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ROBUST NONLINEAR ADAPTIVE CONTROL OF A DC-DC BOOST CONVERTER WITH UNCERTAIN PARAMETERS

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Abstract. *In this work, a robust adaptive nonlinear controller is designed for a boost converter subject to completely unknown uncertainties about the converter nominal parameters. Adaptation laws are designed using an estimator to estimate these uncertainties. The stability analysis is carried out within the framework of the highly nonlinear nonminimum phase system. Using simulation, the effectiveness of the developed controller is compared with the widely popular proportional integral (PI) and sliding mode cascade controller that has shown excellent features in terms of output tracking abilities, robustness to system uncertainties and disturbances. The robustness of the controller developed in this paper is also validated experimentally and tested against large parameter uncertainties, large input voltage variations and step load changes.*
Keywords: Robust adaptive nonlinear control, DC-DC boost converter, Nonminimum phase system, Pulsewidth modulation

1. **Introduction.** There have been a large number of control strategies that have been developed to regulate the output voltage of the boost converter. They include conventional control methodologies based on the application of linear control theory on linearized small-signal models to various nonlinear control schemes such as backstepping technique [1], input-output feedback linearization technique, flatness-based control [2], sliding mode control [3-9], and passitivity-based controllers [10-13]. In the context of PWM-based control of DC-DC boost converters, conventional control strategies based on linear techniques are still widely used. However, linearization of highly nonlinear nonminimum phase systems and application of linear control techniques may lead to poor performance under different operating points, large disturbances and large parameters uncertainties. Nonlinear adaptive controllers that are developed within the framework of the nonlinear model, will in general provide robustness and satisfactory response under a wide range of operating conditions, parameter uncertainties and large disturbances.

Some nonlinear controller designs require a precise knowledge of the converter parameters. However, their robustness to parameter uncertainties and large disturbances has not been tested. Others, assume only the load and/or the external input voltage to be unknown. More practical and robust controller designs take into account incremental variations of the boost converter parameters from their nominal values. However, their implementations require the exact knowledge of the bounds of the parameter uncertainties. There are many practical situations where perfect knowledge of component values is not available due to many factors such as wide manufacturing tolerances, parameter drifts, ageing, measurements errors, temperature and the effect of coil magnetic saturation. In addition, the external input voltage and the loads are also subject to large uncertainties.

In this work, we consider the design of a PWM-based robust nonlinear adaptive controller for the DC-DC boost converter subject to completely unknown parameter uncertainty bounds. The adaptive controller is designed to always guarantee a tight asymptotic regulation of the output voltage regardless of system uncertainties and large external input and load variations. The stability of the closed-loop system is carried out using the highly nonlinear and nonminimum phase model of the boost converter as opposed to the customary linearization techniques that guarantee only local stability. To the best of this author's knowledge, the design of such a robust controller under unknown parametric uncertainties and large disturbances coupled with a rigorous stability analysis carried out within the framework of the nonlinear nonminimun phase model has not been addressed in the control literature.

The proportional integral (PI) in combination with sliding mode control (SMC) presented in [8] is widely applied to switching converters because of its excellent tight output regulation and robustness to parametric uncertainties and large disturbances, and therefore is a very good candidate to perform a comparative simulation with the proposed controller. Finally, we also present experimental results that show that the proposed controller forces the output voltage to tightly track the reference voltage despite large parameter uncertainties, large external input voltage and load step variations. The design method of the robust adaptive controller can be easily extended to buck and buck-boost converters.

The structure of this paper is as follows. In Section 2, adaptation laws for the uncertain parameters and the adaptive controller are designed. In Section 3, the closed-loop stability of the converter is analyzed. Section 4 provides comparative simulation results, while Section 5 provides the plots of the experimental results. Section 6 is the conclusion.

2. Preliminaries and Robust Adaptive Controller Design.

2.1. DC-DC boost converter model. A basic boost converter is shown in Figure 1. Under continuous conduction mode (CCM), the averaged model of the converter is

$$\begin{aligned} \dot{x}_1 &= -(1-u)(a+\Delta a)x_2 + (b+\Delta b) \\ \dot{x}_2 &= (1-u)(c+\Delta c)x_1 - (d+\Delta d)x_2 \end{aligned} \tag{1}$$

where the state variable x_1 represents the average inductor current i_L, x_2 the average capacitor voltage v_o. The parameters a, b, c, d represent the nominal parameters given by

$$a = \frac{1}{L_N}, \quad b = \frac{E_N}{L_N}, \quad c = \frac{1}{C_N}, \quad d = \frac{1}{R_N C_N} \tag{2}$$

and the components L_N, C_N, and R_N represent the known nominal values of the inductor, capacitor and load resistor respectively. The parameter E_N represents the nominal value of the external input voltage. The parameters Δa, Δb, Δc and Δd represent the completely

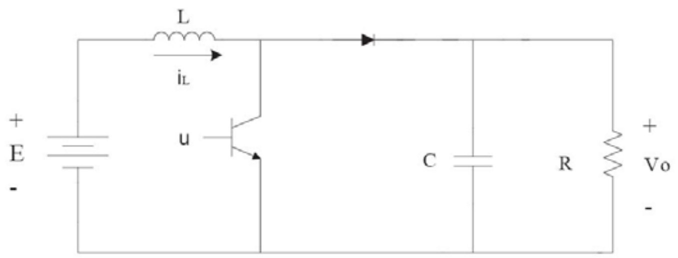

Figure 1. A boost converter

unknown incremental variations of a, b, c and d, respectively. The control input u to the converter is the duty ratio function.

2.2. **Estimator-based adaptation laws.** In this work, both x_1 and x_2 are assumed to be accessible. Nevertheless, an estimator is used to facilitate the design of the parameter adaptation laws for the DC-DC boost converter uncertain parameters. Moreover, it will be seen in the sequel, that a proper choice of the estimator initial condition will always guarantee the existence of a robust adaptive controller that yields a very tight asymptotic regulation of the output voltage despite parameter uncertainties, large external input and load variations. To this end, we consider the following estimator

$$\begin{aligned} \dot{\hat{x}}_1 &= -(1-u)a\hat{x}_2 - (1-u)\widehat{\Delta a}x_2 + b + \widehat{\Delta b} + K_1\tilde{x}_1 \\ \dot{\hat{x}}_2 &= (1-u)c\hat{x}_1 + (1-u)\widehat{\Delta c}\, x_1 - \left(d + \widehat{\Delta d}\right) x_2 + K_2\tilde{x}_2 \end{aligned} \tag{3}$$

where $K_1 > 0$ and $K_2 > 0$ are the estimator gains, $\hat{x}_1$ and $\hat{x}_2$ are the estimates of x_1 and x_2 respectively, $\tilde{x}_1 = x_1 - \hat{x}_1$ and $\tilde{x}_2 = x_2 - \hat{x}_2$. The parameters $\widehat{\Delta a}$, $\widehat{\Delta b}$, $\widehat{\Delta c}$ and $\widehat{\Delta d}$ are the estimates of Δa, Δb, Δc and Δd respectively. Using (1) and (3) we obtain the following error equations

$$\begin{aligned} \dot{\tilde{x}}_1 &= -(1-u)a\tilde{x}_2 - (1-u)\widetilde{\Delta a}\, x_2 + \widetilde{\Delta b} - K_1\tilde{x}_1 \\ \dot{\tilde{x}}_2 &= (1-u)c\tilde{x}_1 + (1-u)\widetilde{\Delta c}\, x_1 - \widetilde{\Delta d}x_2 - K_2\tilde{x}_2 \end{aligned} \tag{4}$$

where $\widetilde{\Delta a} = \Delta a - \widehat{\Delta a}$, $\widetilde{\Delta b} = \Delta b - \widehat{\Delta b}$, $\widetilde{\Delta c} = \Delta c - \widehat{\Delta c}$ and $\widetilde{\Delta d} = \Delta d - \widehat{\Delta d}$.

To generate the adaptation laws we consider the following quadratic Lyapunov function.

$$\begin{aligned} V = {} & \frac{1}{2a}\tilde{x}_1^2 + \frac{1}{2c}\tilde{x}_2^2 + \frac{1}{2a\gamma_1}\widetilde{\Delta a}^2 + \frac{1}{2a\gamma_2}\widetilde{\Delta b}^2 \\ & + \frac{1}{2c\gamma_3}\widetilde{\Delta c}^2 + \frac{1}{2c\gamma_4}\widetilde{\Delta d}^2 \end{aligned} \tag{5}$$

where $\gamma_1 > 0$, $\gamma_2 > 0$, $\gamma_3 > 0$ and $\gamma_4 > 0$ are design parameters. Its time derivative along the trajectories of (4) is

$$\begin{aligned} \dot{V} = {} & -\frac{K_1}{a}\tilde{x}_1^2 - \frac{K_2}{c}\tilde{x}_2^2 - \frac{\widetilde{\Delta a}}{a}\left[(1-u)x_2\tilde{x}_1 + \frac{1}{\gamma_1}\dot{\widehat{\Delta a}}\right] \\ & + \frac{\widetilde{\Delta b}}{a}\left[\tilde{x}_1 - \frac{1}{\gamma_2}\dot{\widehat{\Delta b}}\right] + \frac{\widetilde{\Delta c}}{c}\left[(1-u)x_1\tilde{x}_2 - \frac{1}{\gamma_3}\dot{\widehat{\Delta c}}\right] \\ & - \frac{\widetilde{\Delta d}}{c}\left[x_2\tilde{x}_2 + \frac{1}{\gamma_4}\dot{\widehat{\Delta d}}\right] \end{aligned} \tag{6}$$

The adaptation laws are determined by cancelling the terms in brackets and are given by the following expressions

$$\begin{aligned} \dot{\widehat{\Delta a}} &= -\gamma_1\ (1-u)\ x_2\tilde{x}_1 \\ \dot{\widehat{\Delta b}} &= \gamma_2\ \tilde{x}_1 \\ \dot{\widehat{\Delta c}} &= \gamma_3\ (1-u)\ x_1\tilde{x}_2 \\ \dot{\widehat{\Delta d}} &= -\gamma_4\ x_2\tilde{x}_2 \end{aligned} \tag{7}$$

With the adaptation laws given by (7), we now have

$$\dot{V} = -\frac{K_1}{a}\tilde{x}_1^2 - \frac{K_2}{c}\tilde{x}_2^2 \tag{8}$$

Using Lasalle invariance principle, we can conclude that $\tilde{x}_1 \to 0,\ \tilde{x}_2 \to 0$ asymptotically. Consequently, the estimates given by (7), converge asymptotically and respectively to the constants $\widehat{\Delta a}_{ss}$, $\widehat{\Delta b}_{ss}$, $\widehat{\Delta c}_{ss}$ and $\widehat{\Delta d}_{ss}$.

2.3. **Robust nonlinear adaptive controller design.** We consider the following tracking error

$$\sigma = \hat{x}_1 + \gamma \int_0^t (\hat{x}_2 - V_{ref})\, d\tau \tag{9}$$

where $\gamma > 0$ is a design parameter and V_{ref} is the desired output voltage.

The controller u is derived by differentiating σ with respect to time and setting $\dot{\sigma} = 0$. Using (3) and (9), the controller is

$$u = 1 - \frac{\left(b + \widehat{\Delta b} + K_1 \tilde{x}_1 + \gamma\,(\hat{x}_2 - V_{ref})\right)}{a\hat{x}_2 + \widehat{\Delta a}\ x_2} \tag{10}$$

where $\hat{x}_2(0) > 0$.

The initial condition $\hat{x}_1(0)$ of the estimator is selected as $\hat{x}_1(0) = 0$, then from (9) we have $\sigma(0) = 0$ and since the controller guarantees

$$\dot{\sigma}(t) = 0,\ t \geq 0 \tag{11}$$

then $\sigma(t) = 0$ for all $t \geq 0$. In this case we have

$$\lim_{t\to\infty} x_1(t) = \hat{x}_1(t)|_{t\geq 0} = -\gamma \int_0^t (\hat{x}_2 - V_{ref}) d\tau \tag{12}$$

3. **Stability Analysis.** The evolution of the state estimate $\hat{x}_2$ is determined by substituting the controller (10), the integration term in (12) for $\hat{x}_1$ into the second equation of (3). We also use $x_1 = \hat{x}_1 + \tilde{x}_1$ and $x_2 = \hat{x}_2 + \tilde{x}_2$ to obtain the following integro-differential equation

$$\begin{aligned} \dot{\hat{x}}_2 = {} & \frac{\left(b + \widehat{\Delta b} + K_1\tilde{x}_1 + \gamma\,[\hat{x}_2 - V_{ref}]\right)}{a\hat{x}_2 + \widehat{\Delta a}\ \ (\hat{x}_2 + \tilde{x}_2)} \\ & \times \left[-\gamma\left(c + \widehat{\Delta c}\right)\int_0^t (\hat{x}_2 - V_{ref})\, d\tau + \widehat{\Delta c}\ \tilde{x}_1\right] \\ & - \left(d + \widehat{\Delta d}\right)(\hat{x}_2 + \tilde{x}_2) + K_2\tilde{x}_2 \end{aligned} \tag{13}$$

As $t \to \infty$, we have $\tilde{x}_1 \to 0$, $\tilde{x}_2 \to 0$ and from (7), $\widehat{\Delta a} \to \widehat{\Delta a}_{ss}$, $\widehat{\Delta b} \to \widehat{\Delta b}_{ss}$, $\widehat{\Delta c} \to \widehat{\Delta c}_{ss}$ and $\widehat{\Delta d} \to \widehat{\Delta d}_{ss}$ asymptotically, and then the above integro-differential equation admits as a limit integro-differential equation

$$\dot{\hat{x}}_2 = -\frac{(b^* + c^*(\hat{x}_2 - V_{ref}))\int_0^t (\hat{x}_2 - V_{ref})\, d\tau}{a^*\hat{x}_2} - d^*\hat{x}_2 \tag{14}$$

where

$$\begin{aligned} b^* &= \gamma\left(b + \widehat{\Delta b}_{ss}\right)\left(c + \widehat{\Delta c}_{ss}\right), \quad c^* = \gamma^2\left(c + \widehat{\Delta c}_{ss}\right) \\ a^* &= \left(a + \widehat{\Delta a}_{ss}\right), \quad d^* = \left(d + \widehat{\Delta d}_{ss}\right) \end{aligned} \tag{15}$$

It was found through extensive simulations that the selection of high estimator gains K_1 and K_2 and adaptation gains γ_1, γ_2, γ_3 and γ_4 to be much greater than $1/R_N C_N$ and less than the switching frequency, will ensure that the dynamics of the estimator are made much faster than the dynamics of the closed-loop system. Here, the parameter $R_N C_N$

represents the nominal time constant of the open-loop system. In our simulations and experimental results, the estimator and adaptation gains were chosen to be greater than $\frac{5}{R_N C_N}$. In this case the evolution of $\hat{x}_2$ given by (13) can be accurately described by the evolution of $\hat{x}_2$ given by the limit Equation (14). Also, the asymptotic behavior of the trajectories of (13) converges towards the equilibrium of (14), which will be shown to be an asymptotically stable equilibrium. To this end we define the state variables y_1 and y_2 as

$$y_1 = \int_0^t (\hat{x}_2 - V_{ref})\, d\tau, \quad y_2 = \hat{x}_2 - V_{ref} \tag{16}$$

Using (16) into (14) we obtain

$$\begin{aligned} \dot{y}_1 &= y_2 \\ \dot{y}_2 &= -\frac{(b^* + c^* y_2) y_1}{a^*(y_2 + V_{ref})} - d^*(y_2 + V_{ref}) \end{aligned} \tag{17}$$

The system (17) admits the equilibrium (y_1^*, y_2^*) given by

$$y_1^* = -\frac{a^* d^* V_{ref}^2}{b^*}, \quad y_2^* = 0 \tag{18}$$

To determine the region of attraction of the equilibrium point we consider the Lyapunov function

$$V(y) = \frac{1}{2}\alpha\, (y_1 - y_1^*)^2 + \frac{1}{2}\beta y_2^2 \tag{19}$$

where $y = (y_1, y_2)^T$, and $\alpha > 0$ and $\beta > 0$ are parameters to be determined. Its derivative along the trajectories of (17) is

$$\begin{aligned} \dot{V} = {} & \alpha(y_1 - y_1^*) y_2 - \frac{\beta(b^* y_1 + c^* y_1 y_2) y_2}{a^*(y_2 + V_{ref})} \\ & -\beta d^* y_2^2 - d^* \beta V_{ref} y_2 \end{aligned} \tag{20}$$

After reduction to the same denominator and basic algebraic manipulations and rearrangement, we obtain

$$\begin{aligned} \dot{V}(y) = {} & \frac{-y_2^2\,[-(a^*\alpha - \beta c^*) y_1 + a^*\alpha y_1^* + a^*\beta d^* V_{ref}]}{a^*(y_2 + V_{ref})} \\ & + \frac{(a^*\alpha V_{ref} - \beta b^*) y_1 y_2 - (a^*\alpha y_1^* + a^*\beta d^* V_{ref}) V_{ref}\; y_2}{a^*(y_2 + V_{ref})} \\ & -\beta d^* y_2^2 \end{aligned} \tag{21}$$

We choose the parameters $\alpha > 0$ and $\beta > 0$ to satisfy

$$a^* \alpha V_{ref} = \beta b^* \tag{22}$$

Using (18) and (22) we also have the following equality

$$a^* \alpha y_1^* = -a^* d^* \beta V_{ref} \tag{23}$$

In view of (22) and (23), Equation (21) reduces to

$$\dot{V}(y) = \frac{-y_2^2\,[-(a^*\alpha - \beta c^*) y_1]}{a^*(y_2 + V_{ref})} - \beta d^* y_2^2 \tag{24}$$

With practical boost component values and reference voltages, we generally have $a^*\alpha \gg \beta c^*$. Using (15) and (22) this inequality reduces to $(b + \widehat{\Delta b}_{ss}) \gg \gamma V_{ref}$. Also, $\gamma > 0$ is a design parameter that can be chosen such that the inequality condition is satisfied. We should also mention that $y_2 + V_{ref} = \hat{x}_2 > 0$ for all $t \geq 0$.

If

$$y \in \Omega = \{y|\ y_1 \le 0\} \tag{25}$$

then from (24) we have

$$\dot{V}(y) \le -\beta d^* y_2^2 \le 0 \tag{26}$$

which is negative semi definite. Using LaSalle's invariance theorem, we conclude that $y_2 \to 0$ and therefore $\hat{x}_2 \to V_{ref}$ and $x_2 \to V_{ref}$. Also, substituting $y_2 = 0$ and $\dot{y}_2 = 0$ in (17) yields $y_1 = y_1^*$. Therefore, for any $y = (y_1, y_2)^T$ in the set

$$\Omega = \{y|\ y_1 \le 0\} \tag{27}$$

the equilibrium is locally asymptotically stable.

Next we will determine a region of attraction that will always include our initial conditions and ensures that $x_2 \to V_{ref}$. Let

$$R = \left\{ y|V(y) \le \frac{1}{2}\alpha y_1^{*2} \right\} \tag{28}$$

If $y \in R$ then we must also have $y \in \Omega$ since

$$\frac{1}{2}\alpha(y_1 - y_1^*)^2 + \frac{1}{2}\beta y_2^2 \le \frac{1}{2}\alpha y_1^{*2} \tag{29}$$

implies that $y_1 \le 0$. Please note here that $y_1^* < 0$. Therefore,

$$R \subset \Omega \tag{30}$$

and R is an invariant set that includes the equilibrium point given by (18). Any trajectory starting in R at $t = 0$ will remain in $R \subset \Omega$ for all $t \ge 0$. This is true since in Ω, inequality (26) holds and we have

$$\dot{V}(y(t)) \le 0 \Longrightarrow V(y(t)) \le V(y(0)) \le \frac{1}{2}\alpha y_1^{*2} \tag{31}$$

If we choose the initial condition $\hat{x}_2(0) = V_{ref}$ then from (16) we have $y_2(0) = 0$ and $y_1(0) = 0$ and from (19) $V(y(0)) = \frac{1}{2}\alpha y_1^{*2}$. This implies that $y(0) \in R$ and $y(t) \in R \subset \Omega$ for all $t \ge 0$. In view of (25) and (26), the application of LaSalle's invariance theorem yields $y_2 \to 0$ and consequently $\hat{x}_2 \to 0$ and $x_2 \to V_{ref}$. The choice of $\hat{x}_2(0) = V_{ref}$ will guarantee that the output voltage will always converge to V_{ref}.

4. **Simulation Results.** A Matlab/Simulink simulation is performed to compare the proportional integral (PI) and sliding mode cascade controller developed in [8] with the proposed robust adaptive controller. The nominal parameters of the boost converter are selected as in [8] and are $L_N = 40$mH, $C_N = 4\mu$F, $R_N = 40\Omega$, $E_N = 20$V and $V_{ref} = 50$V. The controller developed in [8] is given by

$$u = 0.5(1 - sign(S)) \tag{32}$$

where the switching manifold S is given by

$$S = x_1 - \frac{V_{ref}^2}{R_N E_N} - K_p e - K_i \int_0^t e(\tau)d\tau \tag{33}$$

with $e = V_{ref} - x_2$. The gains K_p and K_i are obtained through the linearization of the system around the equilibrium and are given [8] as $K_p = -0.0087$ and $K_i = 10.3347$. For the proposed controller, the estimator gains K_1 and K_2 and the adaptation gains γ_1, γ_2, γ_3 and γ_4 are chosen as $5/R_N C_N = 31,250$ with $\gamma = 10$.

To model system uncertainties, we consider the actual parameters of the boost converter, which are unknown to both controllers, to be $E = 75\% E_N = 15$V, $L = 50\% L_N =$

20mH, $C = 500\%C_N = 20\mu$F, and $R = 300\%R_N = 120\Omega$ with $V_{ref} = 35$V. In addition, we assume that the load resistance R varies stepwise from $R = 120\Omega$ to $R = 240\Omega$ at $t = 100$ms and then back to $R = 120\Omega$ at $t = 200$ms. The input voltage E is changed stepwise from 15V to 20V at $t = 300$ms and then back to 15V at $t = 400$ms. Finally, the reference voltage is changed from 35V to 50V at $t = 500$ms.

The performance index used in the comparaison of the two controllers is the integral of the absolute magnitude of the error, IAE, which is defined as in [14] and given by

$$IAE = \int_0^T |e(t)|dt \tag{34}$$

The plots of the output voltage are shown in Figure 2 for the boost converter subject to the controller given in [8] and Figure 3 for the proposed controller.

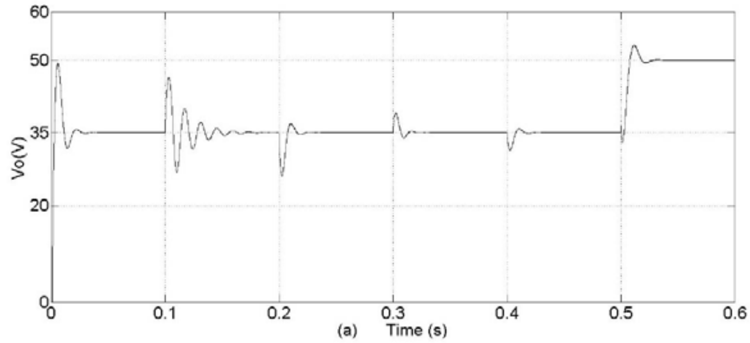

FIGURE 2. Transient output response of the boost converter subject to the controller of [8]

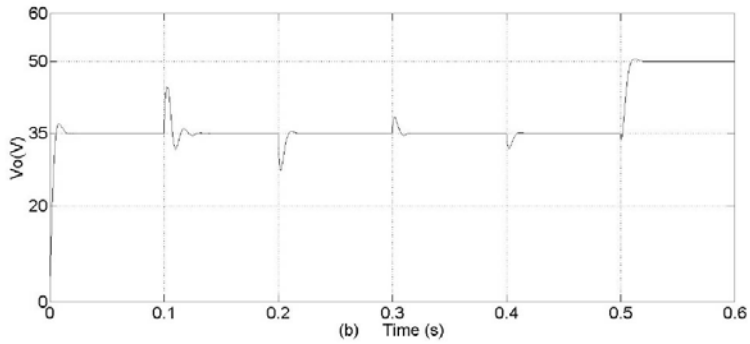

FIGURE 3. Transient output response of the boost converter subject to the proposed controller

As seen from the figures, there is a substantial percent overshoot of 41% at start-up and 6.10% during voltage reference change using the controller of [8] and 5.7% at start-up and 1.14% during voltage reference change using the proposed controller. In addition, the output voltage transient exhibits an oscillatory response during the load step change when using the controller of [8]. Finally, the performance of the controlled system as measured by the output voltage tracking error index IAE is 0.52 with the controller of [8] and 0.30 with the proposed controller.

During start-up, and using the controller given in [8], the percent overshoot of the output voltage as well as the peak of the inductor current are heavily influenced by system uncertainties since they affect the duration of the reaching phase during which the controlled system is sensitive to parameter uncertainties and disturbances. In particular, we should note that the reaching phase is substantially increased with an increase of the value of the actual inductance, and as a result a substantial degradation of the performance of the system response was observed.

Linearization of highly nonlinear nonminimum phase systems, as it is done in [8], and application of linear control techniques may lead to poor performance under different operating points, large disturbances and large parameters uncertainties. Adaptive nonlinear controllers that are developed within the framework of the nonlinear model, as the case of the proposed controller, will in general provide robustness and satisfactory response under a wide range of operating conditions, parameter uncertainties and large disturbances.

5. **Experimental Results.** The boost converter nominal parameters used in the implementation of the robust nonlinear adaptive controller are $L_N = 300\mu$H, $C_N = 280\mu$F, $R_N = 100\Omega$, and $E_N = 8$V.

The actual boost converter parameters used in the experiment and assumed completely unknown in the implementation of the adaptive controller are $L = 180\mu$H, $C = 150\mu$F, $R = 40\Omega$, and $E = 6$V. The switching frequency of the PWM modulator is 200kHz. A prototype of the boost converter was constructed and the adaptive nonlinear controller was implemented using the dSpace 1104 controller board.

The estimator gains and the adaptation gains are chosen to be $100/R_NC_N$. In this case we have $K_1 = K_2 = 3571$ and $\gamma_1 = \gamma_2 = \gamma_3 = \gamma_4 = 3571$. Using $\gamma = 250$ with the estimator initial conditions as $\hat{x}_1(0) = 0$ and $\hat{x}_2(0) = 12$V, the experimental results are depicted in Figures 4, 5, 6 and 7.

Figure 4 shows the response due to a periodic step change of the reference V_{ref} from 8V to 12V and from 12V to 8V. Figure 5 depicts the robustness of the adaptive controller to step load variation. In our case, the load resistor R varies stepwise periodically between 40Ω and 160Ω with $V_{ref} = 12$V. Shown in Figure 6 is the robustness of the controller to the external input voltage change. In this case E undergoes a variation of 67% from its nominal value of 6V to 10V. Finally, shown in Figure 7 is the output voltage when the external input voltage undergoes a variation of -40% from 10V to 6V. These experimental results show the excellent features in terms of robustness and recovery of the proposed adaptive controller to large and unknown parameter uncertainties, large external input voltage variations and load step variations.

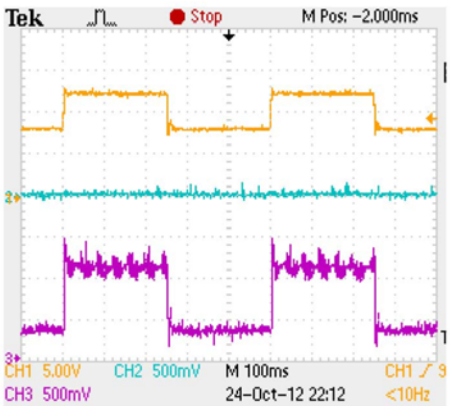

FIGURE 4. Transient response with reference voltage change, V_{ref} varying periodically stepwise between 8V and 12V. Top: Output voltage V_o (5V/div). Middle: Tracking error σ (500 mV/div). Bottom: Inductor current i_L (0.5A/div).

FIGURE 5. Transient response with the load resistor R_o varying periodically stepwise between 40Ω and 160Ω. Top: Output voltage V_o (5V/div). Middle: Tracking error σ (500mV/div). Bottom: Inductor current i_L (0.5A/div).

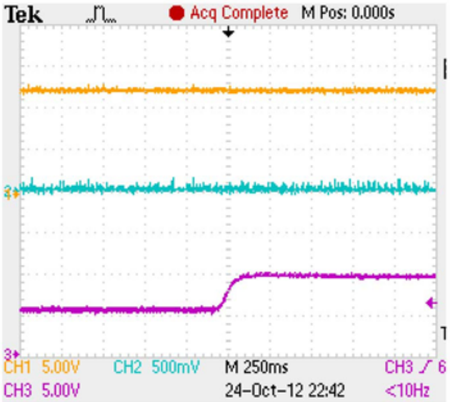

FIGURE 6. Transient response with the input voltage E undergoing a variation of 67% from its nominal value of 6V to 10V. Top: Output voltage V_o (5V/div). Middle: Tracking error σ (500mV/div). Bottom: Input voltage E (5V/div).

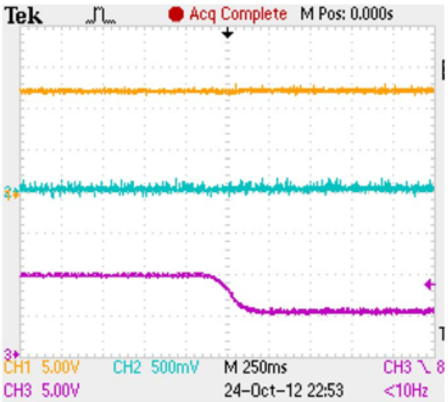

FIGURE 7. Transient response with the input voltage E undergoing a variation of −40% from 10V to 6V. Top: Output voltage V_o (5V/div). Middle: Tracking error σ (500mV/div). Bottom: Input voltage E (5V/div).

6. **Conclusion.** Estimator-based robust adaptive controllers are designed for the boost converter to deal with the unknown incremental variations of the converter parameters from their nominal values. Through experimental results, the controller designed yields a tight output regulation, is robust to uncertain parameters and has excellent recovery features to large input voltage variations and large step load changes.

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