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2018

Jiang, W., Chincholkar, S. H., & Chan, C.-Y. (2018). Modified voltage-mode controller for the quadratic boost converter with improved output performance. IET Power Electronics, 11(14), 2222-2231. doi:10.1049/iet-pel.2018.5037

<https://hdl.handle.net/10356/104276>

<https://doi.org/10.1049/iet-pel.2018.5037>

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A Modified Voltage-Mode Controller for the Quadratic Boost Converter with Improved Output Performance

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Abstract: The control of the quadratic boost converter with unknown load resistance and where only the output voltage is used for feedback purposes is addressed. In the existing voltage-mode controlled converter system, there exists a performance “trade-off” between the transient responses after the onset of a reference input and a load disturbance due to the usage of the traditional PI control scheme. To overcome these problems, a modified voltage-mode controller is proposed. The proposed controller uses a normalized integral action in which the derivative of the integrand is bounded by a user defined constant. The approximate stability analysis of the resultant voltage-mode controlled quadratic boost converter system is carried out to gain some insight into its closed-loop behaviour. Some simulation and experimental results comparing the performance of the proposed controller with that of the existing controller are also provided to show the improvements obtained using the proposed controller.

1 Introduction

The traditional dc-dc boost converter has been widely used to increase the voltage gain in many applications, e.g., renewable energy sources, electric vehicles and high-intensity-discharge lamps for automobiles [1] – [8]. However, theoretically, it can only provide a high voltage gain at some extremely high values of the duty ratio. The operation at high value of the duty ratio results in a low conversion efficiency as well as large voltage and current stresses in the power switch. Moreover, in practical applications, the maximum voltage gain of the traditional boost converter is very limited due to the existence of parasitic resistances [6]. Therefore, the traditional boost converter is not so suitable for high output voltage demand applications. To achieve a high voltage gain with reasonable duty ratios and high conversion efficiency, various high-gain boost converters have been explored in the past few years [9] – [16]. They can be separated as isolated and non-isolated converters. The isolated converters utilize high frequency transformers. The voltage gain can be easily tuned by changing the coil turns of the transformers. However, the leakage inductor problem dramatically affects the overall power conversion efficiency. Moreover, if the isolation property is not required in the applications, the use of isolated converters only reduces the power density and increases the cost of the converter system [17]. In order to achieve high voltage gain without using the transformer, advanced circuitry techniques such as cascaded structure [11], multiplier diode-capacitor circuit [12] – [13], switched capacitor cell [14] and voltage lift [15] – [16] have been employed to develop novel transformer-less boost converters. As such, these converters can step up the output voltage to a high level without affecting the conversion efficiency. They also provide a high power density, low voltage and current stresses on the switch components.

Recently, the regulation aspect of transformer-less boost converters has attracted quite a lot of attention [18] – [21]. In contrast to the buck converters, these boost converters present non-minimum phase dynamic characteristic which make it difficult to design the controller using only output

voltage for feedback purposes [18]. Therefore, the two-loop current mode controllers are usually used for the regulation of the boost converter, in which the converter output voltage is regulated via the control of the inductor current [17] – [20]. This scheme provides an improved settling time and a wider control bandwidth [21]. However, the structures of some of the existing current-mode controllers (such as those used in [18] – [19]) are such that it requires the knowledge of the nominal value of the load resistance to compute the control signal. This makes them unsuitable in practical applications where the value of the load resistance is unknown. To overcome this problem, an adaptive current-mode control law has been proposed [17]. In this control scheme, a real-time estimate of the load conductance was used to generate the control signal.

Because of the variable structure nature of the boost converters, the sliding-mode controllers (SMCs) are widely used in their regulation [22] – [27]. The SMCs provide the converter system robustness against external disturbances such as load changes and line voltage variations [22]. In [23] – [24], the hysteresis-based SMC is used to regulate the output voltage of the high-order boost converters. Even though this controller is easy to implement and provides a robust response over a wide range of operating conditions [23], it requires the use of a variable switching frequency which results in an electromagnetic-interference (EMI) problem as well as the excessive switching losses [25]. These drawbacks of hysteresis-based SMCs can be avoided in SMCs by using pulse-width modulation (PWM) technique [22], [25]. However, the implementation of these SMCs usually demand more state variables for feedback purposes. This incurs the additional cost and complexity due to more number of sensors. Also, due to the multi-current sensors requirement, such a controller is unsuitable for applications where power density limitations cannot accommodate the current sensor [26] – [27].

Recently, some voltage-mode controllers have been proposed for high-order boost converters [28] – [29]. These controllers are based on the simplified parallel-damped passivity-based approach proposed in [30]. The main advantage of these controllers is that they require only the

output voltage measurement for feedback purposes. However, due to the use of the traditional proportional-integral (PI) approach in the controllers of [28] – [29], there exists a “trade-off” between the performances of the transient responses after the onset of reference input and load disturbances. Besides, due to the structure of the controller presented in [28] – [29], there is a risk of division by zero in the control signal. Hence, there still exist many problems in the regulation of high-order boost-type converters that need to be addressed.

In this paper, the voltage-mode control of the quadratic boost converter with unknown load resistance is addressed. The main contribution of this paper is that a modified structure of the existing voltage-mode control law is proposed which overcomes the problems of the existing voltage-mode controllers. Instead of using the traditional integral action based on the output voltage error, the proposed controller uses an integral action which is based on the normalized output voltage error. Also, the integrand is bounded by a user-defined constant. This avoid the extreme changes in the control signal. As a result, the output performance of the controlled converter system is improved and the “trade-off” problem mentioned previously can be alleviated. The stability analysis of the proposed voltage-mode controlled quadratic boost converter system is carried out and the feasibility of the proposed controller is verified. Simulation and experimental results showing the effectiveness of the proposed controller for the quadratic boost dc-dc converter under various operation conditions are also provided.

2 Model of the quadratic boost converter

As compared to the traditional boost converter, the quadratic boost converter shown in Fig. 1 can provide a higher voltage gain using a smaller duty cycle. It also has a better trade-off between efficiency and the operation range of the duty cycle as compared to the cubic-type counterpart [31].

Fig. 1 Quadratic boost converter

The averaged state-space model of the quadratic boost converter operating in the continuous current mode (CCM) is given by [28]:

$$\frac{di_{L1}}{dt} = -\frac{v_C}{L_1}(1-u) + \frac{E}{L_1} \quad (1)$$

$$\frac{di_{L2}}{dt} = \frac{v_C}{L_2} - \frac{v_{Co}}{L_2}(1-u) \quad (2)$$

$$\frac{dv_C}{dt} = \frac{i_{L1}}{C}(1-u) - \frac{i_{L2}}{C} \quad (3)$$

$$\frac{dv_{Co}}{dt} = \frac{i_{L2}}{C_o}(1-u) - \frac{v_{Co}}{RC_o} \quad (4)$$

where i_{L1} , i_{L2} , v_C and v_{Co} represent the average current of inductor L_1 , the average current of inductor L_2 , the average voltage of capacitor C and the average voltage of the output capacitor C_o , respectively. In addition, E and R are the input voltage and load resistance, respectively, and u is the control signal, where $u \in (0,1)$.

Setting (1) – (4) equal to zero, the equilibrium values of the state variables and control signal can be obtained as

$$\begin{aligned} i_{L1\infty} &= \frac{V_d^2}{RE}, & i_{L2\infty} &= \sqrt{\frac{V_d^3}{R^2E}}, & v_{C\infty} &= \sqrt{EV_d}, \\ v_{Co\infty} &= V_d, & u_{\infty} &= 1 - \sqrt{\frac{E}{V_d}} \end{aligned} \quad (5)$$

where $i_{L1\infty}$, $i_{L2\infty}$, $v_{C\infty}$, $v_{Co\infty}$ and u_{∞} are the steady-state values of i_{L1} , i_{L2} , v_C , v_{Co} and u , respectively. The symbol of V_d represents the reference of the converter output voltage.

3 Proposed voltage-mode controller

The structure of the existing voltage-mode controller for the quadratic boost converter is first given. It is described by [28]:

$$u = 1 - \frac{\sqrt{E + K_{Pe}(v_{Co} - V_d) + K_{Ie} \int (v_{Co}(t) - V_d) dt}}{\sqrt{x_d}} \quad (6)$$

$$\frac{dx_d}{dt} = \frac{1}{C_o} [K_{1e}(v_{Co} - x_d) + K_{2e}(V_d - x_d)] \quad (7)$$

where K_{Pe} , K_{Ie} , K_{1e} and K_{2e} are positive controller gains and x_d is an artificial voltage variable. This controller structure is motivated by the structure of the simplified parallel-damped passivity-based controller (PBC) for the traditional boost converter [30]. The structure given by (6) – (7) is derived based on the expression of the equilibrium value of the averaged duty ratio, i.e., $u_{\infty} = 1 - \sqrt{E/V_d}$ (given in (5)) like what was done in [30] for the traditional boost converter. Also, additional integral and proportional actions are incorporated in (6) for improved performance of the controlled system. Although the control law (6) – (7) is able to successfully regulate the quadratic boost converter in the presence of load and line variations [28], there still exist a disadvantage in this control law. In practical applications, the traditional integral action is necessary to reduce the steady-state error in the converter output voltage. Now, since the integrand of the traditional integral action can be very large, there exists a “trade-off” between the desired performances of the output responses in the presence of reference voltage and load variations [17]. Specifically, the use of a large integral gain results in a short settling time and small variation in the output response after the onset of a load disturbance. However, this also leads to an increased overshoot as well as more oscillations in the converter output response in the presence of reference voltage variations. If a small integral gain is used to avoid this problem, the overshoot is suppressed but the settling time of the output response increases.

In order to overcome these disadvantages, a modified voltage-mode controller is proposed in this section. The stability analysis of the resultant voltage-mode controlled system and the feasibility of the controller are also presented.

3.1 Proposed control law

The proposed voltage-mode controller for the quadratic boost converter is described by:

$$u = \frac{\sqrt{z_d} - \sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta}}{\sqrt{V_d}} \quad (8)$$

$$\frac{d\theta}{dt} = -\frac{2\alpha h(v_{C_o} - V_d)}{1 + [\alpha(v_{C_o} - V_d)]^2} \quad (9)$$

$$\frac{dz_d}{dt} = \frac{1}{C_o} [-(K_1 + K_2)z_d + K_1 v_{C_o} + K_2 V_d] \quad (10)$$

where K_P , K_I , K_1 , K_2 , α and h are the positive controller gains defined by the user and z_d is an artificial voltage variable. It can be seen that the denominator of the proposed control law (8) is now a predefined positive constant. Hence, the risk of division by zero in the existing control law (6) is avoided. Also, an integral action based on the normalized voltage error in [17] is adopted in the proposed controller. The integrand is bounded as is shown in the following.

For simplicity, the symbol of n is used to represent (9), that is

$$n = \frac{d\theta}{dt} = -\frac{2\alpha h(v_{C_o} - V_d)}{1 + \alpha^2(v_{C_o} - V_d)^2} \quad (11)$$

The first- and second-order derivatives of (11) with respect to the output voltage error ($v_{C_o} - V_d$) can be derived as:

$$\frac{dn}{dv_e} = -\frac{2\alpha h(1 - \alpha^2 v_e^2)}{(1 + \alpha^2 v_e^2)^2} \quad (12)$$

$$\frac{d^2 n}{dv_e^2} = \frac{4\alpha h(1 - \alpha^2 v_e^2)}{(1 + \alpha^2 v_e^2)^3} + \frac{4\alpha^3 h v_e}{(1 + \alpha^2 v_e^2)^2} \quad (13)$$

where $v_e = (v_{C_o} - V_d)$. By setting (12) to zero, the corresponding inflection point $v_e = \pm \frac{1}{\alpha}$ can be obtained. Then, substituting $v_e = \pm \frac{1}{\alpha}$ into (11) and (13), $\frac{d^2 z}{dv_e^2} < 0$ results when $n = h$ and $\frac{d^2 z}{dv_e^2} > 0$ results when $n = -h$, respectively. Since v_e is the only dynamic variable in n , the absolute value of n , i.e., $|n| = \left|\frac{d\theta}{dt}\right|$, is upper bounded by the user defined constant h . As such, the value of θ will not be extremely large or small when the output voltage deviates from its reference value. This feature will improve the transient response of the output voltage.

3.2 Stability analysis

Substituting (8) – (10) into (1) – (4) yields the closed-loop equations given by

$$\frac{di_{L1}}{dt} = \frac{1}{L_1} (-v_C \bar{u} + E) \quad (14)$$

$$\frac{di_{L2}}{dt} = \frac{1}{L_2} (v_C - v_{C_o} \bar{u}) \quad (15)$$

$$\frac{dv_C}{dt} = \frac{1}{C} (i_{L1} \bar{u} - i_{L2}) \quad (16)$$

$$\frac{dv_{C_o}}{dt} = \frac{1}{RC_o} (Ri_{L2} \bar{u} - v_{C_o}) \quad (17)$$

$$\frac{dz_d}{dt} = \frac{1}{C_o} [-(K_1 + K_2)z_d + K_2 v_{C_o} + K_1 V_d] \quad (18)$$

$$\frac{d\theta}{dt} = -\frac{2\alpha h(v_{C_o} - V_d)}{1 + [\alpha(v_{C_o} - V_d)]^2} \quad (19)$$

where $\bar{u} = 1 - u = 1 - \frac{\sqrt{z_d} - \sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta}}{\sqrt{V_d}}$. By setting (14) – (19) to zero, the unique equilibrium point of the closed-loop dynamics is given by

$$(i_{L1\infty}, i_{L2\infty}, v_{C\infty}, v_{C_o\infty}, z_{d\infty}, \theta_\infty) = \left(\frac{V_d^2}{RE}, \sqrt{\frac{V_d^3}{R^2 E}}, \sqrt{EV_d}, V_d, V_d, 0 \right) \quad (20)$$

where $z_{d\infty}$ and θ_∞ are the steady-state values of z_d and θ , respectively.

Considering the high-order and complexity of the system given by (14) – (19), proving the large-signal stability analysis of the closed-loop system is rather difficult. Hence, the linearization method is used to perform an approximate stability analysis of the voltage-mode controlled system. This method is often used to study the stability of the high-order converter systems [22] – [23], [25].

Now, linearizing (14) – (19) about (20) and yields the following system of the form:

$$\dot{\tilde{x}} = A\tilde{x} \quad (21)$$

where $\tilde{x} = [\tilde{i}_{L1}, \tilde{i}_{L2}, \tilde{v}_C, \tilde{v}_{C_o}, \tilde{z}_d, \tilde{\theta}]^T$, $\tilde{i}_{L1} = i_{L1} - i_{L1\infty}$, $\tilde{i}_{L2} = i_{L2} - i_{L2\infty}$, $\tilde{v}_C = v_C - v_{C\infty}$, $\tilde{v}_{C_o} = v_{C_o} - V_d$, $\tilde{z}_d = z_d - z_{d\infty}$, $\tilde{\theta} = \theta - \theta_\infty$ and

$$A = \begin{bmatrix} 0 & 0 & -\frac{1}{L_1} \sqrt{\frac{E}{V_d}} & -\frac{K_P}{2L_1} & \frac{1}{2L_1} \sqrt{\frac{E}{V_d}} & -\frac{K_I}{2L_1} \\ 0 & 0 & \frac{1}{L_2} & -\frac{2E + K_P V_d}{2L_2 \sqrt{EV_d}} & \frac{1}{2L_2} & -\frac{K_I}{2L_2} \sqrt{\frac{V_d}{E}} \\ \frac{1}{C} \sqrt{\frac{E}{V_d}} & -\frac{1}{C} & 0 & \frac{K_P}{2CR} \sqrt{\frac{V_d}{E}} & -\frac{V_d}{2CER} & \frac{K_I}{2CR} \sqrt{\frac{V_d}{E}} \\ 0 & \frac{1}{C_o} \sqrt{\frac{E}{V_d}} & 0 & \frac{K_P V_d - 2E}{2C_o ER} & -\frac{1}{2C_o R} \sqrt{\frac{V_d}{E}} & \frac{K_I V_d}{2C_o ER} \\ 0 & 0 & 0 & \frac{K_2}{C_o} & -\frac{K_1 + K_2}{C_o} & 0 \\ 0 & 0 & 0 & 2\alpha & 0 & 0 \end{bmatrix}$$

Here, $h = 1$ is used for analysis purposes. The system (21) will be stable if all the eigenvalues of A have negative real parts [32], namely, all the roots of the system characteristic polynomial $P_A(s) = |sI - A| = 0$, where s is a complex variable, lie in the left-hand-side of the complex plane. Since $P_A(s)$ is of high order, it is not easy to use the analytical Routh-Hurwitz stability criterion to ensure system. Instead, the root-locus method [16] – [17] is adopted. In this method, only one controller gain, say, K_P is varied from its initial value until some roots of P_A leave the left half of the complex plane, while the other controller gains K_I , K_1 , K_2 and α are

Fig. 2 Root-locus plots of the controlled converter system

(a) For $0 < K_P < 2.5$, $\alpha K_I = 0.1$, $K_1 = 0.1$ and $K_2 = 0.1$, (b) For $K_P = 0.1$, $0 < \alpha K_I < 8$, $K_1 = 0.1$ and $K_2 = 0.1$, (c) For $K_P = 0.1$, $\alpha K_I = 0.1$, $0 < K_1 < 100$ and $K_2 = 0.1$, (d) For $K_P = 0.1$, $\alpha K_I = 0.1$, $K_1 = 0.1$ and $0 < K_2 < 5$.

fixed. Thus, the stability range of K_P can be obtained. Using the same procedure, the stability range of the other controller gains for a specific operation condition can also be obtained.

For the purpose of illustration, consider the following converter circuit parameter values:

$$\begin{aligned} L_1 &= 680 \mu\text{H}, L_2 = 560 \mu\text{H}, C = C_o = 470 \mu\text{H}, \\ R &= 470 \Omega, E = 10 \text{ V}, V_d = 70 \text{ V} \end{aligned} \quad (22)$$

Substituting (22) into (21) gives the following characteristic polynomial P_A :

$$\begin{aligned} P_A(s) &= s^6 + m_5 s^5 + m_4 s^4 + m_3 s^3 \\ &\quad + m_2 s^2 + m_1 s + m_0 \end{aligned} \quad (23)$$

where

$$m_0 = 7.23 \times 10^{15} (K_1 + K_2) \alpha K_I,$$

$$\begin{aligned} m_1 &= 5.16 \times 10^{14} K_1 - 8.5 \times 10^{14} K_2 \\ &\quad + 3.61 \times 10^{15} (K_1 + K_2) K_P + 3.4 \times 10^{12} \alpha K_I \\ &\quad - 5.42 \times 10^{11} (K_1 + K_2) \alpha K_I, \end{aligned}$$

$$\begin{aligned} m_2 &= 4.1 \times 10^{10} K_1 + 1.43 \times 10^{11} K_2, \\ &\quad + [1.7 \times 10^{12} - 2.71 \times 10^8 (K_1 + K_2)] K_P \\ &\quad - 2.55 \times 10^8 \alpha K_I + 8.1 \times 10^9 (K_1 + K_2) \alpha K_I \\ &\quad + 2.43 \times 10^{11}, \end{aligned}$$

$$\begin{aligned} m_3 &= 1.02 \times 10^{10} K_1 + 8.66 \times 10^9 K_2 \\ &\quad + [4.04 \times 10^9 (K_1 + K_2) - 1.27 \times 10^8] K_P \end{aligned}$$

$$\begin{aligned} &+ 3.8 \times 10^6 \alpha K_I - 67422 (K_1 + K_2) \alpha K_I \\ &+ 1.92 \times 10^7, \end{aligned}$$

$$\begin{aligned} m_4 &= 9631.8 K_1 + 22373 K_2 \\ &\quad + [1.9 \times 10^6 - 33711 (K_1 + K_2)] K_P \\ &\quad - 31.69 \alpha K_I + 4.79 \times 10^6, \end{aligned}$$

$$m_5 = 2127.7 (K_1 + K_2) - 15.84 K_P + 4.52$$

Fig. 2 shows the roots locus of (23) for various gains. To avoid the tedious calculations, the term of αK_I is treated as one variable. In Fig. 2(a), the controller gains used are $0 < K_P < 2.5$, $\alpha K_I = 0.1$, $K_1 = 0.1$ and $K_2 = 0.1$. The arrows indicate the movements of the poles with increasing K_P . It can be seen that two poles are in the right-half complex plane initially. As K_P is increased, the right-half complex plane poles move towards the imaginary axis and the left-half complex plane at $K_P = 0.047$. However, when $K_P = 1.42$, two dominant poles enter and stay in the right-half complex plane. Hence, the system is stable for $0.047 < K_P < 1.42$. Fig. 2(b) shows the root locus of (23) for $K_P = 0.1$, $0 < \alpha K_I < 8$, $K_1 = 0.1$ and $K_2 = 0.1$. In contrast to Fig. 2(a), initially, all the poles shown in Fig. 2(b) are located at the left-half complex plane, and four of them move toward to the imaginary axis as αK_I is increased. Since two of the poles enter the right-half plane at $\alpha K_I = 5.85$, the stability range of K_I for the converter system is $0 < \alpha K_I < 5.85$. Figs. 2(c) and 2(d) show the root loci for $K_P = 0.1$, $\alpha K_I = 0.1$, $0 < K_1 < 100$, $K_2 = 0.1$ and $K_P = 0.1$, $\alpha K_I = 0.1$, $K_1 = 0.1$, $0 < K_2 < 5$, respectively. The system is stable for $0.06 < K_1 < 100$ and $0 < K_2 < 0.18$.

3.3 Feasibility of the proposed controller

Next, the feasibility of the proposed voltage-mode controller described by (8) – (10) is addressed. Equation (8) can be rewritten as:

$$z_d = \left(\sqrt{V_d}u + \sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta} \right)^2 \quad (24)$$

The time derivative of the control signal u in (8) is given by,

$$\dot{u} = \frac{1}{2\sqrt{V_d}} \left(\frac{\dot{z}_d}{\sqrt{z_d}} - \frac{K_P\dot{v}_{C_o} + K_I\dot{\theta}}{\sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta}} \right) \quad (25)$$

where $\dot{z}_d$, $\dot{v}_{C_o}$ and $\dot{\theta}$ are the first-order time derivative of z_d , v_{C_o} and θ , respectively. Now, substituting (4), (18), (19) and (24) into (25) yields (26) (given at the bottom of this page). Setting the state-variables i_{L_2} , v_{C_o} and θ equal to their equilibrium values $i_{L_2\infty}$, $v_{C_o\infty}$ and θ_{∞} , respectively, gives the following remaining dynamics of the closed-loop system:

$$\dot{u} = \frac{1}{2\sqrt{V_d}} \left\{ \frac{(K_1 + K_2)}{C_o} \left[\frac{V_d}{\sqrt{V_d}u + \sqrt{E}} - (\sqrt{V_d}u + \sqrt{E}) \right] - \frac{K_P V_d [\sqrt{V_d}(1-u) - \sqrt{E}]}{R E C_o} \right\} \quad (27)$$

For the purpose of illustration, the circuit parameter values given in (22) are used here. Using (22) in (27) gives the phase-portrait of (27) as shown in Fig. 3.

Fig. 3 ‘Remaining dynamics’ for the voltage-controlled converter

It can be seen that $u_{\infty} = 1 - \sqrt{E/V_d}$ is the unique equilibrium point of the remaining dynamics and it is stable.

The key result of the paper is given in the following proposition:

Proposition: Given a reference voltage V_d , where $E < V_d < \infty$, the proposed control law (8) – (10) together with an appropriate set of controller gains, locally asymptotically stabilizes the averaged model of the quadratic boost converter (1) – (4) to its equilibrium point $(i_{L_1\infty}, i_{L_2\infty}, v_{C_o\infty}, v_{C_o\infty}) = \left(\frac{V_d^2}{RE}, \sqrt{\frac{V_d^3}{R^2 E}}, \sqrt{EV_d}, V_d \right)$ for $0 < R < \infty$.

Fig. 4. Block diagram of the closed-loop quadratic boost dc-dc converter system

$$\dot{u} = \frac{1}{2\sqrt{V_d}} \left\{ \frac{1}{C_o} \left[\frac{K_2 v_{C_o} + K_1 V_d}{\sqrt{V_d}u + \sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta}} - (K_1 + K_2) \left(\sqrt{V_d}u + \sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta} \right) \right] - \frac{[K_P R i_{L_2}(1-u) - v_{C_o}]/(R C_o) + \{2\alpha K_I(v_{C_o} - V_d)/\{1 + [\alpha(v_{C_o} - V_d)]^2\}\}}{\sqrt{E + K_P(v_{C_o} - V_d) + K_I\theta}} \right\} \quad (26)$$

Fig. 5 Output voltage responses of closed-loop converter system: (a) For varying K_p , (b) For varying αK_l , (c) For varying K_l , (d) For varying K_2 .

4 Simulation and experimental results

In this section, some results are provided to show the effectiveness of the proposed controller for regulation of the quadratic boost converter. The guidelines for selecting the controller gains to achieve the desired output response are also presented. For consistency sake, the same set of converter circuit parameter values in (22) were used in both simulation and experiments. The block diagram of the closed-loop converter system is shown in Fig. 4.

4.1 Tuning guidelines for the controller gains

The proposed controller has four adjustable controller gains, viz., K_p , αK_l , K_l and K_2 . It is not obvious how these controller gains can be selected to achieve the desired converter output response. In order to select the controller gains properly, the effects of the controller gains on the converter output responses is required to address. Initially, while ensuring that the regulated converter system is stable with a negligible steady-state error, the controller gains were set as $K_p = 0.12$, $\alpha K_l = 4$, $K_l = 0.2$ and $K_2 = 0.15$.

Fig. 5 shows the converter output responses for various controller gains. These simulation results were obtained from a controlled quadratic boost converter system built in MATLAB/SIMULINK. In these simulations, the value of the load resistance was changed from $R = 470 \Omega$ to $R = 235 \Omega$ at $t = 2$ s. Fig. 5(a) shows the converter output

responses for various values of K_p . It can be seen that increasing K_p reduces the overshoot, settling time and oscillations in the transient response after the onset of a load change. As such, a considerably large value of K_p can be first selected to achieve a fast response after the onset of load disturbances. Next, the converter output responses for different values of αK_l are given in Fig. 5(b). It is noted that increasing αK_l results in a smaller voltage variation and a faster convergence to the desired output voltage after the onset of a load change. As such, a sufficiently large value of αK_l can be chosen to further improve the output response after the onset of load disturbances. Fig. 5(c) shows the output voltage responses for various values of K_l . As the value of K_l increases, the overshoot and the settling time of the transient response increases. Therefore, after tuning the load change response using K_p and αK_l , a small value of K_l can be used to improve the start-up transient response. The output voltage response for varying K_2 is shown in Fig. 5(d).

Next, the effects of α and K_l on the output voltage response were investigated individually. The output voltage responses for a constant value of αK_l but varying values of α and K_l are shown in Fig. 6.

It is evident that for a constant αK_l , varying the values of α and K_l did not have much effect on the transient response in the presence of a load disturbance. But a large value of K_l (or a small value of α) resulted in a large overshoot during startup. In view of this, a small value of K_l can be employed

Fig. 6 Output voltage response for constant αK_I but varying α and varying K_I

to further improve the start-up transient response. This is the

Fig. 7 Output voltage responses of the regulated quadratic dc-dc boost converter (the solid blue line and the dashed red line are output responses of the quadratic boost converter obtained using the existing voltage-mode controller, while the dotted black line is the output response of the quadratic boost converter obtained using the proposed controller)

main feature of the proposed controller. Since the overshoot during the startup can be suppressed without affecting the response after the onset of a load disturbance, the aforementioned “trade-off” problem in the existing voltage-mode controllers can be overcome.

4.2 Comparison with existing voltage-mode controller

In order to show the advantages of the proposed controller over the existing voltage-mode controller, a comparative study involving the two controllers was carried out. The same set of circuit parameter values given by (22) was used. Also, the load resistance R was changed from 470Ω to 235Ω at $t = 2$ s.

Fig. 7 shows the output voltage responses of the quadratic boost converter using two controllers. The responses shown by the solid blue line and dashed red line are obtained using the existing voltage-mode controller (6) –

(7) for different values of the controller gains. It is evident that there exists a trade-off between the transient performances after the onset of a step reference input and the load disturbances. The output voltage response of the quadratic boost converter obtained using the proposed controller is shown by the dotted black line in Fig. 7. It is seen that after the onset of a reference input and a load resistance change, good output transient responses were achieved. The simulation results indicate that the proposed controller provides a better performance as compared to that obtained using the existing voltage mode controller.

4.3 Experimental results

In this section, some experimental results to show the effectiveness and advantages of the proposed controller in regulating the quadratic boost converter are provided. A laboratory prototype of the quadratic boost converter was built in which the circuit parameter values used were those in (22) and the switching frequency used was 30 kHz. Besides, the controller gains used in the experiments are the same as those given in the previous section, i.e., $K_P = 0.12$, $K_I = 40$, $\alpha = 0.1$, $K_1 = 0.2$ and $K_2 = 0.15$.

4.3.1 Using existing voltage-mode controller: Fig. 8 shows the output voltage responses of the quadratic boost converter obtained using the existing voltage mode controller (6) – (7). Fig. 8(a) shows the output voltage for a step change in the reference voltage V_d from 60 V to 70 V. The overshoot of the output voltage was relatively large, around 10% of the desired value of the output voltage. Besides, the settling time of the voltage response is around 0.15 s. Fig. 8(b) shows the output voltage response for an input voltage change from $E = 10$ V to $E = 7$ V. Again, the overshoot in the transient response was large, i.e., around 17% of the desired value of the output voltage, and the output response was restored to its desired value in 0.7 s. The output voltage responses in the presence of load resistance changes are shown in Figs. 8(c) and 8(d), where the output power P_{out} was changed from 10.4 W to 20.8 W in Fig. 8(c) and then back to 10.4 W in Fig. 8(d). In both figures, the output voltage deviation was around 8% of its nominal value.

4.3.2 Using proposed voltage-mode controller: Next, the performance of the proposed voltage-mode controller (8) – (10) was investigated. Fig. 9(a) shows the output response in the presence of a step change in the reference voltage V_d from 60 V to 70 V. As compared to the case of the existing voltage-mode controller, the output voltage was quickly restored to its desired reference value with a small overshoot (around 2% of the desired voltage) and a small settling time (smaller than 0.1 s). Fig. 9(b) shows the output response in the presence of an input voltage change. The overshoot is now smaller and the settling time of the output response is shorter. Figs. 9(c) and 9(d) show the output voltage responses in the presence of the output power disturbance mentioned in the preceding section. The deviations from the desired output voltage were now smaller.

The above results show that the proposed voltage-mode controller can successfully regulate the quadratic boost converter in the presence of load, input and reference voltage variations. Moreover, the experimental results are in

agreement with the analytical analyses and simulation results. It is evident from the results that the proposed

voltage-mode controller is superior to the existing voltage-mode controller.

Fig. 8 Output voltage v_o of the quadratic converter system using the existing voltage mode controller
(a) Output voltage response in the presence of reference voltage V_d changes from 60 V to 70 V
(b) Output voltage response in the presence of input voltage E changes from 10 V to 7 V
(c) Output voltage response in the presence of output power changes from 10.4 W to 20.8 W
(d) Output voltage response in the presence of output power changes from 20.8 W to 10.4 W

Fig. 9 Output voltage v_o of the quadratic converter system using the proposed controller.
a Output voltage response in the presence of reference voltage V_d changes from 60 V to 70 V
b Output voltage response in the presence of input voltage E changes from 10 V to 7 V
c Output voltage response in the presence of output power changes from 10.4 W to 20.8 W
d Output voltage response in the presence of output power changes from 20.8 W to 10.4 W

5 Conclusion

A modified voltage-mode control law for the regulation of the quadratic boost converter was proposed. The proposed controller employs an integral action based on the normalized output voltage error. Some simulation and experimental results were also provided and they confirm that the proposed control law is able to regulate the quadratic boost converter over a wide range of operating conditions and the control performance is considerably better as compared to that obtained using the existing voltage-mode controller.

6 References

- [1] He, X., Li, W.: 'Review of nonisolated high-step-up dc/dc converters in photovoltaic grid-connected applications', *IEEE Trans. Power Electron.*, 2011, **58**, (4), pp.1239-1250.
- [2] Wu, G., Ruan, X., Ye, Z.: 'Review of nonisolated high-step-up dc/dc converters in photovoltaic grid-connected applications', 2015, **62**, (1), pp. 383-393.
- [3] Ahrabi, R.R., Ardi, H., Elmi, M., Ajami, A.: 'A novel step-up multiinput dc-dc converter for hybrid electric vehicles application', *IEEE Trans. Power Electron.*, 2017, **32**, (5), pp. 3549-3561.
- [4] Saadat, P., Abbaszadeh, K.: 'A single-switch high step-up dc-dc converter based on quadratic boost', *IEEE Trans. Ind. Electron.*, 2016, **63**, (12), pp. 7733-7742.
- [5] Guo, F., Wen, C., Mao, J., and Song, Y.: 'Distributed Cooperative Secondary Control for Voltage Unbalance Compensation in an Islanded Microgrid', *IEEE Trans. Ind. Inf.*, 2015, **11**, (5), pp. 1087-1088.
- [6] Zhang, Y., Liu, J., Dong, Z., Wang, H., Liu, Y. F.: 'Dynamic performance improvement of diode-capacitor-based high step-up dc-dc converter through right-half-plane zero elimination', *IEEE Trans. Power Electron.*, 2017, **32**, (8), pp. 6532-6543.
- [7] Ismail, E.H., Al-Saffar, M. A., Sabzali, A.J., Fardoun, A.A.: 'A family of single-switch pwm converters with high step-up conversion ratio', *IEEE Trans. Circuits Syst. I, Reg. Papers*, 2008, **55**, (4), pp. 1159-1171.
- [8] Walker, G. R., Sernia, P. C.: 'Cascaded DC -DC converter connection of photovoltaic modules', *IEEE Trans. Power Electron.*, 2004, **19**, (4), pp. 1130 -1139.
- [9] Zeng, J., Qiao, W., Qu, L.: 'A single-switch LCL-resonant iso-lated DC-DC converter', *Proc. IEEE Energy Convers. Congr. Expo.*, USA, Sep. 2013, pp. 5496-5502.
- [10] Nguyen, M.-K., Lim, Y.-C., Choi, J.-H., Cho, G.-B.: 'Isolated high step-up dc-dc converter based on quasi-switched-boost network', *IEEE Trans. Ind. Electron.*, 2016, **63**, (12), pp. 7553-7562.
- [11] Nejad, M. L., Poorali, B., Adib, E., Birjandi, A. A. M.: 'New cascade boost converter with reduced losses', *IET Power Electron.*, vol. 9, no. 6, pp. 1213-1219, 2016.
- [12] Mayo-Maldonado, J. C., Salas-Cabrera, R., Rosas-Caro, J. C., Leon-Morales, D., Salas-Cabrera, E. et al., 'Modelling and control of a dc-dc multilevel boost converter', *IET Power Electron.*, 2011, **4**, (6), pp. 693-700.
- [13] Prabhala, V. A., Fajri, K. P. V., Gouribhatla, S. P., Baddipadiga, B. P., Ferdowsi, M.: 'A dc-dc converter with high voltage gain and two input boost stages', *IEEE Trans. Power Electron.*, 2016, **31**, (6), pp. 4206-4215.
- [14] Tan, S.-C., Bronstein, S., Nur, M., Lai, Y. -M., Ioinovici, A., Chi, K. T.: 'Variable structure modeling and design of switched-capacitor converters', *IEEE Trans. Circuits Syst. I, Reg. Papers*, 2009, **56**, (9), pp. 2132-2142.
- [15] Luo, F. L., Ye, H.: 'Super-lift boost converters', *IET Power Electron.*, 2014, **7**, (9), pp. 1655-1664.
- [16] Luo, F. L., Ye, H.: 'Positive output super-lift converters', *IEEE Trans. on Power Electron.*, 2003, **18**, (1), pp. 105-113.
- [17] Chan, C.Y., Chincholkar, S. H., Jiang, W.: 'Adaptive current-mode control of a high step-up dc-dc converter', *IEEE Trans. Power Electron.*, 2017, **32**, (9), pp. 7297-7305.
- [18] Chan, C.Y., Chincholkar, S. H.: 'Investigation of current-mode controlled cascade boost converter systems: dynamics and stability issues', *IET Power Electron.*, 2016, **9**, (5), pp. 911-920.
- [19] Chan, C.Y.: 'Comparative study of current-mode controllers for a high-order boost dc-dc converter', *IET Power Electron.*, 2014, **7**, (1), pp. 237 - 243.
- [20] Leyva-Ramos, J., Ortiz-Lopez, M., Diaz-Saldierna, L., Morales-Saldana, J.: 'Switching regulator using a quadratic boost converter for wide dc conversion ratios', *IET Power Electron.*, 2009, **2**, (5), pp. 605-613.
- [21] Tang, W. F., Lee, C., Ridley, R. B.: 'Small-signal modeling of average current-mode control', *IEEE Trans. on Power Electron.*, 1993, **8**, (2), pp. 112-119.
- [22] Chincholkar, S. H., Chan, C.-Y.: 'Design of fixed-frequency pulswidth-modulation-based sliding-mode controllers for the quadratic boost converter', *IEEE Trans. Circuits Syst. II, Exp. Briefs*, 2017, **64**, (1), pp. 51-55.
- [23] Lopez-Santos, O., Martinez-Salamero, L., Garcia, G., Valderrama-Blavi, H., Sierra-Polanco, T.: 'Robust sliding-mode control design for a voltage regulated quadratic boost converter', *IEEE Trans. Power Electron.*, 2015, **30**, (4), pp. 2313-2327.
- [24] Tan, S.-C., Lai, Y.-M., Chi, K. T.: 'General design issues of sliding-mode controllers in dc-dc converters', *IEEE Trans. Ind. Electron.*, 2008, **55**, (3), pp. 1160-1174.
- [25] Tan, S., Lai, Y.M., Tse, C. K.: 'A unified approach to the design of PWM-based sliding-mode voltage controllers for basic DC-DC converters in continuous conduction mode', *IEEE Trans. Circuits Syst. I, Reg. Pap.*, 2006, **53**, (13), pp. 1816-1827.
- [26] Tan, S., Lai, Y.M., Tse, C. K.: 'Indirect sliding mode control of power converters via double integral sliding surface', *IEEE Trans. Power Electron.*, 2008, **23**, (2), pp. 600-611.
- [27] Tan, S.-C., Lai, Y. M.: 'Constant-frequency reduced-state sliding-mode current controller for Cuk converters', *IET Power Electron.*, 2008, **1**, (4), pp. 466-477.
- [28] Chan, C.-Y.: 'Investigation of voltage-mode controller for cascade boost converter', *IET Power Electron.*, 2014, **7**, (8), pp. 2060-2068.
- [29] Chan, C.-Y.: 'Analysis and experimental study of an output feedback controller for a high-order boost dc-dc converter', *IET Power Electron.*, 2013, **6**, (7), pp. 1279-1287.
- [30] Chan, C.Y.: 'Simplified parallel-damped passivity-based controllers for dc-dc converters', *Automatica*, 2008, **44**, (11), pp. 2977-2980.
- [31] López-Santos, O. Martínez-Salamero, L., García, G., Valderrama-Blavi, H., Mercuri, D. O.: 'Efficiency analysis of a sliding-mode controlled quadratic boost converter', *IET Power Electron.*, 2013, **6**, (2), pp. 364-373.
- [32] Khalil, H.: 'Nonlinear Systems' (Pearson, NJ, USA, 2000, 3rd edn.).