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Abstract: A comparative study of two adaptive current-mode controllers for a high-order hybrid-type dc–dc boost converter is presented. The implementation of the traditional current-mode controller for this converter requires the knowledge of the nominal value of the load resistance to compute the control signal. As such, it is unable to handle systems with uncertain loads well. To address this, an adaptive law is used to estimate the load conductance in order to generate the reference current input. In this adaptive law, the derivative of the estimator is optimised as well as bounded. Moreover, the converter has two inductor currents which can be used for feedback purposes. Considering this, two adaptive current-mode controllers using the input and output inductor currents of the converter are separately designed to find the most appropriate inductor current for the implementation of the proposed controller. Finally, some simulation and experimental results comparing the performance of the adaptive controller using the output inductor current with that of the traditional current-mode controller are also presented.

1 Introduction

Theoretically, the conventional boost converter can produce a very high output voltage when its duty ratio approaches unity. However, operating at high values of the duty ratio is not desirable as it puts an unrealistic cap on the conduction times of the switching devices and leaves less room for control [1]. To address this drawback, many high-order boost-type dc–dc converters have been proposed in the past few years [1–10]. These include N -stage cascade boost converters [2], the quadratic converters based on a single active switch [3], converters with coupled inductors to achieve a high step-up voltage gain [4, 5], converters with active-clamp circuits [6, 7], high step-up dc–dc converters [8], dc–dc boost converter combined with voltage multiplier cell [9], and hybrid-type dc–dc converters [10].

The focus of the paper is on the hybrid-type dc–dc converters in which the switching capacitor/inductor structures are combined with the conventional converters like the boost converter and the buck–boost converter to achieve the high voltage transfer gain. This converter gives high-voltage gain and offers lower energy in the magnetic elements resulting in reduced weight, size and cost of the inductors and the overall power supply [10]. Despite such various advantages offered by these converters, there is a scarcity of works addressing their regulation problem.

Like most other boost-derived topologies, a hybrid-type dc–dc boost converter is a non-minimum phase system due to the presence of right-half plane zeroes in its control-to-output voltage transfer function. This transfer function, which is derived by linearising the converter model around its steady-state equilibrium point, makes it slightly difficult to design the controller of the converter using a single voltage-loop [11, 12]. To solve this problem, an indirect approach of control in which the output voltage is regulated via the inductor current control can be employed. In [13], the current-mode controller of the hybrid-type dc–dc boost converter has been addressed. Even though this control scheme offers ease of implementation and inherent overcurrent protection, it has a certain drawback. In this controller, an external current reference is required to compute the control signal. Since the value of this reference signal is calculated using the nominal value of the load resistance, the control law may not be used in applications where the value of the load resistance is unknown. In [14–17], several output feedback control laws for

some high-order dc–dc converters have been addressed. This kind of control law has certain advantages such as no need of current sensor and a good transient response over a wide range of operating conditions. However, its implementation is quite complex and the control law is not generic for all high-order dc–dc converter topologies.

Sliding-mode (SM) control is another popular non-linear control methodology for the dc–dc converters [18–25]. The implementation of the traditional hysteresis-modulation-based SM controller of several high-order dc–dc converters has been addressed in [18, 19], and it has various advantages such as its ease of implementation and robust performance against the load and line variations. However, the main drawback of this scheme is that it demands the variable switching frequency for its implementation which may lead to excessive switching losses, inductor losses and electromagnetic-interference generation [20]. To overcome these problems, the fixed-frequency pulse-width-modulation-based SM controller has been employed to regulate some high-order dc–dc converters, and it offers a good dynamic response over a wide range of operating conditions [20, 21]. However, it is difficult to achieve a good steady-state regulation using this approach if only a single integral term acting on the output voltage error is used in the sliding surface of these controllers [22]. If an additional double integral action is used [22–25] to alleviate this drawback, then the implementation of the controller requires more computations and becomes rather complex. In summary, to select an appropriate indirect controller for the regulation of the hybrid-type dc–dc boost converter is quite challenging and some further investigations are required to solve this problem.

In this paper, the regulation of the hybrid-type dc–dc boost converter using the adaptive current-mode controllers is investigated. This controller solves the problem of the traditional current-mode controller, in that it is unable to handle the systems with unknown loads, by employing an estimator of the load conductance to compute the reference inductor current. The value of this estimator is computed using an adaptive law whereby the derivative of the estimator is both optimised and bounded [26]. Moreover, for the hybrid-type dc–dc converter, two inductor currents could be used for feedback purposes. However, only one of them will be used to achieve output voltage regulation. The previous study has shown that the output inductor current is preferred for the traditional linear current-mode control of this

Fig. 1 Hybrid-type dc-dc boost converter and its operating stages
(a) Overall converter circuit topology, (b) Circuit configuration when switch 'S' is on, (c) Circuit configuration when switch 'S' is off

converter [13]. In this present work, two adaptive non-linear controllers using the input and output inductor currents are investigated to find the most suitable inductor current for the controller design. The stability analysis of the proposed adaptive current-mode controlled converter systems is also presented. Finally, some simulation and experimental results are provided to show the effectiveness of the proposed controller to regulate the hybrid-type high-order dc-dc converter.

2 State-space model of the hybrid-type dc-dc boost converter

Fig. 1 shows the circuit topology of the hybrid-type dc-dc boost converter. The operating principle of the converter is given in [10], and thus, will not be covered here.

Now, using $C_1 = C_2 = C$ and $x_4 = x_5$, the average state-space model of the converter operating in the continuous conduction mode can be obtained as [13]

$$\frac{d\bar{x}}{dt} = A\bar{x} + BE \quad (1)$$

where $\bar{x}^T = [\bar{x}_1 \ \bar{x}_2 \ \bar{x}_3 \ \bar{x}_4]$ represents the vector of averaged state variables of the system. Here, $\bar{x}_1$ and $\bar{x}_2$ represent the average value of the inductor currents flowing through L_1 and L_2 , respectively, and $\bar{x}_3$ and $\bar{x}_4$ are the average voltages across capacitors C_o and C_1 , respectively. Also, matrices A and B are given by

$$A = \begin{bmatrix} 0 & 0 & 0 & -\frac{1-u}{L_1} \\ 0 & 0 & -\frac{1}{L_2} & \frac{1+u}{L_2} \\ 0 & \frac{1}{C_o} & -\frac{1}{RC_o} & 0 \\ \frac{1-u}{2C} & -\frac{1+u}{2C} & 0 & 0 \end{bmatrix}, \quad B = \begin{bmatrix} \frac{1}{L_1} \\ 0 \\ 0 \\ 0 \end{bmatrix}$$

Here, the symbol u denotes the control signal of the switch 'S' such that $u \in [0, 1]$.

By setting (1) to zero, the following equilibrium values are obtained:

$$\begin{aligned} X_1 &= \frac{V_d^2}{RE}, \quad X_2 = \frac{V_d}{R}, \quad X_3 = V_d, \\ X_4 &= \frac{V_d + E}{2}, \quad U = \frac{V_d - E}{V_d + E} \end{aligned} \quad (2)$$

where X_1, X_2, X_3, X_4 and U denote the equilibrium values of the averaged state variables $\bar{x}_1, \bar{x}_2, \bar{x}_3, \bar{x}_4$ and u , respectively. The symbols E and V_d represent the input voltage and the reference converter output voltage, respectively.

3 Traditional current-mode control

The traditional current-mode controller (of the form used in [11–13]) is given first to demonstrate its shortcoming to regulate the systems with unknown loads.

In [13], a linear current-mode controller for the hybrid-type dc-dc boost converter has been proposed. The controller using the output inductor current for feedback purposes is given by

$$u = U - K_P(x_2 - X_2) - K_I \int (x_3(\tau) - V_d) d\tau \quad (3)$$

where K_P and K_I are the positive gains of the controller and $X_2 = V_d/R$ represents the reference value of the output inductor current. The main drawback of this controller is that it requires the nominal value R of the load resistance to compute the control signal. Therefore, this control law may not be applicable in the application where R is unknown. To solve this problem, the non-linear adaptive current-mode controller is proposed in the following section.

4 Adaptive current-mode control

In this section, the design of the non-linear adaptive current-mode controllers for the hybrid-type boost converter is presented.

4.1 Proposed adaptive control law

The adaptive current-mode control law for the hybrid-type boost converter is given by

$$u = U - K_c [x_i - \hat{X}_i(\hat{\theta})], \quad i = 1, 2 \quad (4)$$

where K_c is the gain of the adaptive controller, U is given by (2), x_i is the inductor current of the converter whose reference value $\hat{X}_i(\hat{\theta})$ is achieved using the estimator $\hat{\theta}$ of the load conductance. The estimator $\hat{\theta}$ is obtained using the adaptive law given by [26]

$$\frac{d\hat{\theta}}{dt} = -\frac{2\beta m e_3}{1 + \beta^2 e_3^2} \quad (5)$$

where β and m are the positive controller gains and $e_3 = x_3 - V_d$ denotes the output voltage error.

Setting $d\hat{\theta}/dt = h$, the first- and second-order time derivatives of h with respect to e_3 are:

$$\frac{dh}{de_3} = -\frac{2\beta m(1 - \beta^2 e_3^2)}{(1 + \beta^2 e_3^2)^2}, \quad (6a)$$

$$\frac{d^2h}{de_3^2} = \frac{4\beta m(1 - \beta^2 e_3^2)}{(1 + \beta^2 e_3^2)^3} + \frac{4\beta^3 m e_3}{(1 + \beta^2 e_3^2)^2} \quad (6b)$$

By setting (6a) to zero, the inflection points of (6) can be obtained as $e_3 = \pm(1/\beta)$. Substituting this into (5) and (6b), we get $d^2h/de_3^2 < 0$ when $h = m$ and $d^2h/de_3^2 > 0$ when $h = -m$. Thus, m and $-m$ present the global maximum and global minimum of h ,

respectively. Therefore, $\left| \frac{d\hat{\theta}}{dt} \right|$ is optimised and bounded by a user defined maximum value m .

4.2 Adaptive current-mode controller using input inductor current

Unlike the conventional boost converter, the hybrid-type dc–dc boost converter presents more than one inductor current for feedback purposes. Thus, when using current-mode control of the converter, it is necessary to select the most appropriate inductor current for the controller design. The choice of the inductor current not only determines the range of controller parameters to ensure system stability, but it also affects the dynamic response of the controlled converter [11–13]. Considering this, a detailed comparative study of two non-linear adaptive current-mode controllers using the input and output inductor currents of the converter has been conducted. The adaptive current-mode controller using the input inductor current for feedback purposes is first studied.

The control law using the input inductor current is given by

$$u = U - K_c [x_1 - \hat{X}_1] \quad (7)$$

where

$$\hat{X}_1 = \frac{\hat{\theta} V_d^2}{E} \quad (8)$$

Here, $\hat{X}_1$ is the estimated value of X_1 and $\hat{\theta}$ is obtained using (5). To gain an insight into the adaptive current-mode controlled system, the stability analysis is now provided.

The following errors are defined:

$$e_1 = \bar{x}_1 - X_1, \quad e_2 = \bar{x}_2 - X_2, \quad e_3 = \bar{x}_3 - X_3, \quad e_4 = \bar{x}_4 - X_4, \quad (9)$$

$$\tilde{\theta} = \hat{\theta} - \frac{1}{R}$$

Substituting (7)–(9) into (1) yields the error dynamics described by

$$\frac{de_1}{dt} = \frac{1}{L_1} [-(1 - u_{e1})e_1 - (1 - u_{e1})X_4 + E] \quad (10a)$$

$$\frac{de_2}{dt} = \frac{1}{L_2} [-e_3 + (1 - u_{e1})e_4 - X_3 + (1 - u_{e1})X_4] \quad (10b)$$

$$\frac{de_3}{dt} = \frac{1}{C_o} \left(e_2 - \frac{1}{R} e_3 \right) \quad (10c)$$

$$\frac{de_4}{dt} = \frac{1}{2C} [(1 - u_{e1})e_1 - (1 + u_{e1})e_2 + (1 - u_{e1})X_1 - (1 + u_{e1})X_2] \quad (10d)$$

$$\frac{d\tilde{\theta}}{dt} = -\frac{2\beta m e_3}{1 + \beta^2 e_3^2} \quad (10e)$$

where

$$u_{e1} = U - K_c \left(e_1 - \frac{\tilde{\theta} V_d^2}{E} \right).$$

The equilibrium point of (10) can be obtained as

$$e_{1\infty} = e_{2\infty} = e_{3\infty} = e_{4\infty} = \tilde{\theta}_{\infty} = 0 \quad (11)$$

Linearisation of (10) about the equilibrium point (11) yields the following linearised system:

$$\dot{z} = M_{in} z \quad (12)$$

where $z^T = [z_1 \ z_2 \ z_3 \ z_4 \ z_5]$, $z_i = e_i - e_{i\infty}$, $i = 1, \dots, 4$, $z_5 = \tilde{\theta} - \tilde{\theta}_{\infty}$ and (see (13)) The linearised system (12) will be stable if the coefficients of the characteristic polynomial $p_{in}(s) = |sI - M_{in}|$, where s is a complex variable, meet the Routh-Hurwitz stability criterion. For the purpose of illustration, consider the following set of circuit parameters:

$$E = 5 \text{ V}, \quad V_d = 25 \text{ V}, \quad L_1 = L_2 = 680 \mu\text{H}, \quad (14)$$

$$C_1 = C_2 = C_o = 220 \mu\text{F}, \quad R = 470 \Omega$$

Now, using these values into (13), the corresponding characteristic polynomial is given by

$$p_{in}(s) = s^5 + a_4 s^4 + a_3 s^3 + a_2 s^2 + a_1 s + a_0 \quad (15)$$

where (see equation below) and $K_a = \beta m$.

The Routh table for (15) is given by

$$\begin{array}{ccccc} s^5 & 1 & a_3 & a_1 & \\ s^4 & a_4 & a_2 & a_0 & \\ s^3 & b_1 & b_2 & 0 & \\ s^2 & c_1 & c_2 & 0 & \\ s^1 & d_1 & 0 & 0 & \\ s^0 & e_1 & 0 & 0 & \end{array} \quad (16)$$

where

$$b_1 = \frac{a_3 a_4 - a_2}{a_4}, \quad b_2 = \frac{a_1 a_4 - a_0}{a_4}, \quad c_1 = \frac{b_1 a_2 - b_2 a_1}{b_1},$$

$$c_2 = a_0, \quad d_1 = \frac{c_1 b_2 - c_2 b_1}{c_1}, \quad e_1 = a_0.$$

According to the Routh-Hurwitz stability criterion, system (12) is asymptotically stable if and only if all the coefficients of the characteristic polynomial (15), namely $a_i (i = 0, \dots, 4)$, and the first column of the Routh table (16), namely b_1, c_1, d_1 and e_1 , are greater than zero. Since all $a_i (i = 0, \dots, 4)$, b_1, c_1, d_1 and e_1 are functions of K_a and K_c , a system stability region can be determined in the $K_c - K_a$ plane. If K_a and K_c are selected in system stability region, the linearised system (12) is asymptotically stable. Solving $a_0 > 0$, $a_2 > 0$, $a_3 > 0$ and $a_4 > 0$ yields $K_c > 0$ and $K_a > 0$. Then, a second-degree relation between K_a and K_c , which is given as a hyperbola

$$M_{in} = \begin{bmatrix} -\frac{K_c X_4}{L_1} & 0 & 0 & -\frac{1-U}{L_1} & \frac{K_c X_4 V_d^2}{L_1 E} \\ -\frac{K_c X_4}{L_2} & 0 & -\frac{1}{L_2} & \frac{1+U}{L_2} & \frac{K_c X_4 V_d^2}{L_2 E} \\ 0 & \frac{1}{C_o} & -\frac{1}{RC_o} & 0 & 0 \\ \frac{(1-U) + K_c(X_1 + X_2)}{2C} & -\frac{1+U}{2C} & 0 & 0 & \frac{K_c(X_1 + X_2) V_d^2}{2CE} \\ 0 & 0 & -2\beta m & 0 & 0 \end{bmatrix} \quad (13)$$

Fig. 2 Stability region for system using adaptive current-mode controller based on input inductor current

in the $K_c - K_a$ plane, is obtained by solving $b_1 > 0$. Finally, by solving $c_1 > 0$ and $d_1 > 0$, two high-degree relations between K_a and K_c are also identified. The stability region of linearised system (12) using the circuit parameters given in (14) is shown in Fig. 2. The plots of $a_1 > 0$ and $c_1 > 0$ are quite far from the origin, and thus, are not shown in the figure. The shaded area represents the stability region. It is evident that, for the non-linear adaptive controller designed using the input inductor current, the ranges of the controller gains to ensure system stability are quite narrow.

4.3 Adaptive current-mode controller using output inductor current

This section presents the non-linear adaptive current-mode controller using the output inductor current x_2 for the feedback purpose. The adaptive control law is given by

$$u = U - K_c[x_2 - \hat{X}_2] \quad (17)$$

where

$$\hat{X}_2 = \hat{\theta}V_d \quad (18)$$

Here, $\hat{X}_2$ is the estimated value of X_2 . Like what was done previously in Section 4.2, substituting (9) and (17)–(18) into (1)

yields a set of dynamic equations with a unique equilibrium point (11). The corresponding linearised model has a coefficient matrix M_{out} given by

$$\dot{z} = M_{out}z \quad (19)$$

where (see equation below) Using the same set of circuit parameter values given in (14), the characteristic polynomial $p_{out}(s) = |sI - M_{out}|$ can be obtained as

$$p_{out}(s) = s^5 + \hat{a}_4s^4 + \hat{a}_3s^3 + \hat{a}_2s^2 + \hat{a}_1s + \hat{a}_0 \quad (20)$$

where (see equation below) and $K_a = \beta m$.

Following the same procedure used in the previous section, the stability region of (19) can be obtained and is shown in Fig. 3, where $\hat{b}_1$, $\hat{c}_1$ and $\hat{d}_1$ are the coefficients in the first column of the Routh table for (20). Again, the shaded area represents the stability ranges for K_c and K_a . Since K_c and K_a are positive, the conditions of $\hat{a}_0 > 0$, $\hat{a}_2 > 0$ and $\hat{a}_4 > 0$ are naturally achieved. As such, these conditions are not shown in Fig. 3.

It can be seen from Fig. 3 that the adaptive current-mode controller using the output inductor current leads to a wider range of controller gains to be used to give a stable system. This allows the designer to vary the controller gains over a wider range to achieve the desired output response. Therefore, the adaptive controller using the output inductor current should be preferred over the controller using the input inductor current.

4.4 Validation of results

In order to verify the theoretical conclusions obtained in Sections 4.2 and 4.3, some simulations were carried out using MATLAB. The output response of the adaptive current-mode controller using the input inductor current was compared with that of the adaptive controller using the output inductor current. The same set of circuit parameter values given in (14) was used to obtain the results.

Figs. 4a and b show the transient responses of the adaptive controller using the input inductor current. From Fig. 4a, it is seen that as the value of K_c increases, the oscillations in the transient response were reduced and the settling time of the response was much shorter. However, there is a limit to the maximum value of K_c which can be used to ensure system stability, and the response becomes unstable even for very small values of K_c (see Fig. 5b).

To solve this problem, the adaptive controller using the output inductor current was used. Figs. 4c and d show the transient output response obtained. It can be seen that even though there are

$$\begin{aligned} a_4 &= 2.21 \times 10^4 K_c + 9.67, & a_3 &= 5.69 \times 10^3 K_c + 1.63 \times 10^7 \\ a_2 &= 3.93 \times 10^{11} K_c + 2.51 \times 10^{10} K_c K_a + 9.34 \times 10^7 \\ a_1 &= 4.75 \times 10^{12} K_c - 2.02 \times 10^{12} K_c K_a + 2.48 \times 10^{12}, & a_0 &= 5.59 \times 10^{16} K_c K_a. \end{aligned}$$

$$M_{out} = \begin{bmatrix} 0 & -\frac{K_c X_4}{L_1} & 0 & -\frac{1-U}{L_1} & \frac{K_c X_4 V_d}{L_1} \\ 0 & -\frac{K_c X_4}{L_2} & -\frac{1}{L_2} & \frac{1+U}{L_2} & \frac{K_c X_4 V_d}{L_2} \\ 0 & \frac{1}{C_o} & -\frac{1}{RC_o} & 0 & 0 \\ \frac{1-U}{2C} & \frac{K_c(X_1 + X_2) - (1+U)}{2C} & 0 & 0 & -\frac{K_c(X_1 + X_2)V_d}{2C} \\ 0 & 0 & -2\beta m & 0 & 0 \end{bmatrix}$$

$$\begin{aligned} \hat{a}_4 &= 2.21 \times 10^4 K_c + 9.67, & \hat{a}_3 &= -1.56 \times 10^6 K_c + 1.63 \times 10^7 \\ \hat{a}_2 &= 4.91 \times 10^{10} K_c + 5.01 \times 10^9 K_c K_a + 9.34 \times 10^7 \\ \hat{a}_1 &= 4.91 \times 10^{11} K_c - 4.04 \times 10^{11} K_c K_a + 2.48 \times 10^{12}, & \hat{a}_0 &= 1.12 \times 10^{16} K_c K_a. \end{aligned}$$

oscillations for small values of K_c , these oscillations can be suppressed by increasing K_c to a sufficiently large value without losing the system stability. This is in agreement with theoretical conclusion that the range of controller gains to ensure system stability increases considerably when the output inductor current was used for feedback purposes. This confirms that the adaptive controller using the output inductor current should be preferred over the adaptive controller using the input inductor current.

5 Simulation and experimental results

In this section, simulation and experimental results are provided to show the effectiveness of the proposed controller using the output inductor current for the regulation of the hybrid-type high-order dc–dc boost converter. The following parameter values of the converter circuit were used in both simulations and experiments.

$$E = 5 \text{ V}, \quad V_d = 25 \text{ V}, \quad L_1 = L_2 = 680 \mu\text{H}, \quad C_1 = C_2 = C_o = 470 \mu\text{F}, \quad R = 950 \Omega$$

Fig. 3 Stability region for system using adaptive current-mode controller based on output inductor current

Fig. 4 Output responses of the controlled hybrid-type dc–dc boost converter

(a) Controller using input inductor current with $K_c = 0.0005$, $K_a = 0.1$ ($\beta = 0.1$ and $m = 1$) (red dotted) and $K_c = 0.0015$, $\beta = 0.1$ and $m = 1$ (blue), (b) Controller using input inductor current with $K_c = 0.005$, $K_a = 0.1$ ($\beta = 0.1$ and $m = 1$), (c) Controller using output inductor current with $K_c = 0.1$, $K_a = 0.1$ ($\beta = 0.1$ and $m = 1$), (d) Controller using output inductor current with $K_c = 2$, $K_a = 0.1$ ($\beta = 0.1$ and $m = 1$)

5.1 Simulation results

In order to show the merits of the proposed controller, a comparison study between the traditional current-mode controller (3) and the proposed controller (17) was carried out. In this comparison study, the value of the load resistance was changed from $R = 950 \Omega$ to $R = 470 \Omega$ at $t = 2 \text{ s}$ and it restored to $R = 950 \Omega$ at $t = 3 \text{ s}$.

Fig. 5 shows the output voltage responses of the controlled hybrid-type boost converter using different controllers. The dashed blue waveform shows the output response obtained using the traditional current-mode control while the solid red waveform shows the output response obtained using the proposed controller. It can be seen from Fig. 5a that when the small value of the integral gain K_1 of the traditional current-mode controller, i.e. $K_1 = 0.1$ was employed, both controllers provide similar control performance at the converter start-up stage. However, the proposed controller has better performance in the presence of load disturbances as compared to that obtained using the traditional current-mode controller. When the value of the integral gain K_1 was increased to $K_1 = 0.8$, it can be seen from Fig. 5b that even though both controller have similar performance in the presence of load disturbances, the proposed controller provides an output voltage response with a smaller overshoot and a shorter settling time at the converter start-up stage.

Considering both Figs. 5a and b, it is evident that when the converter is regulated by the traditional current-mode controller, there exists a ‘trade-off’ between the transient performances at the converter start-up stage and that after the onset of the load disturbances. However, this trade-off problem is avoided if the proposed controller is adopted. Both excellent transient response at the start-up stage and that after the onset of load disturbances can be achieved simultaneously. Hence, the proposed controller is more suitable for regulating the hybrid-type dc–dc boost converter than its traditional counterpart.

5.2 Experimental results

To verify the validity of the proposed adaptive controller to regulate the practical hybrid-type boost converter, a laboratory prototype of the closed-loop converter system was built. Also, to

Fig. 5 Output voltage responses of the regulated hybrid-type boost converter (the dashed blue line is the output response obtained using the traditional current-mode controller while the solid red line is the output response obtained using the proposed controller based on the output inductor current)

(a) $K_P = 0.2$, $K_I = 0.1$ for the traditional controller; $K_c = 0.05$, $\beta = 0.5$, $m = 0.12$ for the proposed controller, (b) $K_P = 0.2$, $K_I = 0.8$ for the traditional controller; $K_c = 0.05$, $\beta = 0.5$, $m = 0.12$ for the proposed controller

Fig. 6 Traditional current-mode control

(a) Output response in the presence of load changes for $K_P = 3$ and $K_I = 1.5$, (b) Output response in the presence of load changes for $K_P = 3$ and $K_I = 7$

compare the performance of the proposed adaptive controller with that of the traditional current-mode controller, the latter was also built. The controllers were implemented using analog components. More specifically, the LEM LTS-6NP current transducer was used to measure the inductor current. The division and square functions were achieved by using the combination of AD633 and AD711. The optical-coupler HCPL3140 was used to drive the MOSFET IRFP250 at a switching frequency of 20 kHz. In order to avoid the saturation problem in the analog implementations, the feedback gain of the output voltage was set as 0.1, and, correspondingly, the value of the reference output voltage V_d and that of input voltage E were also set at 0.1 times their nominal value.

5.2.1 Traditional current-mode controller: In this part, the regulation performance of the traditional current-mode controller (3) is presented. Figs. 6a and b show the output voltage responses in the presence of load changes for $K_P = 3$ and $K_I = 1.5$ and $K_P = 3$ and $K_I = 7$, respectively. The load resistance was changed from $R = 950 \Omega$ to $R = 470 \Omega$ and then back to $R = 950 \Omega$. As can be seen from Fig. 6a, the worst-case overshoot and settling time of the load change response were $\sim 16\%$ of V_d and ~ 1.2 s, respectively, when the traditional current-mode controller was employed. When the value of K_I was increased to improve the load change response, the overshoot and settling time in the transient output response were found to be $\sim 28\%$ and ~ 1.2 s, respectively (see Fig. 6b). As such, there was a trade-off between the qualities of the transient response and the load change response when traditional current-mode controller (3) was used.

5.2.2 Adaptive current-mode controller: Based on the conclusion of Section 4.3, the controller (17) using the output

inductor current was implemented, with $\beta = 1$, $m = 1$ and $K_c = 3.33$. Fig. 7a shows the output voltage start up response and the output response when the load resistance was changed from $R = 950 \Omega$ to $R = 470 \Omega$, and then back to $R = 950 \Omega$. As compared to the output responses given in Figs. 6a and b, the ‘trade-off’ problem was avoided and a faster output voltage response with a reduced overshoot was obtained for both the start-up response and the load change response. The settling time and overshoot of the start-up output voltage response were reduced to ~ 0.6 s and $\sim 4\%$ of V_d , respectively, and the settling time of the load change response was reduced to ~ 0.2 s when the proposed controller was employed. The output voltage response under the input voltage changes is described in Fig. 7b. Again, the output voltage was restored to its nominal value (25 V) after the onset of the input voltage changes with a considerably small value of variation. Fig. 7c shows the output voltage response when V_d was changed from 25 to 15 V and then back to 25 V. A good voltage tracking was achieved. These results show that the proposed adaptive current-mode controller is competent to regulate the high-order boost converter as well as providing a better performance when compared with the traditional current-mode controller. The experimental results are in good agreement with the simulation results.

6 Conclusions

Non-linear adaptive current-mode controllers for the regulation of the hybrid-type high-order boost converter in the presence of an unknown load were presented. The adaptive law, in which the derivative of the estimator is optimised, is used to estimate the conductance of the load. A comparative study involving the adaptive controllers using the input and output inductor currents

Fig. 7 Adaptive current-mode control

(a) Output response in the presence of load changes, (b) Output response in the presence of input voltage changes, (c) Output response in the presence of reference input changes

was carried out. The controller based on the output inductor current was found to be more suitable for regulating the hybrid-type high-order boost converter. Besides, the stability regions of the closed-loop converter system were investigated. Finally, some simulation and experimental results comparing the performance of the traditional and the proposed adaptive current-mode controllers were obtained and they show that the latter is more superior to the former.

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8 References

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