modulation.

I. INTRODUCTION

The many advantages of sliding mode control are well reported, founded, and illustrated, in the existing literature. The sliding mode control technique is, fundamentally, a state space-based discontinuous feedback control technique. The lack of complete knowledge of the state vector components forces the designer to use asymptotic state observers, of the Luenberger, or of the sliding mode type [1], or perhaps to resort to direct output feedback control schemes (See [2]). Unfortunately, the first approach is not robust with respect to unforeseen exogenous perturbation inputs, even if they happen to be of the “classical type” (by this we mean: steps, ramps, parabolas, etc). The second approach is quite limited in nature and it is not directly applicable in a host of non-minimum phase systems.

Generically speaking, state space based sliding mode techniques fall into the unmatched perturbation input case, while output feedback control techniques do not suffer such realistic drawback. For general background on sliding mode control, we refer the reader to the seminal book by Utkin, [3], the recent books by Utkin, Gulder and Shi [1] and that by Edwards and Spurgeon [2]. Recent developments, advances and applications, of the sliding mode control area can be found in the book by Perruquetti and Barbot [4].

In this article, we propose a new approach for the synthesis of sliding mode feedback control schemes for a popular switched DC-to-DC power converter. We propose a sliding mode implementation, based on an analog version of $\Sigma$-$\Delta$ modulators, of an average dynamic output feedback controller design scheme which is known to be available for switched power converters. We show that the use of analog $\Sigma$-$\Delta$-modulators' allows for the switched synthesis of any feedback controller which has been synthesized.

1 A complete account of $\Delta$-modulators, and their simplest modification: $\Sigma$-$\Delta$ modulators, extensively used in analog signal encoding, which never benefited from the theoretical basis of sliding mode control, is found in the classical book by Steele [5] and in the excellent book by Norsworthy et al [6].
from an average viewpoint (*i.e.* assuming that the control input continuously takes values on a closed subset of the real line, usually restricted to be the closed interval [0,1]). We show that a \(\Sigma-\Delta\)-modulator can be used to translate such a continuous design into a discontinuous one with the property that the "equivalent output" signal of the modulator, in an ideal sliding mode sense, precisely matches the modulator’s input signal generated by the continuous output feedback controller.

When we combine \(\Sigma-\Delta\)-modulation with dynamic output feedback control, the result is that the required sliding motion is dynamically synthesized using only the input and the output of the system while retaining all the desirable essential features of the average devised controller (robustness, adaptability, perturbation rejection properties, etc). The proposed approach points to a systematic approach to controller design for switched systems based on average designs. In this respect, the proposed approach is quite different from that found in Yeung et al [7] where the sliding surface is synthesized in terms of (filtered) differential polynomials acting on inputs and outputs (See also [8] for a yet different perspective). As an additional outcome, the scheme here presented requires no "matching conditions" whatsoever. In this article, we particularly advocate a class of dynamic output feedback control for the average controller design in DC-to-DC power converters: namely, passivity based control.

Section 2 presents a review of an analog \(\Sigma-\Delta\)-modulator and its connection with sliding mode control schemes when the actual system input signal takes values in a discrete set of the "ON-OFF" form, *i.e.*, in the discrete set \(\{0,1\}\). Section 3 deals with some generalities on how to synthesize a switched sliding mode controller on the basis of a given continuous feedback controller design. Section 4 concentrates on a direct application of \(\Sigma-\Delta\)-modulators in the implementation of a continuous passivity based controller for the "boost" converter. In sections 5 we present a practical modification of the ideal \(\Sigma-\Delta\)-modulator, which allows a finite commutation frequency to be selected by the user. Section 6 deals with the conclusions of the article.

**II. \(\Sigma-\Delta\)-MODULATORS**

Consider the basic block diagram of Fig. 1, reminiscent of a \(\Sigma-\Delta\)-modulator block but with a binary valued forward nonlinearity, taking values in the discrete set \(\{0,1\}\). For ease of reference we address such a block simply as a \(\Sigma-\Delta\)-Modulator. The following theorem summarizes the relation of the considered modulator with sliding mode control while establishing the basic features of its input output performance.

**Theorem 2.1** Consider the \(\Sigma-\Delta\)-modulator of Fig. 1. Given a sufficiently smooth, bounded, signal \(\mu(t)\), then the integral error signal, \(e(t)\), converges to zero in a finite time, \(t_0\), and, moreover, from any arbitrary initial value, \(e(t_0)\), a sliding motion exists on the perfect encoding condition surface, represented by \(e = 0\), for all \(t > t_0\), provided the following encoding condition is satisfied for all \(t\),

\[
0 < \mu(t) < 1
\]

**Proof.** From the figure, the variables in the \(\Sigma-\Delta\)-modulator satisfy the following relations:

\[
e(t) = u(t) - \mu(t)
\]

\[
u = \frac{1}{2}[1 + \text{sign}(e)]
\]

The quantity \(\dot{e}\) is given by

\[
\dot{e} = e\left[\mu - \frac{1}{2}(1 + \text{sign}(e))\right] = -|e|\left[\frac{1}{2}(1 + \text{sign}(e)) - \mu \text{sign}(e)\right]
\]

For \(e > 0\) we have \(\dot{e} = -e(1 - \mu)\), which, according with the assumption in (1) leads to \(\dot{e} < 0\). On the other hand, when \(e < 0\), we have \(\dot{e} = -|e|\mu < 0\). A sliding regime exists then on \(e = 0\) for all time \(t\) after the hitting time \(t_0\) (see [3]). Under ideal sliding, or encoding, conditions, \(e = 0\), \(\dot{e} = 0\), we have that the, so called, equivalent value of the switched output signal, \(u\), denoted by \(u_{eq}(t)\) satisfies \(u_{eq}(t) = \mu(t)\).

An estimate of the hitting time \(t_0\) is obtained by examining the modulator system equations with the worst possible bound for the input signal \(\mu\) in each of the two conditions: \(e > 0\) and \(e < 0\), along with the corresponding value of \(u\). Consider then \(e(t_0) > 0\) at time \(t = 0\). We have for all \(0 < t \leq t_0\),

\[
e(t) = e(0) + \int_{0}^{t} (\mu(\sigma) - u(\sigma))d\sigma
\]

\[
\leq e(0) + \int_{0}^{t} \sup_{t \in [0,t]} \mu(\tau) - 1
\]

\[
< e(0) + t \left[\sup_{t \in [0,t]} \mu(t) - 1\right] (3)
\]
Since \( e(t_0) = 0 \), we have:

\[
    t_h \leq \frac{e(0)}{1 - \sup_t \mu(t)} \tag{4}
\]

The average \( \Sigma \Delta \)-modulator output \( u_{eq} \), ideally yields the modulator’s input signal \( \mu(t) \) in an equivalent control sense [3]. The role of the above described \( \Sigma \Delta \)-modulator in sliding mode control schemes, avoiding full state measurements, and using average based controllers will be clear from the developments presented below.

### III. USE OF A \( \Sigma \Delta \)-MODULATOR IN THE SLIDING MODE CONTROL IMPLEMENTATION OF AN AVERAGE FEEDBACK CONTROLLER DESIGN

Suppose we have a smooth nonlinear system of the form \( \dot{x} = f(x) + ug(x) \) with \( u \) being a (continuous) control input signal that, due to some physical limitations, requires to be strictly bounded by the closed interval \([0,1]\). Suppose, furthermore, that we have been able to specify a dynamic output feedback controller of the form \( u = -\kappa(y, \zeta) \), \( \zeta = \varphi(y, \zeta, X) \), with desirable closed loop performance features. Assume, furthermore, that for some reasonable set of initial states of the system (and of the dynamic controller), the values of the generated feedback signal function, \( u(t) \), are uniformly strictly bounded in \([0,1]\).

If an additional implementation requirement entitles now that the control input \( u \) of the system is no longer allowed to continuously take values within the interval \([0,1]\), but that it may only take values in the discrete set \( \{0,1\} \), the natural question is: how can we now implement the previously derived continuous controller, so that we can recover, possibly in an average sense, the desirable features of the derived dynamic output feedback controller design in view of the newly imposed actuator restriction?

The answer is clearly given by the average features of the previously considered \( \Sigma \Delta \)-modulator. Recall, incidentally, that the output signal of such a modulator is restricted to take values, precisely, in the discrete set \( \{0,1\} \). Thus, if the output of the designed continuous controller, call it \( \mu(t) = u_{eq}(t) \), is fed into the proposed \( \Sigma \Delta \)-modulator, the output signal of the modulator reproduces, on the average, the required control signal \( u_{eq} \), but in a switched manner.

We have then the following general result concerning the control of nonlinear systems through sliding modes synthesized on the basis of an average feedback controller and a \( \Sigma \Delta \)-modulator. We only deal with the dynamic output feedback controller case for the stabilization problem around an equilibrium. The result, however, can also be easily extended to be valid for trajectory tracking problems.

**Theorem 3.1.** Consider the following smooth nonlinear single input, \( n \)-dimensional system: \( \dot{x} = f(x) + ug(x) \), with the smooth scalar output map, \( y = h(x) \). Assume the dynamic smooth output feedback controller \( u = -\kappa(y, \zeta) \), \( \zeta = \varphi(y, \zeta, X) \), with \( \zeta \in \mathbb{R}^n \), locally (globally, semiglobally) asymptotically stabilizes the system towards a desired constant equilibrium state, denoted by \( X \). Assume, furthermore, that the control signal, \( u \), is uniformly strictly bounded by the closed interval \([0,1]\) of the real line. Then the closed loop system:

\[
    \dot{x} = f(x) + ug(x)  \\
    y = h(x)  \\
    u_{eq}(y, \zeta) = -\kappa(y, \zeta, X)  \\
    \zeta = \varphi(y, \zeta, X)  \\
    u = \frac{1}{2} [1 + \text{sign} \ e]  \\
    \dot{e} = u_{eq}(y, \zeta) - u
\]

exhibits an ideal sliding dynamics which is locally (globally, semiglobally) asymptotically stable to the same constant state equilibrium point, \( X \), of the system.

**Proof.** The proof of this theorem is immediate upon realizing that under the hypothesis on the average control input, \( u_{eq} \), the previous theorem establishes that a sliding regime exists on the manifold \( e = 0 \). Under the invariance conditions, \( e = 0 \), \( \dot{e} = 0 \), which characterize ideal sliding motions (See Sira-Ramírez [9]), the corresponding equivalent control, \( u_{eq} \), associated with the system satisfies: \( u_{eq}(t) = u_{eq}(t) \). The ideal sliding dynamics is then represented by

\[
    \dot{x} = f(x) + u_{eq} g(x)  \\
    y = h(x)  \\
    u_{eq}(y, \zeta) = -\kappa(y, \zeta, X)  \\
    \zeta = \varphi(y, \zeta, X)
\]

which is assumed to exhibit the desired constant state \( X \) as a locally (globally, semiglobally) asymptotically stable equilibrium point.

**Remark 3.2.** Note that the \( \Sigma \Delta \) modulator state, \( e \), can be initialized at the value \( e(t_0) = 0 \). This implies that the induced sliding regime exists uniformly for all times after \( t_0 \). Hence, no reaching time of the sliding surface, \( e = 0 \), is required. This practical feature is adopted throughout.
IV. CONTROL OF THE “BOOST” CONVERTER CIRCUIT

4.1 The “boost” converter model, its average model and a passivity-based control via energy modification and damping injection

Consider the “boost” converter circuit, shown in Fig. 2. The system is described by the set of equations

\[
\begin{align*}
L \frac{di}{dt} &= -(1-u)v + E \\
C \frac{dv}{dt} &= (1-u)i - \frac{v}{R}
\end{align*}
\]

where \(i\) represents the inductor current and \(v\) is the output capacitor voltage. The control input \(u\), representing the switch position function, is a discrete-valued signal taking values in the set \(\{0, 1\}\). The system parameters are constituted by: \(L\), which is the inductance of the input circuit; \(C\), the capacitance of the output filter and \(R\), the output load resistance. The external voltage source has the constant value \(E\). We assume that the circuit is in continuous conduction mode, i.e., the average value of the inductor current never drops to zero, due to load variations.

The state normalization and time scale transformation:

\[
x_1 = \frac{i}{E \sqrt{L/C}}, \quad x_2 = \frac{v}{E}, \quad \tau = \frac{t}{\sqrt{LC}}
\]

yields the following normalized model:

\[
\begin{align*}
\dot{x}_1 &= -(1-u)x_2 + 1 \\
\dot{x}_2 &= (1-u)x_1 - \frac{x_2}{Q}
\end{align*}
\]

where the “ \(\cdot\)” represents derivation with respect to the normalized time, \(\tau\). The circuit “quality” parameter, denoted by \(Q\), is given by the strictly positive quantity, \(R \sqrt{C/L}\). The variable \(x_1\) is the normalized inductor current, \(x_2\) is the normalized output voltage and \(u\) represents the switch position function.

In order to obtain a suitable average controller, assume for a moment that the normalized “boost” converter Eqs. (7) represent a continuous system (i.e., an average system) where \(u\) may take values in the closed interval \([0, 1]\) of the real line.

When we consider \(y = x_1\) as the normalized system output, such an output is a minimum phase output, whereas when the output is the capacitor voltage \(x_2\) such an output is a non-minimum phase output (see [10]).

Given a constant average control input \(u = U \in [0, 1]\), it follows from the average model Eqs. (7) that the corresponding equilibrium values for the average input current, \(\bar{x}_1\), and average output voltage, \(\bar{x}_2\), are

\[
\bar{x}_1 = \frac{1}{(1-U)^2 Q}, \quad \bar{x}_2 = \frac{1}{(1-U)}
\]

using these expressions, we can write

\[
\bar{x}_1 = \frac{1}{Q} \bar{x}_2^2
\]

Since the voltage \(x_2\) is a non-minimum phase output, it is preferable to indirectly control the capacitor voltage through the inductor current \(x_1\). Using (9), we can find the desired steady state of the current in order to reach the desired steady state of voltage.

Let the normalized stored energy function of the system, and its gradient vector, be given by:

\[
H = \frac{1}{2}(x_1^2 + x_2^2) \Rightarrow \frac{\partial H}{\partial x} = \begin{pmatrix} x_1 \\ x_2 \end{pmatrix}
\]

Consider then the following Generalized Hamiltonian Canonical representation of the “boost” converter (see Sira-Ramírez [11]):

\[
\dot{x} = J(u) \frac{\partial H}{\partial x} + R \frac{\partial H}{\partial x} + \varepsilon
\]

\[
y = h(x) = x_1 = e^x \frac{\partial H}{\partial x}
\]

where

\[
J(u) + J^T(u) = 0
\]

\[
R = R^T \leq 0
\]

Indeed, the system (7) may be written in the following form:

\[
\begin{pmatrix}
\dot{x}_1 \\
\dot{x}_2
\end{pmatrix} = \begin{pmatrix}
0 & -(1-u) \\
(1-u) & 0
\end{pmatrix} \begin{pmatrix}
x_1 \\
x_2
\end{pmatrix} + \begin{pmatrix}
0 & 0 \\
0 & -1/Q
\end{pmatrix} \begin{pmatrix}
x_1 \\
x_2
\end{pmatrix} + \begin{pmatrix}
0 \\
1
\end{pmatrix}
\]

We now design a dynamic output feedback controller via the Energy Shaping and Damping Injection method (see...
H. Sira-Ramírez and R. Silva-Ortigoza: Sliding Mode \( \Sigma-\Delta \) Modulation Control of the Boost Converter

We propose the following tracking error energy function with respect to a set of desired states \( x_d \) generated by an exogenous system which is a copy of the system including some suitable damping. We have,

\[
H_d(x-x_d) = \frac{1}{2}(x-x_d)^T (x-x_d)
\]

The function \( H_d(x-x_d) \) exhibits the following time derivative:

\[
\dot{H}_d(x-x_d) = \frac{\partial H(x-x_d)}{\partial (x-x_d)} (\dot{x} - \dot{x}_d)
\]

(11)

with:

\[
\dot{x}_d = J(u) \frac{\partial H(x_d)}{\partial x_d} + R \frac{\partial H(x_d)}{\partial x_d} + e - R_d \frac{\partial H(x-x_d)}{\partial (x-x_d)}
\]

where, the term: \( R_d \frac{\partial H(x-x_d)}{\partial (x-x_d)} \) represents the damping injection term, and \( R_d \) is a negative semidefinite matrix.

The exogenous system corresponding to the desired state variables \( x_d \) is thus given by

\[
\begin{bmatrix}
\dot{x}_{1d} \\
\dot{x}_{2d}
\end{bmatrix} = \begin{bmatrix}
0 & -(1-u) \\
(1-u) & 0
\end{bmatrix} \begin{bmatrix}
x_{1d} \\
x_{2d}
\end{bmatrix} + \begin{bmatrix}
0 \\
0 - \frac{1}{Q}
\end{bmatrix} \begin{bmatrix}
x_{1d} \\
x_{2d}
\end{bmatrix} + \begin{bmatrix}
1 \\
0
\end{bmatrix} \begin{bmatrix}
x_1 - x_{1d} \\
x_2 - x_{2d}
\end{bmatrix}
\]

(12)

Defining: \( e = x-x_d \), we have:

\[
\dot{e} = J(u) \frac{\partial H(e)}{\partial e} + [R + R_d] \frac{\partial H(e)}{\partial e} + e - R_d \frac{\partial H(x-x_d)}{\partial (x-x_d)}
\]

(13)

where \( \xi_2 \), has been used in replacement of the variable \( x_{2d} \).

We thus have:

\[
H_d(e) = -R_d \xi_2^2 - \frac{1}{Q} \xi_2^2 \leq -2 \min \left\{ R, \frac{1}{Q} \right\} \left( \frac{1}{2} e^2 + e_d^2 \right)
\]

\[
= -2 \min \left\{ R, \frac{1}{Q} \right\} H(e) < 0
\]

(14)

It follows that the vector \( x(t) \) exponentially asymptotically converges towards the exogenous (desired) trajectory \( x_d(t) \).

4.2 Passivity-based control via energy modification and damping injection of the “boost” converter implemented through a \( \Sigma-\Delta \)-modulator

We propose to use the average designed passivity-based controller (13) in a sliding mode implementation using a \( \Sigma-\Delta \)-modulator, as discussed in Section 3.

\[
u = \frac{1}{2} [1 + \text{sign} \ e]
\]

\[
\dot{e} = u_{av} - u
\]

\[
u_{av} = -\frac{1}{\xi_2} \left[ 1 + R(x_1 - \bar{x}_1) \right] + 1
\]

(15)

Figure 3 illustrates the indirect output voltage regulation for a typical “boost” converter circuit \( L=20 \times 10^{-3} \) H, \( C=20 \times 10^{-6} \) Farad, \( R=30 \Omega, E=15V \) with a desired steady state voltage of \( \bar{x}_1 = 2 \) (corresponding to \( v=30V \)), and \( \bar{x}_1 = 4.216 \) (corresponding to \( i = 2A \)). These parameter values yield \( Q = 0.9486 \) and the time normalization factor was obtained to be \( \sqrt{LC} = 6.32 \times 10^{-4} \).

Figure 4 depicts some computer simulations portraying the closed loop response of the system when the passivity-based controller is implemented using a \( \Sigma-\Delta \)-modulator. The controller and the system parameters were chosen to be exactly the same as in the previous simulation.

V. A PROPOSED PRACTICAL CIRCUIT \( \Sigma-\Delta \)-MODULATOR IMPLEMENTATION

A practical solution for the implementation of the ideal \( \Sigma-\Delta \)-modulator would not markedly affect the key results. For this, we propose a practical modification of the \( \Sigma-\Delta \)-modulator, which allows a finite frequency to be selected, defined by the user as shown in Fig. 5. It is important to notice that this practical modification arises from the introduction of a classical modulator PWM inside the classical \( \Sigma-\Delta \)-modulator, in order to limit the operation frequency of the converter.
Figure 6 depicts some computer simulations portraying the closed loop response of the system when the passivity-based controller is implemented using the modified Σ-Δ-modulator presented in the Fig. 5. The controller and the system parameters were chosen to be exactly the same as in the previous simulation for the boost converter, for a frequency of 50kHz. Moreover, a closed loop to the commuted control signal and its average is presented in the Fig. 6.

VI. CONCLUSION

Average feedback controller designs usually represent the desirable equivalent control in sliding mode control implementations. The exact synthesis of the equivalent control is not physically possible in systems commanded by switches. Knowledge of the feedback law defining the equivalent control leads to consider a linear partial differential equation, for the sliding surface, stating that the closed loop vector field should be orthogonal to the sliding surface gradient. However, it is still not obvious how to synthesize a sliding surface, that corresponds to a given equivalent control, due to the indeterminacy, and arbitrariness, of the boundary conditions in the defining linear partial differential equation that needs to be solved.

In this article, we have demonstrated that the use of classical Σ-Δ-modulators can solve the sliding mode control implementation problem arising from an average feedback controller design in a rather efficient manner. The proposed approach retains, in an average sense, the desirable features of the designed average feedback controller. When the proposed controllers are synthesized using only inputs and outputs, as in GPI control, or in output feedback control, the explicit asymptotic estimation of the plant’s state becomes unnecessary and, moreover, the matching conditions, intimately related to the state space representation of the system, are no longer needed.

We have used the Σ-Δ-modulator implementation of a sliding mode controller for a given average passivity based controller for the “boost” DC power converter, resulting in a dynamic output feedback controller.

We have also provided a proposal of the practical implementation of the Σ-Δ-modulator, which allows to limit the commutation frequency of the system, generating very similar results to those obtained when we use the classical Σ-Δ-modulator that produces an ideal infinite commutation frequency.

Other interesting non-linear switched controlled systems, such as jet controlled satellites, Un-interruptible Power Supplies, Active Filters, etc., may immediately benefit from the sliding mode feedback controller design framework based on Σ-Δ-modulators and nonlinear (dynamic or static) output feedback controllers. In this respect, the methods and techniques of current nonlinear systems theory (for instance, geometric, differential...
algebraic, flatness, passivity, energy methods, $H_\infty$, etc.) become also readily available for the class of switched controlled systems.

REFERENCES


Fig. 6. Closed loop response of a “boost” power converter to a $\Sigma\Delta$-modulator implementation for a finite frequency of 50kHz.

Hebertt Sira-Ramírez obtained the degree of Electrical Engineer from the Universidad de Los Andes (Mérida-Venezuela), which he joined as an Instructor professor, in 1970. He obtained the Master’s and PhD degrees from the Massachusetts Institute of Technology (Cambridge, USA) in 1974 and 1977, respectively. He is a retired professor from the Universidad de Los Andes where we worked for 28 years and served as: Head of the Department of Control Systems Engineering, Head of the Postgraduate School in Automatic Control and as a Vicerector of the University. Since 1998 he has been with Cinvestav-IPN, a research institution in Mexico City, where he is a Titular Researcher. He is a Level IV Researcher of the Venezuelan PPI organization and a Level III Researcher in the SNI organization in Mexico. Dr. Sira-Ramírez is interested in the theory and applications of discontinuous feedback control and algebraic systems theory for identification, state estimation and fault detection. He is a co-author of the books: Passivity Based Control of Euler-Lagrange Systems (Springer-Verlag, London, 1998), Differentially Flat Systems (Marcel Dekker, Boston, 2004), Algebraic Methods in Flatness, Signal Processing and State Estimation (Lagares, México, 2003) and Control de Sistemas No Lineales (Pearson-Wiley, Spain, 2005). He has served as member of several Editorial Boards in journals and conferences devoted to Automatic Control. He has published over one hundred papers in credited journals, 20 book chapters and nearly 200 articles in international conferences.

Ramón Silva-Ortigoza received the B.S. degree in electronics from Universidad Autónoma de Puebla, Puebla, Mexico, in 1999, and the M.S degree in electrical engineering from CINVESTAV-IPN, Department of Electrical Engineering, Section of Mechatronics, D.F, Mexico, in 2002. He is currently working toward the Ph.D. degree in electrical engineering. M.S. Silva-Ortigoza is interested in the theory and applications of automatic control with emphasis in controller design techniques for Power Electronics devices.