# Fast Adaptive Finite-Time Voltage Regulation Control Algorithm for a Buck Converter System

Yingying Cheng, Haibo Du, Chen Yang, Zuo Wang, Jinping Wang, and Yigang He

***Abstract*—A new voltage regulation control algorithm is designed for the buck-type dc-dc converter system in the presence of unknown input voltage and load variations. To enhance the voltage regulation time by using the finite-time control theory, a new fast voltage regulation control algorithm is designed that can guarantee that the output voltage converges to the reference voltage in a finite time. To address the unknown input voltage and load variations, two finite-time convergent observers are designed to estimate the unknown parameters in a finite time. Finally, an adaptive finite-time control algorithm is developed. Compared with the PI control algorithm, experimental results show that the proposed algorithm has a faster regulation performance and stronger performance on load-variation rejection.**

***Index Terms*—Adaptive control, buck converter, finite-time control, finite-time estimator.**

## I. Introduction

DC–DC converters are important power electronic devices which have been widely used in uninterruptible power supplies, dc motor drives and other devices and applications that require optimal efficient power conversion. Usually, the dc–dc converter can be classified into three types, i.e., buck type, boost type, and buck–boost type [1]. With the development of modern industrial technology, the performance requirement of dc power is increasing, e.g., high-quality, lightweight, reliable, adaptive, and efficient power supplies. Usually, there are two main issues in the study of dc–dc converter, i.e., topological structure and control strategy. The main aim of this brief is to design a new control algorithm to improve the system performance.

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From the control viewpoint, the dc–dc converter is a typical switching system and many control algorithms are based on the linear control theory [2]–[5]. To further improve the control performance, many different nonlinear control tools were employed to design nonlinear control algorithms. In [6]–[8], the sliding mode control method was employed to design variable structure control algorithms. In [9], the fuzzy control method was used to design control system for dc–dc converter. In [10], the Lyapunov function-based-design method was used to design a saturation control algorithm. Based on the sampled-data output feedback theory, an output-based nonlinear control algorithm was proposed for a buck-type dc–dc generalized proportional-integral observer (GPIO) converter [11]. Based on the extended state observer and the sliding mode control method, a disturbance compensation-based sliding mode control law was designed in [12] to improve the ability of disturbance rejection. In [13], the GPIO-based composite control approach was employed to design a controller for the dc–dc buck converter.

Note that for most of existing feedback control algorithms, the closed-loop system for dc–dc converter is asymptotically stable, which means that the output voltage converge to the reference voltage asymptotically. For dc–dc converter, it is very important to improve the system's transient response. To this end, this brief employs a new developed nonlinear control method (called finite-time control method which guarantees that the system states converge to the equilibrium in a finite time [14], [15]) to design voltage regulation algorithm. That is to say that the output voltage reaches the reference voltage in a finite time. Since dc–dc converters are subject to many kinds of disturbances such as parameter uncertainties, load variations, and input voltage variations, how to enhance the ability of disturbance rejection of closed-loop system is also an important issue in the literature. Besides the fast convergence, the finite-time control usually offer strong disturbance rejection properties [14], [16].

As of the many advantages finite-time control offers, this brief employs it to design the control system for a buck-type dc–dc converter. To achieve the objective that the output voltage converge to the reference voltage in a finite time, the design procedure is divided into three steps. At the first step, a time-scale transformation is used for the average state space model of the buck converter system. Then, based on the saturation finite-time control theory, a new fast voltage regulation control algorithm is designed which can guarantee that the output voltage converges to the reference voltage in a finite time. As the saturation constraint is considered during the controller design, the duty ratio function of the converter satisfies the constraint between 0 and 1. For the unknown load or input

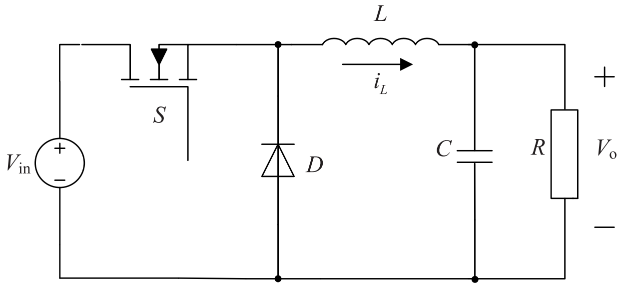


Fig. 1. DC–DC buck converter.

voltage, a finite-time convergent observer is designed to estimate the unknown load or input voltage. Finally, an adaptive finite-time control (AFC) algorithm is developed.

## II. Preliminaries and Problem Formulation

### A. Buck-Type DC–DC Converter

The buck-type dc–dc converter is shown in Fig. 1. $V_{\text{in}}$ is a dc input voltage source, $S$ is a controlled switch, $D$ is a diode, $V_o$ is sensed output voltage, and $L, C, R$ are the inductance, capacitance, and load resistance, respectively. Based on the average state space model [8], the dynamic equation for the buck converter is

$$\dot{i}_L = \frac{1}{L}(\mu V_{\text{in}} - V_o), \quad \dot{V}_o = \frac{1}{C}\left(i_L - \frac{V_o}{R}\right) \tag{1}$$

where $\mu$ is the duty ratio function as well as the control input and $\mu \in [0, 1]$. Let $V_{\text{ref}}$ be the desired dc output voltage and $x_1 = V_{\text{ref}} - V_o$ be the output voltage error. It follows from (1) that the error dynamic equation is:

$$\begin{aligned} \dot{x}_1 &= x_2 = -\dot{V}_o \\ \dot{x}_2 &= -\frac{1}{\text{LC}}x_1 - \frac{1}{\text{RC}}x_2 - \frac{V_{\text{in}}}{\text{LC}}\mu + \frac{V_{\text{ref}}}{\text{LC}}. \end{aligned} \tag{2}$$

Based on this model, the control objective is to design a control law such that the tracking error $x_1$ converges to zero. Note that the most of existing control algorithms only guarantee that the output voltage converges to the reference voltage asymptotically. Here, using the proposed results in this brief, we can guarantee that the output voltage tracks the reference voltage in a finite time.

### B. Time-Scaling Coordinate Changes

For the convenience of controller design, a time-scaling coordinate change is first employed to (2). Let $t = Ms$ with a positive constant $M$, which will be determined later. Define the following coordinate change:

$$\begin{aligned} z_1(s) &= x_1(Ms) = x_1(t) \\ z_2(s) &= Mx_2(Ms) = Mx_2(t) \\ u(s) &= M^2\mu(Ms) = M^2\mu(t) \end{aligned} \tag{3}$$

under which we obtain

$$\begin{aligned} \frac{dz_1(s)}{ds} &= \frac{dx_1(Ms)}{ds} = Mx_2(Ms) = z_2(s) \\ \frac{dz_2(s)}{ds} &= M\frac{dx_2(Ms)}{ds} \\ &= -\frac{1}{\text{LC}}M^2 z_1(s) - \frac{1}{\text{RC}}Mz_2(s) - \frac{V_{\text{in}}}{\text{LC}}u(s) \\ &\quad + \frac{V_{\text{ref}}}{\text{LC}}M^2. \end{aligned} \tag{4}$$

For the sake of statement, denote

$$b_1 = \frac{1}{\text{LC}}M^2, b_2 = \frac{1}{\text{RC}}M, f(s) = -\frac{V_{\text{in}}}{\text{LC}}u(s) + \frac{V_{\text{ref}}}{\text{LC}}M^2 \tag{5}$$

which leads to

$$\frac{dz_1(s)}{ds} = z_2(s), \frac{dz_2(s)}{ds} = -b_1 z_1(s) - b_2 z_2(s) + f(s) \tag{6}$$

where $b_1 > 0, b_2 > 0$, and $f(s)$ is the controller.

### C. Some Useful Definitions and Lemmas

*Definition 1:* Denote $\text{sig}^\alpha(x) = \text{sign}(x)|x|^\alpha$, where $\alpha \geq 0$, $x \in R$, $\text{sign}(\cdot)$ is the standard sign function.

*Definition 2:* Define a class of new saturation functions

$$\text{sat}_\alpha(x) = \begin{cases} \text{sign}(x), & \text{for } |x| > 1 \\ \text{sig}^\alpha(x), & \text{for } |x| \leq 1 \end{cases} \tag{7}$$

where $0 \leq \alpha \leq 1, x \in R$.

*Definition 3 (Homogeneity [17]):* Consider the system

$$\dot{x} = f(x),\ f(0) = 0,\ x(0) = x_0,\ x \in R^m \tag{8}$$

where $f : U_0 \mapsto R^m$ is continuous on an open neighborhood $U_0$ of the origin. Define dilation $(r_1, \ldots, r_m) \in R^m$ with $r_i > 0, i = 1, \ldots, m$. Let $f(x) = [f_1(x), \ldots, f_m(x)]^T$ be a continuous vector field. $f(x)$ is said to be homogeneous of degree $k \in R$ with respect to dilation $(r_1, \ldots, r_m)$ if, for any given $\varepsilon > 0$

$$f_i\left(\varepsilon^{r_1}x_1, \ldots, \varepsilon^{r_m}x_m\right) = \varepsilon^{k+r_i}f_i(x), i = 1, \ldots, m, \forall x \in R^m$$

where $k > -\min\{r_i, i = 1, \ldots, m\}$. System $\dot{x} = f(x)$ is said to be homogeneous if $f(x)$ is homogeneous.

*Lemma 1 [18]:* Consider the following system:

$$\dot{x} = f(x) + \widehat{f}(x),\quad f(0) = 0,\quad x \in R^m \tag{9}$$

where $f(x)$ is a continuous homogeneous vector field of degree $k < 0$ with respect to $(r_1, \ldots, r_m)$ and $\widehat{f}(x)$ satisfies $\widehat{f}(0) = 0$. Assume that $x = 0$ is an asymptotically stable equilibrium of the system $\dot{x} = f(x)$. Then $x = 0$ is a locally finite-time stable equilibrium of system (9) if

$$\lim_{\varepsilon \to 0} \frac{\widehat{f_i}(\varepsilon^{r_1}x_1, \ldots, \varepsilon^{r_m}x_m)}{\varepsilon^{r_i+k}} = 0, i = 1, 2, \ldots, m, \forall x \neq 0. \tag{10}$$

In addition, if system (9) is globally asymptotically stable and locally finite-time stable, then this system is globally finite-time stable.

## III. Design of Saturated Finite-Time Controller

In this section, the main objective is to design a saturated finite-time controller for system (6). In what follows, when there is no confusion, the function $\dot{z}(s)$ represents the derivative of $z(s)$ with respect to variable $s$.

*Theorem 1:* For system (6), if the controller is designed as

$$f = -k_1 \mathrm{sat}_{\alpha_1}(z_1) - k_2 \mathrm{sat}_{\alpha_2}(z_2) \tag{11}$$

where $k_1 > 0, k_2 > 0, 0 < \alpha_1 < 1, \alpha_2 = 2\alpha_1/(1+\alpha_1)$, then the system (6) state will be stabilized to zero in a finite time, i.e., $(z_1(s), z_2(s)) \to 0$ in a finite time.

*Proof:* The proof is divided into two steps. First, it will be proven that closed-loop system is globally asymptotically stable. Then, at the second step, it will be shown that the closed-loop system is locally finite-time stable.

*Step 1 (Proof of Global Asymptotical Stability):* A candidate Lyapunov function for system (6) is chosen as

$$V = k_1 \int_0^{z_1} \mathrm{sat}_{\alpha_1}(\rho) d\rho + \frac{1}{2} b_1 z_1^2 + \frac{1}{2} z_2^2. \tag{12}$$

Since

$$\frac{d \int_0^{z_1} \mathrm{sat}_{\alpha_1}(\rho) d\rho}{ds} = \mathrm{sat}_{\alpha_1}(z_1)\dot{z}_1$$

then

$$\begin{aligned} \dot{V}|_{(6)} &= k_1 \mathrm{sat}_{\alpha_1}(z_1) z_2 + b_1 z_1 z_2 - b_1 z_1 z_2 - b_2 z_2^2 + z_2 f \\ &= k_1 \mathrm{sat}_{\alpha_1}(z_1) z_2 - b_2 z_2^2 + z_2 f. \end{aligned} \tag{13}$$

Under the designed controller (11), one obtains

$$\begin{aligned} \dot{V}|_{(6)-(12)} &= k_1 \mathrm{sat}_{\alpha_1}(z_1) z_2 - b_2 z_2^2 \\ &\quad - k_1 z_2 \mathrm{sat}_{\alpha_1}(z_1) - k_2 z_2 \mathrm{sat}_{\alpha_2}(z_2) \\ &= -b_2 z_2^2 - k_2 z_2 \mathrm{sat}_{\alpha_2}(z_2). \end{aligned} \tag{14}$$

Due to

$$z_2 \mathrm{sat}_{\alpha_2}(z_2) = \begin{cases} |z_2|, & \text{for } |z_2| > 1 \\ |z_2|^{\alpha_2+1}, & \text{for } |z_2| \le 1 \end{cases}$$

then one concludes that $\dot{V} \le 0$.

Define set $\Psi = \{(z_1, z_2) | \dot{V} \equiv 0\}$. It follows from (14) that $\dot{V} \equiv 0$ means $z_2 \equiv 0$ and $\dot{z}_2 \equiv 0$. By (6) and (11), one obtains

$$-b_1 z_1 - k_1 \mathrm{sat}_{\alpha_1}(z_1) \equiv 0. \tag{15}$$

Note that, $b_1 > 0, k_1 > 0$, then $x_1 \equiv 0$. Thus, based on LaSalle's invariant principle [19], it can be concluded that $(z_1(s), z_2(s)) \to 0$ as $s \to 0$. That is to say that the close-loop system (6) with (11) is globally asymptotically stable.

*Step 2 (Proof of Local Finite-Time Stability):* Rewrite the closed-loop system (6) and (11) as

$$\dot{z}_1 = z_2, \dot{z}_2 = -k_1 \mathrm{sat}_{\alpha_1}(z_1) - k_2 \mathrm{sat}_{\alpha_2}(z_2) + g(z_1, z_2) \tag{16}$$

where $g(z_1, z_2) = -b_1 z_1 - b_2 z_2$.

Since system (16) is globally asymptotically stable, then the system state $(z_1, z_2)$ will converge to the region $|z_1| \le 1, |z_2| \le 1$ in a finite time and stay there for ever. After then, there will be no saturation constraint and the system (16) can be rewritten as

$$\dot{z}_1 = z_2, \dot{z}_2 = -k_1 \mathrm{sig}^{\alpha_1}(z_1) - k_2 \mathrm{sig}^{\alpha_2}(z_2) + g(z_1, z_2). \tag{17}$$

For the nominal pat of system (17), that is

$$\dot{z}_1 = z_2, \dot{z}_2 = -k_1 \mathrm{sig}^{\alpha_1}(z_1) - k_2 \mathrm{sig}^{\alpha_2}(z_2) \tag{18}$$

by choosing Lyapunov function as

$$W = k_1 \int_0^{z_1} \mathrm{sig}^{\alpha_1}(\rho) d\rho + \frac{1}{2} z_2^2 \tag{19}$$

the derivative is

$$\dot{W}|_{(18)} = -k_2 z_2 \mathrm{sig}^{\alpha_2}(z_2) = -k_2 |z_2|^{\alpha_2+1}. \tag{20}$$

Similar to step 1, it can be first proved that the system (18) is asymptotically stable.

In addition, note that $0 < \alpha_1 < 1, \alpha_2 = 2\alpha_1/(1+\alpha_1)$. According to Definition 3, it can be verified that system (18) is homogeneous of degree $m = (\alpha_1 - 1)/2 < 0$ with $r_1 = 1$, $r_2 = (\alpha_1 + 1)/2$.

Next it will be shown that $\lim_{\varepsilon \to 0} (g(\varepsilon^{r_1} z_1, \varepsilon^{r_2} z_2)/\varepsilon^{r_2+m}) = 0$, for $\forall (z_1, z_2) \ne (0, 0)$. According to the definition of function $g(\cdot)$, we get

$$\begin{aligned} \lim_{\varepsilon \to 0} \frac{g(\varepsilon^{r_1} z_1, \varepsilon^{r_2} z_2)}{\varepsilon^{r_2+m}} &= \lim_{\varepsilon \to 0} \frac{-b_1 \varepsilon^{r_1} z_1 - b_2 \varepsilon^{r_2} z_2}{\varepsilon^{r_2+m}} \\ &= \lim_{\varepsilon \to 0} \left(-b_1 \varepsilon^{r_1 - r_2 - m} z_1 - b_2 \varepsilon^{-m} z_2\right). \end{aligned}$$

As $r_1 > r_2, m < 0$, then

$$\lim_{\varepsilon \to 0} \left(-b_1 \varepsilon^{r_1 - r_2 - m} z_1 - b_2 \varepsilon^{-m} z_2\right) = 0 \tag{21}$$

that is, $\lim_{\varepsilon \to 0} (g(\varepsilon^{r_1} x_1, \varepsilon^{r_2} x_2)/\varepsilon^{r_2+m}) = 0$.

Thus, according to Lemma 1, it can be concluded that system (17) is globally finite-time stable, i.e., system (16) is locally finite-time stable.

By the proposed results in steps 1 and 2, it can be found that the system (6) under the controller (11) is globally finite-time stable. The proof is completed.

*Remark 1:* Obviously, the proposed controller (11) is bounded. Specifically, from the definition (2), it can be found that $|f| \le k_1 + k_2$. Hence, the bound of a proposed controller can be adjusted through $k_1, k_2$ to any small level to accommodate the actuation bound in practical implementation.

## IV. Design of Fast Finite-Time Voltage Regulation Control Algorithm

*Theorem 2:* For the buck-type dc–dc converter (1), if the duty ratio is designed as

$$\mu(t) = \frac{V_{\mathrm{ref}}}{V_{\mathrm{in}}} + \frac{\mathrm{LC}}{M^2 V_{\mathrm{in}}} \left[k_1 \mathrm{sat}_{\alpha_1}(x_1(t)) + k_2 \mathrm{sat}_{\alpha_2}(M \cdot x_2(t))\right] \tag{22}$$

where $k_1 > 0, k_2 > 0, 0 < \alpha_1 < 1, \alpha_2 = 2\alpha_1/(1+\alpha_1)$, then the output voltage $V_o$ converges to the reference voltage $V_{\mathrm{ref}}$ in a finite time.

## V. Design of Fast Adaptive Finite-Time Voltage Regulation Control Algorithm

In this section, the case when the load resistance and the input voltage are unknown will be considered. First, by (1) and (2), to obtain the state $x_2$, it is required to know the precise value of load resistance $R$. However, in practice, the load and

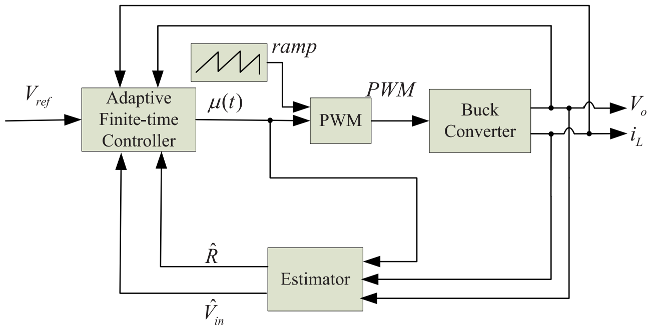


Fig. 2. Block diagram of adaptive finite-time voltage control algorithm.

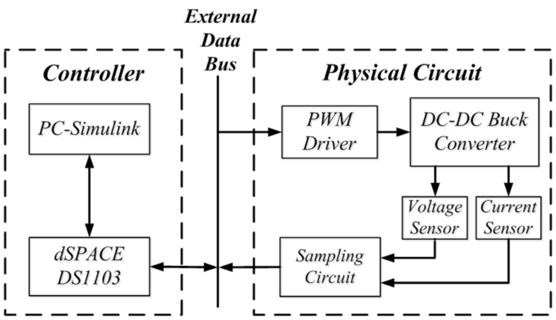


Fig. 3. Configuration of the experimental system of the buck-type dc–dc converter.

the input voltage are usually time-varying and unknown. To this end, the technique of finite-time convergent observer will be employed to estimate the unknown load resistance and input voltage in a finite time.

Denote $\hat{R}$ as the estimate value of the unknown load resistance $R$. For the sake of statement, let $\theta = -1/R$ (which is an unknown parameter) and $\hat{\theta}$ as its estimate value.

*Theorem 3:* For the buck-type dc–dc converter with an unknown load resistance, if the estimator is designed as

$$\begin{aligned}\dot{\hat{V}}_o &= \frac{1}{C}\left(i_L + \hat{\theta} V_o\right) + l_1 V_o \text{sig}^{\beta_1}\left(V_o - \hat{V}_o\right)\\ \dot{\hat{\theta}} &= l_2 V_o \text{sig}^{\beta_2}\left(V_o - \hat{V}_o\right)\end{aligned} \tag{23}$$

then the estimate value $\hat{\theta}$ (or $\hat{R} = -1/\hat{\theta}$) will converge to the real value $\theta$(or $R$) in a finite time, where $l_1, l_2$ are appropriate positive gains, $0.5 < \beta_1 < 1, \beta_2 = 2\beta_1 - 1$. ■

*Proof:* Define the estimate error

$$e_1 = V_o - \hat{V}_o, e_2 = \theta - \hat{\theta}. \tag{24}$$

Noting that $\theta$ is a constant, then it follows from (23) that:

$$\dot{e}_1 = \frac{V_o}{C} e_2 - l_1 V_o \text{sig}^{\beta_1}(e_1), \dot{e}_2 = -l_2 V_o \text{sig}^{\beta_2}(e_1). \tag{25}$$

Based on [20, Th. 3.1], there are appropriate gains such that the error system (25) is finite-time convergent. That is to say that the estimate value $\hat{\theta}$ will converge to the real value $\theta$ in a finite time, i.e., the load resistance estimate value $\hat{R} = -1/\hat{\theta}$ will converge to the real value $R$ in a finite time. ■

Similarly, we can design the following finite-time estimator to estimate the unknown input voltage in a finite time.

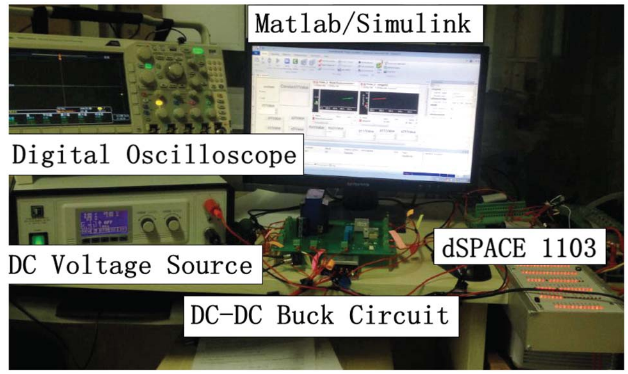


Fig. 4. Experimental test system setup of the buck-type dc–dc converter.

*Theorem 4:* For the buck-type dc–dc converter with an unknown input voltage, if the estimator is designed as

$$\begin{aligned}\dot{\hat{i}}_L &= -\frac{V_o}{L} + \frac{\mu}{L}\hat{V}_{\text{in}} + l_3 \mu \text{sig}^{\beta_3}\left(i_L - \hat{i}_L\right)\\ \dot{\hat{V}}_{\text{in}} &= l_4 \mu \text{sig}^{\beta_4}\left(i_L - \hat{i}_L\right)\end{aligned} \tag{26}$$

then the estimate value $\hat{V}_{\text{in}}$ will converge to the real value $V_{\text{in}}$ in a finite time, where $l_3, l_4$ are appropriate positive gains, $0.5 < \beta_3 < 1, \beta_4 = 2\beta_3 - 1$.

With the help of Theorems 3 and 4, when the load resistance and the input voltage are unknown, we can design a fast AFC algorithm as follows:

$$\begin{aligned}\mu(t) &= \frac{V_{\text{ref}}}{\hat{V}_{\text{in}}} + \frac{\text{LC}}{M^2 \hat{V}_{\text{in}}}\left[k_1 \text{sat}_{\alpha_1}(x_1) + k_2 \text{sat}_{\alpha_2}\left(M\hat{x}_2\right)\right]\\ &= \frac{V_{\text{ref}}}{\hat{V}_{\text{in}}} + \frac{\text{LC}}{M^2 \hat{V}_{\text{in}}}\Bigg[k_1 \text{sat}_{\alpha_1}(V_{\text{ref}} - V_o)\\ &\qquad + k_2 \text{sat}_{\alpha_2}\left[\frac{M}{C}\left(\frac{V_o}{\hat{R}} - i_L\right)\right]\Bigg].\end{aligned} \tag{27}$$

Fig. 2 shows the control scheme for buck converter by using the proposed AFC algorithm.

## VI. Experiment Results

To evaluate the performance of the proposed method, the experimental setup system for a buck dc–dc converter was built. Based on the MATLAB/Simulink and dSPACE 1103, the configuration and experimental test setup are shown in Figs. 3 and 4, respectively, where the frequency for triangular wave is chosen as 20 kHz.

### A. Parameters

The components values of the buck converter are given as: input voltage $V_{\text{in}} = 3$ V, inductance $L = 0.1$ mH, capacitance $C = 100\ \mu$F, load resistance $R = 10\ \Omega$, and desired output voltage $V_{\text{ref}} = 1.5$ V.

To have a comparison, two kinds of control algorithms are employed, i.e., one is the proposed fast AFC algorithm (27), the other one is PI control algorithm. As for the choice of controller gains, they were chosen using trial and error methods based on a large number of simulations and experiments. The gains that provided the most satisfactory performance were chosen. For the results detailed, the PI controller gains are: $K_P = 0.08$ and $K_I = 0.7$. For the AFC algorithm, the control gains are: $M = 0.0001, k_1 = 0.13, k_2 = 1.5, \alpha_1 = 0.5, l_1 = 300, l_2 = 100, \beta_1 = 0.55$.

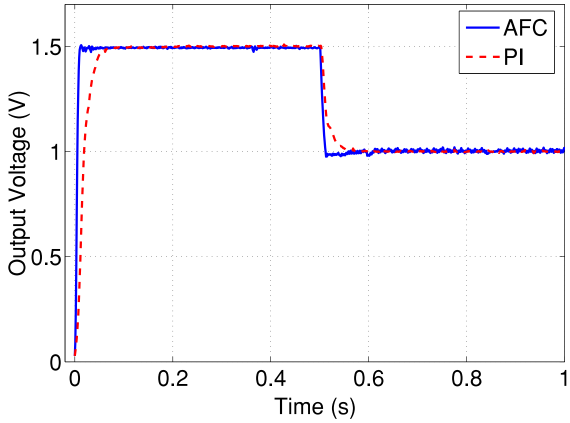


Fig. 5. Experiment results for the output voltage under two control algorithms and different reference voltages. (a) AFC. (b) PI control.

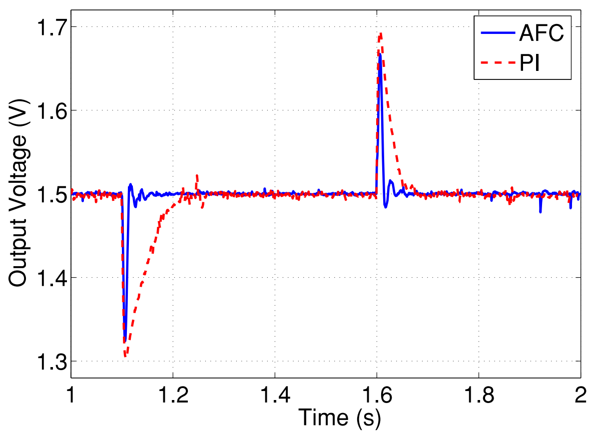


Fig. 6. Experiment results for the output voltage under two control algorithms and load variations. (a) AFC. (b) PI control.

### *B. Dynamical Responses Under Different Reference Voltages*

In this case, the load is kept unchanged and the reference voltage is changed from 1.5 to 1 V at 0.5 s. The responses under two different control algorithms are given in Fig. 5. It can be found that the proposed fast AFC algorithm can offer a faster transient response.

### *C. Dynamical Responses Under Load Variations*

In this case, the load resistance is changed as

$$R = \begin{cases} 10, & 0 \leq t \leq 1.1\text{s} \\ 5, & 1.1\text{s} < t \leq 1.6\text{s} \\ 10, & t > 1.6\text{s} \end{cases} \tag{28}$$

and the other parameters are kept unchanged. The response curves for output voltage under two control algorithms are shown in Fig. 6 from which it can be found that, under the proposed control algorithm, the response rate is faster and the voltage fluctuation is smaller.

Hence, compared with the PI control algorithm, the proposed fast AFC algorithm in this brief has a faster regulation performance and stronger performance on load-variation rejection.

## VII. Conclusion

This brief has studied the voltage regulation problem for the buck-type dc–dc converter. Using the adaptive control method and finite-time control method, a novel voltage regulation algorithm has been proposed. Theoretical and experimental results have been presented to verify the effectiveness of the proposed method.

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