

Robust-Optimal Output-Voltage Control of Buck Converter using Fuzzy Adaptive Weighted Combination of Linear Feedback Controllers

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Abstract: This paper presents a computationally-intelligent adaptive weighted controller combination scheme to optimize the output-voltage regulation capability of a low power DC-DC buck converter. The proposed scheme beneficially combines two linear feedback controllers; namely, Proportional-Integral-Derivative (PID) controller and Linear-Quadratic-Regulator (LQR). The PID controller provides control effort based on the error-dynamics of output-voltage. Wherein, the term regarding the error-derivative is replaced with the information of capacitor-current to nullify the effects of noise injected by the derivative action during transients. The LQR provides optimal control decisions by utilizing the state-feedback of inductor-current and output-voltage. The outputs of PID controller and LQR are linearly combined by computing their weighted sum. The fixed weightages associated with each controller cannot compensate the parametric uncertainties and load-step transients. Therefore, the weightages are adaptively self-tuned via a hyperbolic tangent function of error in output voltage. The performance of weighted control scheme is also investigated by augmenting it with a fuzzy inference system that directly captures the variations in output-voltage and capacitor-current to adaptively self-tune the weightages. The performances of aforementioned weighted controllers are comparatively analyzed via credible real-time experiments. The fuzzy weighted controller yields time-optimal control effort during step-reference tracking and offers minimum-time transient recovery during load variations.

Keywords: DC-DC buck converter, linear quadratic regulator, proportional-integral-derivative controller, adaptive weighted control, fuzzy inference system.

1. INTRODUCTION

The buck converters are switch-mode electrical power supplies that step-down an input DC voltage signal and regulate the output DC voltage signal at a desired level, even if the load-impedance or input-voltage changes (Tahri et al., 2012). They are used in computer systems, DC machine drives, communication equipment, solar photovoltaic systems, and battery chargers due to their low power consumption and high efficiency (Ghosh and Banerjee, 2015; Akter et al., 2015).

Extensive research has been done to synthesize robust and optimal feedback controllers to further optimize the voltage regulation performance of the buck converters (Guo et al., 2009; Mariethoz et al., 2010; Lakshmi and Raja, 2014; Lindiya et al., 2015; Lee et al., 2016; Du et al., 2017; Batouba et al., 2017; Hossain et al., 2018). The buck converter uses a high-frequency switch to reduce the DC component of voltage to the required level. In addition, it consists of an inductor-capacitor filter that removes the undesired harmonics from the output-voltage signal (Rashid, 2009). The buck converter is a bilinear system because it presents a different linear-circuit topology during the on-time and off-time of the switch, in each time-period (Olalla et al., 2011). This periodic change in the circuit configuration leads to a

discontinuous behavior in the converters. Such variable structure systems are generally controlled with the aid of Sliding-Mode-Controllers (SMCs), wherein, the switching device is operated to track a given sliding surface in the phase space (Guldemir, 2011; Ling et al., 2016). The SMCs offer robust control effort and compensate the effects of nonlinearities associated with the system (Naik and Mehta, 2017). However, the superior performance of the SMCs comes at a cost of significant control energy (Dastidar, 2010). The SMCs tend to inject chattering in the system that may result in the premature wear and tear of the switching device (Utkin and Lee, 2006). The computational synthesis of a well-postulated SMC is quite cumbersome.

The linear feedback controllers have also been rigorously investigated by the researchers to effectively control the output of buck converters (Pedroso et al., 2013a). The linear controllers, either model-free or model-based, can be easily synthesized. Owing to their shortcomings in handling nonlinear and complex dynamical electro-mechanical systems, a lot of research has been done in robustifying their performance by augmenting them with additional tools. Among these techniques, the Linear-Quadratic-Regulator (LQR) and Proportional-Integral-Derivative (PID) controllers have simple yet robust control structures (Moreira et al., 2011; Hernandez et al., 2016; Debra et al., 2007; Chang et al.,

2017). When individually used, each of these schemes offer certain advantages and disadvantages (Yaseen, 2017). The LQR delivers optimal control decision by minimizing a quadratic cost function that directly captures the deviations occurring in the state trajectories and control effort (Abbas et al., 2010). Once the averaged mathematical model of the system is correctly identified, the optimal state-feedback gain vector can be easily computed offline using the Algebraic Riccati Equation (Lindiya et al., 2016). Although the LQR does not directly depend on the error-dynamics, the quadratic cost function delivers a large control effort for a small change occurring in the state-trajectories, which is quite beneficial in improving the transient recovery and error convergence-rate of the response. However, it yields a poor steady-state performance and exhibits significant over-shoots in the response (Saleem and Urwa, 2017). Being a model-based technique, LQR is also prone to be affected by the modeling errors. On the contrary, the PID controller is a model-free scheme (Anbarasi and Muralidharan, 2016). It is the weighted sum of error, time-integral of error, and time-derivative of error (Jalilvand et al., 2010; Seshagiri et al., 2016). Despite its simple structure, it yields an effective and robust control effort due to its dependence on the variation in error-dynamics. The integral action eliminates the steady state error, damps oscillations and suppresses the overshoots/undershoots. However, the integral damping degrades the error convergence rate of the response (Zhang et al., 2017). The slow response-times can be improved by introducing error-derivative control term. The derivative action improves the transitional times of the response and offers crude error-prediction based on the gradient of the error variations (Shang et al., 2009). However, it also amplifies and injects high frequency noise in the system that leads to fluctuations in the steady-state response. Despite its demerits, the importance of derivative action cannot be ignored. The error-derivative term can be indirectly retained in the PID control law by using the information of output capacitor current. The information provided by the capacitor current is synonymous to error-derivative (Kapat and Krein, 2012a). The PID controller formulation employing capacitor current to emulate the derivative control action is denoted as Geometric-PID controller, and it has been rigorously used to yield time-optimal control effort for DC-DC converters (Kapat and Krein, 2012b). The Geometric-PID controller can also compensate the hysteresis effect exhibited by parasitic impedances in the circuit.

This research paper contributes to the methodic synthesis of a robust and time-optimal control scheme for output-voltage regulation of buck converters using adaptive combination of two linear feedback controllers. As discussed earlier, each linear controller offers a unique set of attributes, and hence a trade-off has to be made between optimality, robustness, and design flexibility while selecting a particular control scheme (Abdullah et al., 2015). By using a hybrid of the PID controller and LQR, the favorable features rendered by each controller can be selectively harnessed based on the error-dynamics of the system in each sampling interval (Sun and Gan, 2010). The control effort provided by the individual linear controllers can be beneficially combined with the objective that they optimize the controller performance in

time-domain, enhance robustness against bounded exogenous disturbances, offer fast load-transient recovery, and minimize the cost of control signal (Pitel and Krein, 2009, Carradini et al., 2010). The linear combination of the PID controller and LQR has been done in the past by computing their weighted sum to control nonlinear dynamical systems (Salim et al., 2014; Bagheri et al., 2016). However, the fixed weightages associated with each controller may not always improve the control effort as desired. The two controllers impede each other's correctional efforts when the system encounters uncertain parametric variations. This phenomenon renders oscillations in the system's response as it transits from transient-state to steady-state, or vice-versa. In order to overcome this problem, the weightages are updated in each sampling interval via an online adaptation mechanism (Pedroso et al., 2013b). An adaptive weighted control scheme is presented in (Pedroso et al., 2013c) that uses a Hyperbolic-Tangent-Function (HTF) of error to dynamically adjust the weightages, and hence the contribution offered by the PID controller and LQR, in each sampling interval. The proposed scheme significantly improves the time-domain response and disturbance rejection capability of the system. The weighted collaboration scheme can be further synergized by using a more elaborate set of pre-defined rules that could aid in compensating the nonlinear dynamics of the system. The desired objectives can be achieved by augmenting the online adaptation mechanism with computational intelligence techniques (Nizami and Mahanta, 2016; Lian et al., 2017). Hence, in this research, a well-postulated Fuzzy-Inference-System (FIS) is also investigated as an online adaptation mechanism for the dynamic adjustment of weightages (Salam et al., 2011; Ang et al., 2017; Saleem et al., 2017). The proposed system captures the variations in output-voltage error and capacitor-current, in conjunction with a carefully orchestrated fuzzy rule-base, to render time-optimal reference tracking performance and enhanced disturbance-attenuation capability under load-transient conditions in the system.

The rest of the paper is organized as follows: Section 2 presents the details of the experimental setup of buck converter. The theoretical background of the PID controller and LQR is presented in Section 3. The synthesis of HTF and FIS based adaptive weighted control schemes are discussed in Section 4. The experimental tests and the comparative performance assessment of their results are summarized in Section 5. The paper is concluded in the Section 6.

2. EXPERIMENTAL SETUP

The buck converter transforms a higher DC voltage to lower DC voltage level. The circuit diagram is shown in Fig. 1.

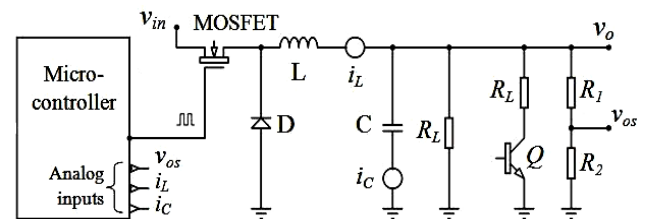


Fig. 1. DC-DC buck converter circuit

The input voltage, v_{in} , to the circuit is passed through a high-frequency switching transistor, an N-channel MOSFET, that chops down the applied v_{in} into a rectangular waveform. The low-pass passive filter formed by the inductor-capacitor network allows only the DC component of the rectangular waveform to appear at the output. The output voltage, v_o , of the converter is given by (1). It is varied by changing the duty-cycle, d , of the MOSFET as given by (2).

$$v_o = d \times v_{in} \quad (1)$$

such that,

$$d = \frac{t_{on}}{t_{on} + t_{off}} \quad (2)$$

where, t_{on} and t_{off} refers to the on-time and off-time of switching period of MOSFET, respectively. In case of any unprecedented changes in the load-resistance, R_L , or v_{in} , the controller appropriately changes the duty-cycle of the switching period in order to maintain the v_o at the reference value. The diode, D , is reversed biased during the on-time of switch. Thus, the entire input current is supplied to the capacitor, C , and R_L via the inductor, L . In addition, the current charges L as well. During the off-time of switch, the input current supply from the source to the circuit is cut-off. Hence, L utilizes its stored energy to supply the current to C and R_L . The diode stays forward-biased and completes the circuit during the off-time. Prior to the complete discharging of L , the MOSFET is turned-on again (Rashid, 2009). The system's hardware setup and mathematical model are presented in the following sub-sections.

2.1 Hardware Setup

The programmable buck convertor module, used for experimentation, is shown in Fig. 2. The state-variables, i_L (inductor-current) and v_o , are measured in real-time using on-board current-sensing resistor of 0.01Ω (less than the value of inductor's Equivalent-Series-Resistance) and voltage-divider circuit (formed by R_1 and R_2), respectively. An 8-bit embedded microcontroller, ATMEGA-328, acts as a relay to serially communicate sensor-measurements and control-signals between the buck converter circuit and the control system software. The microcontroller acquires and digitizes the analog sensor-readings via its analog input channels. The sensor-readings are sampled at 250 Hz. The microcontroller serially transmits the acquired sensor data, at 9600 bps, to the control algorithm that is implemented in a LabVIEW based computer application (Demirtas and Gezer, 2010). The application graphically represents the variations in v_o in real-time and generates appropriate control commands, $u(t)$. These commands are serially transmitted back to the embedded microcontroller that converts them into high-frequency Pulse-Width-Modulated (PWM) signals, $d(t)$. A switching frequency of 150 kHz is used in this research. The PWM signals are applied to the MOSFET (IRF540N) via a dedicated driver circuit. The driver circuit amplifies the PWM signal to turn on the MOSFET. Additionally, it optically isolates the digital control circuit from the power electronic circuit to prevent it from the inductive-kick phenomenon.

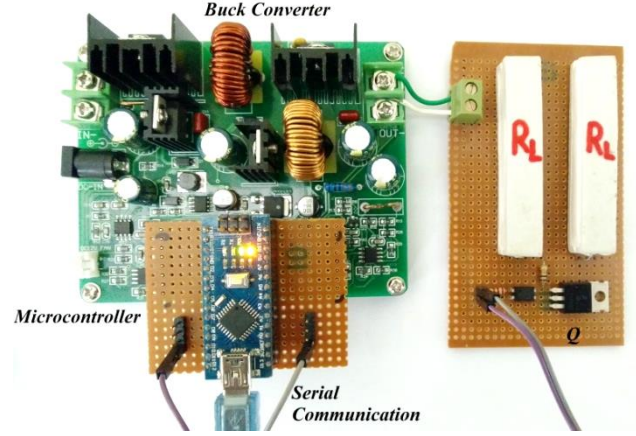


Fig. 2. Experimental setup

The circuit also consists of a cyclically switched resistance (CSR) that is connected in parallel with the R_L . The CSR consists of a resistor identical to R_L connected in series with an NPN power bipolar junction transistor (TIP122). This transistor is denoted as Q in Fig. 1 and Fig. 2. The transistor Q remains “turned-on” under normal conditions, leading to an overall load-resistance of $0.5R_L$. When the transistor Q is turned-off, it introduces an incremental load-step transient in the response of v_o , and vice-versa. A regulated 15.0 V DC signal is applied as v_{in} to the circuit.

2.2 Mathematical Model

The state-space model of a linear dynamical system is given by (3) and (4).

$$\dot{x}(t) = \mathbf{A}x(t) + \mathbf{B}u(t) \quad (3)$$

$$y(t) = \mathbf{C}x(t) + \mathbf{D}u(t) \quad (4)$$

where, $x(t)$ is the state-vector, $y(t)$ is the output-vector, $u(t)$ is the control input signal, $\mathbf{A}$ is the system matrix, $\mathbf{B}$ is the input matrix, $\mathbf{C}$ is the output matrix, and $\mathbf{D}$ is the feed-forward matrix. The vectors, $x(t)$ and $u(t)$, are given by (5).

$$x(t) = [v_o(t) \quad i_L(t)]^T, \quad u(t) = d(t) \quad (5)$$

where, $d(t)$ is the instantaneous value of duty-cycle. Correspondingly, the matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, and $\mathbf{D}$ of the buck converter's state-space model are identified in (6), (Priewasser et al., 2014).

$$\mathbf{A} = \begin{bmatrix} -\frac{1}{C(0.5R_L + r_c)} & \frac{0.5R_L}{C(0.5R_L + r_c)} \\ -\frac{0.5R_L}{L(0.5R_L + r_c)} & -\frac{(r_L + r_c || 0.5R_L)}{L} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ \frac{v_{in}}{L} \end{bmatrix}, \quad \mathbf{C} = [1 \quad 0], \quad \mathbf{D} = 0 \quad (6)$$

The design parameters of the DC-DC buck converter used in this research are identified in Table 1.

3. THEORATICAL BACKGROUND OF THE CONTROLLERS

This section presents the theoretical background of the individual control schemes.

Table 1. Design parameters of buck-converter circuit

Parameters	Symbol	Values
Voltage divider resistance	R_1, R_2	150 k Ω
Load resistance	R_L	22 Ω
Inductance	L	330 μ H
Capacitance	C	1000 μ F
Equivalent-Series-Resistance (ESR) of capacitor	r_c	0.08 Ω
Equivalent-Series-Resistance (ESR) of inductor	r_L	0.05 Ω
Equivalent-Series-Inductance (ESL) of capacitor	L_c	100 nH

3.1 PID Controller

The PID controllers are widely used in control industry due to their simple structure and ability to deliver effective control effort (Bhatti et al., 2015; Astom and Hagglund, 2016). The PID controller is a weighted linear combination of error, integral-of-error, and derivative-of-error (or v_o , since v_{ref} is constant). The PID control law and the error are expressed in (7) and (8), respectively.

$$u_{pid}(t) = k_p e(t) + k_i \left(\int_0^t e(\tau) d\tau \right) - k_d (v_o(t)) \quad (7)$$

$$e(t) = v_{ref} - v_o(t) \quad (8)$$

The weightages of the aforementioned error-dynamics are referred to as the proportional gain (k_p), integral gain (k_i), and derivative gain (k_d), respectively. In this research, the PID gains are heuristically tuned by minimizing the Integral-Time-Absolute-Error (ITAE) criterion, shown in (9), (Vijaykumar and Manigandan, 2016; Taha et al., 2018).

$$J_1 = \int_0^t \tau |e(\tau)| d\tau \quad (9)$$

The heuristically optimized values of k_p , k_i , and k_d used in this research are 1.79 V^{-1} , 8.68 $(Vs)^{-1}$, and $2.49 \times 10^{-6} sV^{-1}$, respectively. The integral action in the PID control law serves to eliminate the steady-state error and damps the oscillation or overshoots caused by the transients or random perturbations in the system. However, it slows down the response as well. The derivative action is usually added in the control law to improve the system's phase margin and asymptotic convergence. However, the derivative control significantly amplifies (and injects) the high frequency noise signals to the closed-loop system. Additionally, the practical parasitic impedance(s), such as the ESR and ESL of the capacitor further limit the performance of the derivative control term in PID controller. These parasitic impedances introduce large overshoots (and undershoots) during large-signal transients, and discrete jumps at switching transitions with alternating polarities during small-signal transients or steady-state. In order to solve these problems, a geometric

formulation of PID controller is derived (Kapat and Krein, 2012a). The Geometric-PID controller replaces the $v_o(t)$ term in the control law with the instantaneous information of the capacitor current, i_c , (Kapat and Krein, 2012b). The geometric PID formula uses the state-feedback of v_o and i_c . The practical model of the output capacitor is given by (10).

$$v_o(t) = L_c \frac{d^2 i_c(t)}{dt^2} + r_c \frac{di_c(t)}{dt} + \frac{i_c(t)}{C} \quad (10)$$

With the inclusion of random disturbances in the capacitor model, the capacitor current, i_c , of the buck converter is expressed according to (11).

$$i_c(t) \approx \frac{r_c}{r_c C - L_c} \left(e^{\frac{-t}{r_c C}} - e^{\frac{-r_c t}{L_c}} \right) C (v_o(t)) \quad (11)$$

If the values of parasitics (ESR and ESL) are considered negligibly small, then the capacitor current becomes equal to $C(v_o(t))$. Consequently, the Geometric-PID control law is given by (12).

$$u_{pid}(t) = k_p e(t) + k_i \left(\int_0^t e(\tau) d\tau \right) - \frac{k_d}{C} i_c(t) \quad (12)$$

The current i_c is sensed by using a current-sensing power resistor of 0.01 Ω (less than the value of capacitor's ESR). Its effect on the closed-loop bandwidth and stability is insignificant (Kapat and Krein, 2012a). The i_c neither shows any overshoots nor any discrete jumps during steady-state. Moreover, it also feeds forward the variations in the load-current and i_L .

3.2 Linear Quadratic Regulator

The LQR is an optimal controller that uses the linear state-space model of the dynamical system along with the full state-feedback of the system to deliver optimal control decisions (Lewis et al., 2012). It achieves this optimality by minimizing a quadratic performance index, given by (13).

$$J_2 = \int_0^t x^T(\tau) \mathbf{M} x(\tau) + u^T(\tau) \mathbf{R} u(\tau) d\tau \quad (13)$$

where, the $\mathbf{M}$ and $\mathbf{R}$ are the state- and control-penalty matrices, respectively. These matrices are chosen such that; $\mathbf{M}$ is positive semi-definite and $\mathbf{R}$ is positive definite. In this research, the $\mathbf{M}$ and $\mathbf{R}$ matrices are tuned by minimizing the ITAE criterion shown in (9). The resulting matrices are expressed in (14).

$$\mathbf{M} = \begin{bmatrix} 4.81 \times 10^6 & 0 \\ 0 & 3.37 \times 10^7 \end{bmatrix}, \quad \mathbf{R} = 100 \quad (14)$$

The state-feedback control law used to generate the optimal control commands is given by (15).

$$u_{lqr}(t) = -\mathbf{K}x(t) \quad (15)$$

where, $\mathbf{K}$ is denoted as the state-feedback gain vector. The gain vector relocates the open-loop poles of the system to

synthesize an optimal controller. The state-feedback gain vector, $\mathbf{K}$, is given by (16).

$$\mathbf{K} = \mathbf{R}^{-1} \mathbf{B}^T \mathbf{P} \quad (16)$$

where, $\mathbf{P}$ is a symmetric positive definite matrix that is evaluated using the Algebraic Riccati Equation, as shown in (17).

$$\mathbf{A}^T \mathbf{P} + \mathbf{P} \mathbf{A} - \mathbf{P} \mathbf{B} \mathbf{R}^{-1} \mathbf{B}^T \mathbf{P} + \mathbf{M} = 0 \quad (17)$$

According to the performance index, the applied control signal is proportional to the square of state variations; therefore, if the variations are large, the minimization is faster and the LQR system converges quickly to deliver an optimal gain vector. Consequently, the transition time of the response and error convergence improves significantly while utilizing minimum energy. The optimal gain vector is evaluated offline using the expression given (16). The evaluate gain vector is given by (18).

$$\mathbf{K} = [165.93 \quad 579.68] \quad (18)$$

4. WEIGHTED CONTROLLER COMBINATION SCHEME

Owing to their design limitations, each of the aforementioned control schemes offer unique attributes in controlling v_o of the buck converter. Individually they are capable of enhancing only a few performance parameters. Thus, the two control schemes are beneficially combined by taking their weighted sum, as shown in (19), in order to fulfill the desired control objectives.

$$u(t) = w \times u_{lqr}(t) + (1 - w) \times u_{pid}(t) \quad (19)$$

where, w is the weightage, such that $w \in [0, 1]$.

4.1 Fixed Weighted Controller (f-WC)

Initially, the w is tuned by minimizing the ITAE criterion given in (9). The value of w thus selected is 0.659. This controller-variant is denoted as the fixed-Weighted-Controller (f-WC). The block diagram of the f-WC is shown in Fig. 3.

4.2 Adaptive Weighted Control Scheme

The controller combination can be further synergized by utilizing an adaptive weighted control mechanism.

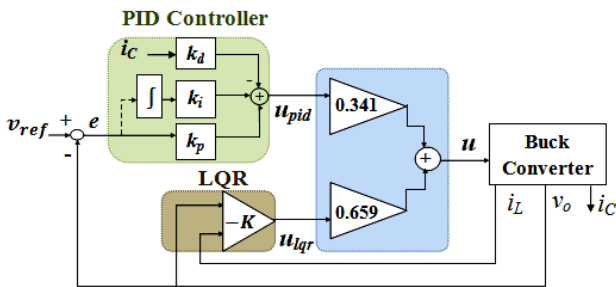


Fig. 3. Fixed Weighted Controller (f-WC)

Wherein the weightages are adaptively modulated using a predefined set of rules that depends on the error-dynamics of the system. The proposed scheme effectively combines the efficiencies of each controller and suppresses their deficiencies, so that only the best features of the available control resources are utilized at a given instant. Additionally, it ensures a smoother transition between the controllers. In this research, two different online adaptation mechanisms are investigated for dynamic adjustment of w ; namely, HTF and FIS.

A. Hyperbolic Adaptive Weighted Controller (HAWC)

The weightages of the WSM can be automatically updated by using a nonlinear function of output voltage error, as proposed in (Pedroso et al., 2013). The nonlinear scaling of the individual control schemes enables the overall control mechanism to quickly adapt to variations in system dynamics and improve the reference-tracking accuracy. Therefore, a smooth nonlinear function is needed to change the values of w automatically with respect to error-variations. The refined control objectives are explained as follows: If the error is large, the value of w should appropriately increase the control contribution of the LQR so that the convergence-rate of the response is significantly improved. For smaller errors (as the response approaches the reference voltage), the PID controller contribution should be increased so that the steady-state fluctuations are eliminated and the oscillations (or overshoots) are suppressed. Owing to these objectives, an HTF is chosen for adaptive collaboration of the two controllers, as given by (20). The waveform of HTF is shown in Fig. 4.

$$w = \tanh |e(t)| \quad (20)$$

The control system architecture of the Hyperbolic Adaptive Weighted Controller (HAWC) is illustrated in Fig. 5.

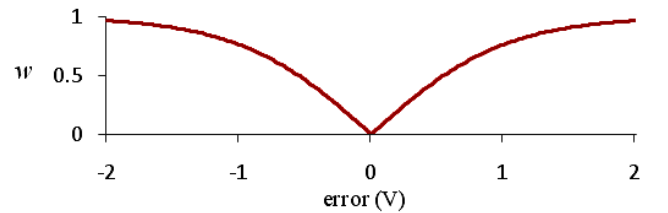


Fig. 4. Waveform of HTF

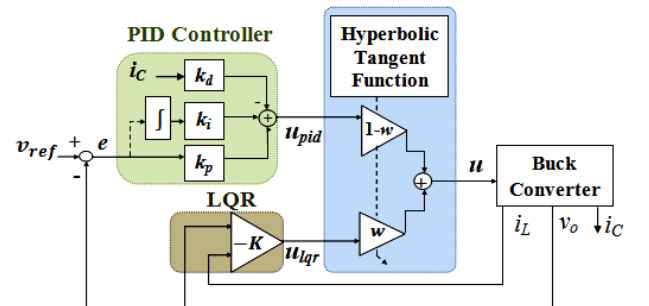


Fig. 5. Hyperbolic Adaptive Weighted Controller (HAWC)

B. Fuzzy Adaptive Weighted Controller (FAWC)

The HTF provides a formidable platform for adaptive tuning of w . However, since it depends on the error only, the HTF may not ensure efficient commutation between the appropriate controllers in order to effectively handle unprecedented dynamic variations occurring in the system. To further enhance the controller's performance, a two-input and one-output fuzzy inference system (FIS) is employed for the real-time self-tuning of w . The FIS is a model-free technique that optimizes the hyper-parameters according to a set of heuristically synthesized logical (if-then-else) rules (Lian et al., 2006; Martinez et al., 2012; Mahendran et al., 2016). The fuzzy rule base is carefully synthesized to include correctional efforts regarding various control scenarios emerging in the system. Unlike the HTF system in [18], the pre-defined logical rule base of FIS depends on two state-variables, $e(t)$ and $i_c(t)$, that act as its inputs. The inclusion of $e(t)$ and $i_c(t)$ in the fuzzy rule-base increases the degrees-of-freedom of the controller. This feature allows for flexible controller design and an agile transition between the constituent controllers under parametric uncertainties. The information of error-derivative, provided by $i_c(t)$, in conjunction with the elaborate rule-base helps in minimizing the convergence-time during load transients, followed by damping of oscillations to allow for a smoother settling [24]. Hence, this phenomenon enhances the time-optimal control effort of the system as well as its robustness against the bounded exogenous disturbances. The w is the output of FIS.

The functionality of the FIS is presented as follows. First of all, the input signals are normalized between -1 and 1. The normalized input signals are fuzzified into linguistic variables. The linguistic variables of input membership functions (MF), for e and i_c , are defined as; Negative-Big (NB), Negative-Medium (NM), Negative-Small (NS), Zero (Z), Positive-Small (PS), Positive-Medium (PM), and Positive-Big (PB). Hence, there a total of 49 rules. The linguistic variables of the output MF are; Zero (Z), Small (S), Medium (M), Big (B), and Very-Big (VB). The rule-base is presented in Table 2. After the fuzzification of the inputs, the implication method is applied followed by the aggregation of the fuzzy outputs using the maximum-minimum fuzzy inference technique, as shown in (21).

$$\mu_i = \min(f_i(\hat{e}), f_i(\hat{i}_c)) \quad (21)$$

where, μ is the grade-value of MF, $f(x)$ is the triangular MF, i is the number of rule, and $\hat{e}$ and $\hat{i}_c$ are the normalized values of e and i_c , respectively. The triangular MF, $f(x)$, is given by (22).

$$f_i(x) = \begin{cases} 1 + \frac{x - c_i}{b_i^-}, & -b_i^- \leq x - c_i \leq 0 \\ 1 - \frac{x - c_i}{b_i^+}, & 0 \leq x - c_i \leq b_i^+ \\ 0, & \text{otherwise} \end{cases} \quad (22)$$

where, b_i^- , c_i , and b_i^+ are the left-half width, center, and right-half width of the MF, respectively. The waveforms of the input and output MF are shown in Fig. 6 and 7, respectively.

Table 2. Fuzzy Rule-Base for Adaptive Self-tuning of Weightage

w		i_c						
		NB	NM	NS	Z	PS	PM	PB
e	NB	VB	VB	B	B	B	VB	VB
	NM	VB	B	M	M	M	B	VB
	NS	B	B	M	S	M	B	B
	Z	B	M	S	S	S	M	B
	PS	B	B	M	S	M	B	B
	PM	VB	B	M	M	B	B	VB
	PB	VB	VB	B	B	B	VB	VB

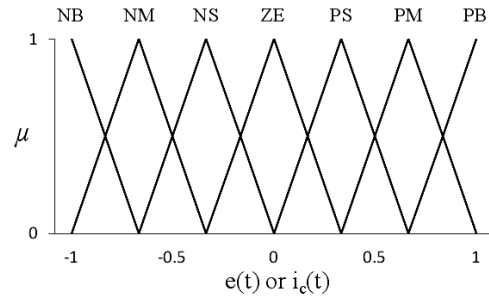


Fig. 6. Waveform of input MF

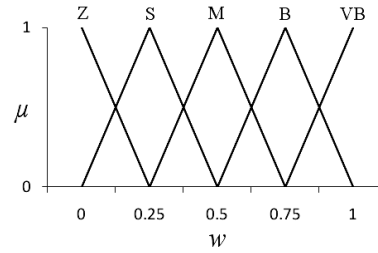


Fig. 7. Waveform of output MF

The aggregated output fuzzy set is de-fuzzified to compute a crisp value of w via the center-of-area method, given by (23).

$$w = \frac{\sum_{i=1}^n \mu_i \hat{c}_i}{\sum \mu_i} \quad (23)$$

where, $\hat{c}_i$ is the center of output MF. The block diagram of Fuzzy Adaptive Weighted Controller is shown in Fig. 8.

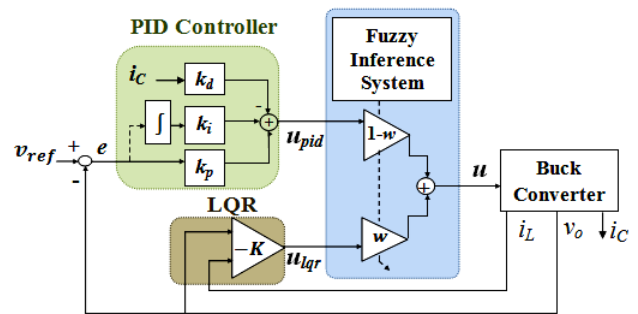


Fig. 8. Fuzzy Adaptive Weighted Controller (FAWC)

5. TESTS AND RESULTS

The voltage regulation performance of the PID controller, LQR, f-WC, HAWC, and FAWC is tested via the following two ‘hardware-in-the-loop’ experiments.

Test A: The performance of the controllers in tracking a step-reference of 5.0 V is tested. The resulting variations in v_o , exhibited by each controller, are shown in Fig. 9, 10, 11, 12, and 13. The corresponding variations in w , in HAWC and FAWC, are shown in Fig. 14 and 15, respectively.

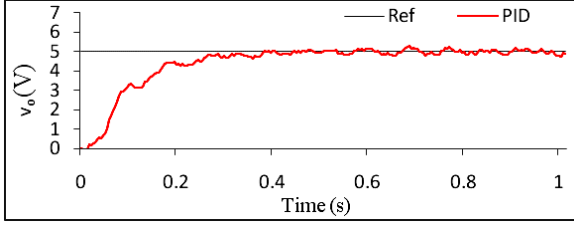


Fig. 9. Step-response exhibited by PID controller

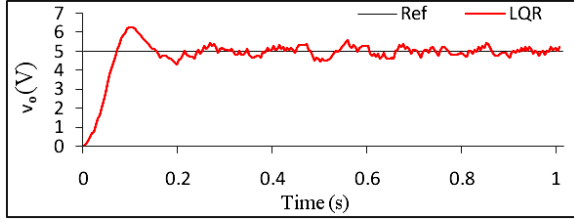


Fig. 10. Step-response exhibited by LQR

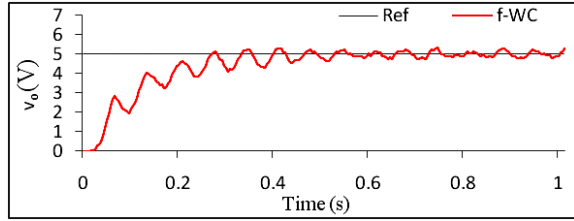


Fig. 11. Step-response exhibited by f-WC

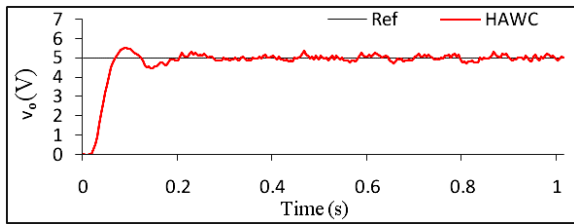


Fig. 12. Step-response exhibited by HAWC

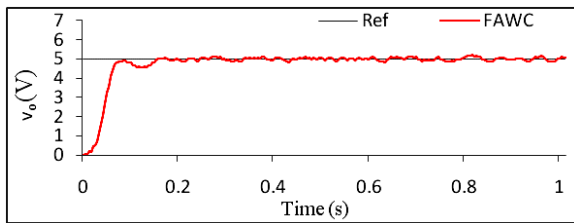


Fig. 13. Step-response exhibited by FAWC

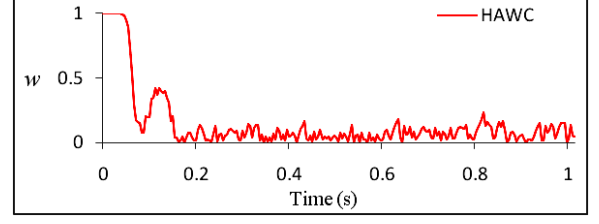


Fig. 14. Variations in ‘w’ for HAWC

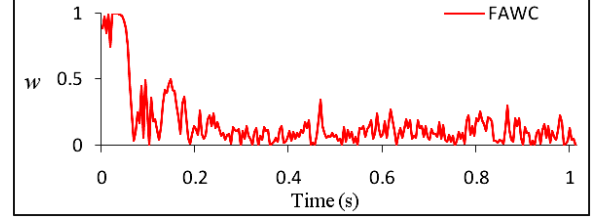


Fig. 15. Variations in ‘w’ for FAWC

Test B: The voltage regulation performance of each controller is tested under load disturbance conditions. The load-step transient resembles the impulsive disturbance and modeling error (since R_L is part of the buck converter’s state-space representation). The load-step transient is introduced by turning-off the transistor Q at $t \approx 1.25$ s. The abrupt variations exhibited by v_o for a 100% step increment in R_L are illustrated in Fig. 16, 17, 18, 19, and 20. The corresponding variations in the w for HAWC and FAWC are shown in Fig. 21 and 22, respectively.

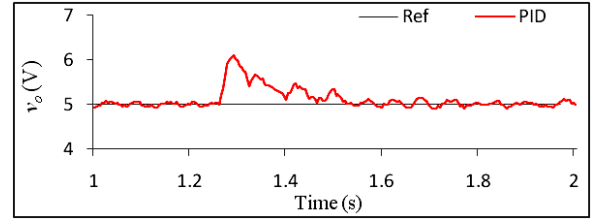


Fig. 16. Response of PID controller under load-step transient

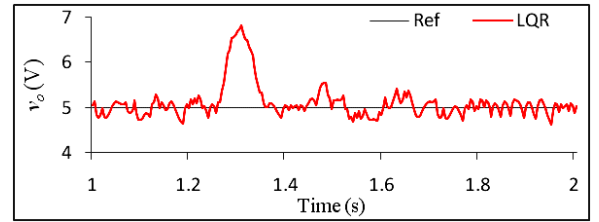


Fig. 17. Response of LQR under load-step transient

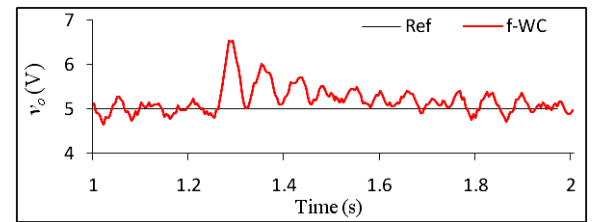


Fig. 18. Response of f-WC under load-step transient

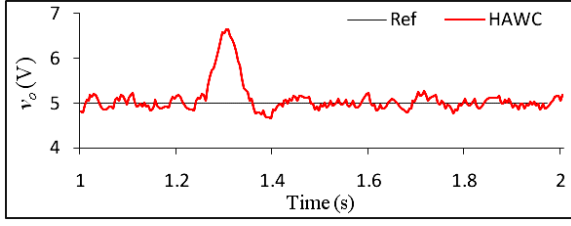


Fig. 19. Response of HAWC under load-step transient

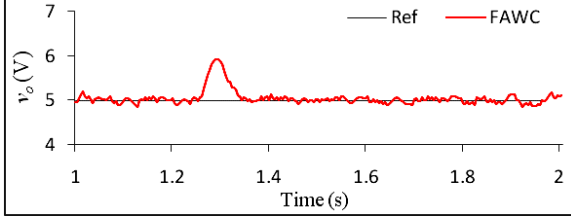


Fig. 20. Response of FAWC under load-step transient

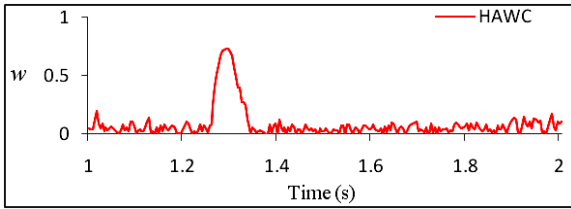


Fig. 21. Variations in 'w' for HAWC under load-step transient

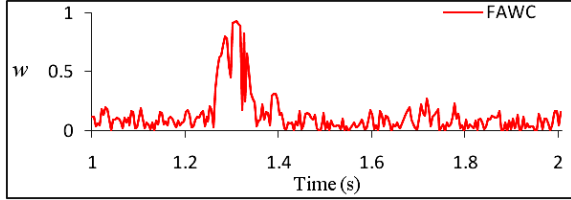


Fig. 22. Variations in 'w' for FAWC under load-step transient

The performance assessment of the results of Test-A and Test-B is summarized in Table 3. The FAWC surpasses other controllers by demonstrating significant improvement in the overall performance of the system. It exhibits the most time-optimal control effort while tracking the step-reference. It exhibits superior disturbance-rejection capability by quickly converging to the reference, while rendering minimum overshoot and negligible oscillations. In both testing scenarios, the variation behavior of 'w' clearly shows that FAWC is more responsive to the changes in system dynamics than HAWC. Hence, the proposed FAWC demonstrates considerable improvement over the HAWC scheme implemented in [17]. Since w is fixed, the f-WC does not deliver the most optimal overall control effort. It commutes abruptly between the two constituent controllers which leads to persistent oscillations in the response as it converges to v_{ref} .

Table 3. Summary of test results

Controller	Test A				Test B	
	t_r (s)	OS (V)	t_s (s)	E_{ss} (V)	OS (V)	t_s (s)
PID	0.28	0.02	0.45	0.10	1.10	0.26
LQR	0.11	1.27	0.25	0.28	1.82	0.12
f-WC	0.24	0.08	0.51	0.22	1.53	0.41
HAWC	0.10	0.54	0.20	0.18	1.65	0.18
FAWC	0.11	0.01	0.18	0.09	0.98	0.08

t_r = rise-time, OS = over-shoot, t_s = settling-time, E_{ss} = root-mean-squared steady-state error

6. CONCLUSION

This paper address the importance of computational intelligence to improve the control performance and compensate the nonlinearities associated with a complex dynamical system, using conventional linear feedback controller. Different collaborative control mechanisms are investigated to optimize the voltage regulation of a buck converter. The LQR delivers optimal control effort by minimizing the deviations in state-trajectories and control effort. The PID controller renders improved reference tracking capability and damps the steady-state errors caused by the parametric uncertainties and parasitic impedances. The LQR and PID controller work together to yield improved robustness against exogenous disturbances, convergence-rate, and time-domain response. Each controller variant is tested in real-time. The comparative assessment of the test-results clearly validates the superior performance of the FAWC over other controllers. Unlike the HTF technique, the fuzzy inference system in FAWC uses two state-variables, e and ic , to appropriately update w . This increases the degrees-of-freedom and flexibility of the mechanism to generate optimal control commands. In future, other intelligent self-tuning mechanisms can be investigated to further enhance the performance of weighted adaptive control schemes for DC-DC converters. Different soft-computing, meta-heuristic, and evolutionary optimization algorithms can be investigated to select the M and R weighting matrices of the quadratic cost function. The proposed adaptive control mechanism can also be tested on other DC-DC switch-mode power electronic converters; such as, boost, buck-boost, flyback, Cuk, Sepic, or Zeta converters. The Linear-Quadratic-Gaussian controllers and Fractional-Order PID controllers also can be intelligently combined and investigated to further enhance the robustness and time-optimality of the system's reference-tracking response and load-transient compensation.

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