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A high efficiency input/output magnetically coupled interleaved buck–boost converter with low internal oscillation for fuel-cell applications: Small signal modeling and dynamic analysis

Vahid Samavatian *, Ahmad Radan

Electrical and Computer Engineering Faculty, K.N. Toosi University of Technology, Seyed Khandan Bridge, 1431714191 Tehran, Iran

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ABSTRACT

Dynamic behavior of DC–DC converters plays a crucial role in stability of renewable energy exploitation systems. This paper presents small signal modeling of an input/output magnetically coupled interleaved buck–boost converter for fuel-cell applications to help the designers with the better understanding of converter dynamics. Aiming to have a continuous converter transfer function for a smooth transition between the operation modes and an improved inner dynamics, a damping network and an input/output coupling have been added to the interleaved structure of well-known cascaded buck–boost converter. Having the same step-up/step-down voltage transfer ratio, smooth transition and improved inner dynamics make this converter quite suitable for renewable energy applications. The paper presents a small signal ac equivalent circuit model of the proposed converter based on state space averaging (SSA) method. Simulation results show remarkable improvements in converter dynamic behavior in both time and frequency domains. Prototype setup of 360 W and 36 V output voltage for a fuel cell with a brand of "FCgen 1020ACS" Ballard Power Systems, Inc. was implemented. Experimental results are presented to verify the theoretical model and its expected merits.

Introduction

Fuel cell (FC) energy sources demonstrate an unregulated output voltage with a wide range of variations under different load conditions [1–3]. In order to provide a constant load voltage an interface DC–DC converter is necessarily required [4–6].

Requirements such as low ripple input/output current, capability of high power density, high efficiency and fast dynamic response to load and input voltage changes have to be also met by choosing a suitable interface DC–DC converter [4–11].

Despite many single-active-switch buck–boost DC–DC converters such as Sepic, Cuk, conventional inverting buck–boost and flyback converters which can be drawn into consideration for this application, the non-inverting buck–boost converter, constructed by combining a boost and a buck circuit in cascade with two independently controllable switches, is a popular choice offering high efficiency and low component stresses [12–21].

A major problem of double-active-switch buck–boost converters is the presence of a dissociation in control-to-output transfer function called dead zone at the changing boundary of operation modes where input voltage is around output voltage [22–25]. This dissociation can make the controller confused and consequently leads to a disability in tracking the desired output voltage. A possible solution for improving the converter dynamic behavior in boundary mode is employing one of the dead zone avoidance and mitigation (DZAM) methods such as memory-less nonlinearity technique and/or tri-interval technique [26,27]. Dynamic and stability response improvement can be achieved using DZAM at expense of losing simplicity of control system.

A better possible solution reported by authors in the first part of this paper [32] is using a magnetic coupling between I/O inductors in double-active-switch buck–boost converter.

Comparing the features of both aforementioned methods, DZAM have two major problems including dead zone detection and dead zone mitigation implementing procedure. Accordingly, DZAM needs some algorithms to detect that whether operation point is in the dead zone or not. Performing such an algorithm can complicate the controlling system leading to cost increase. In addition, for implementing the conventional DZAM methods, there would be the requirement of extra switching leading to efficiency reduction [26]. On the other hand, not only does the magnetic coupling between I/O inductors remove above barriers, but also play a

* Corresponding author. Tel.: +98 21 88462408; fax: +98 21 88462066.
E-mail addresses: vahidsamavatian@ee.kntu.ac.ir, vahidsamavatian_90@yahoo.com (V. Samavatian), radan@eetd.kntu.ac.ir (A. Radan).

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Nomenclature

A	system state matrix
B	system input vector
C_1, C_2, C_d	interface, output and damping capacitor
D_{12}	steady state duty cycle of boost stage
D_{34}	steady state duty cycle of buck stage
$d_{12}(t)$	duty cycle of boost stage
$d_{34}(t)$	duty cycle of buck stage
$G_{vod12}(s)$	small signal control-to-output transfer functions with respect to $\hat{d}_{12}$
$G_{vod34}(s)$	small signal control-to-output transfer functions with respect to $\hat{d}_{34}$
i_{L1}, i_{L2}	inductor 1 and2 currents
i_{Lm1}, i_{Lm2}	magnetizing inductor 1 and 2 currents
$M(u)$	converter voltage transfer ratio
$M(D_{12}, D_{34})$	converter voltage transfer ratio
P_{Rd}	power dissipation in damping network resistor
$S_1 \sim S_4$	$Q_1 \sim Q_4$ signal activations
T_s	switching period
u	unified controlling variable
U	system input vector
v_{C1}, i_{C1}	capacitor 1 voltage and current
v_{C2}, i_{C2}	capacitor 2 voltage and current
v_{Cd}, i_{Cd}	damping network capacitor voltage and current
v_g, i_g	input voltage and current
v_{L1}, v_{L2}	inductor 1 and 2 voltages
v_{Lm1}, v_{Lm2}	magnetizing inductor 1 and 2 voltages
v_{out}, i_{out}	output voltage and current
$\hat{x}$	converter space vector

supplementary role for removing and damping the right-half-plane (RHP) zeros in the boost operation mode [32].

Other drawbacks of most DC–DC converters mentioned earlier are associated with poor efficiency and slow dynamic response to load and input voltage changes due to a RHP zero presence in continuous conduction mode (CCM). In many applications, especially in FCs, there are lots of output voltage changes; therefore fast dynamic response of interface converter is absolutely required [28–30].

The topology called KY buck–boost converter which improves dynamic response through RHP zero elimination in CCM has been proposed in [23] for solving these two major problems. However, this solution introduces employing of four power switches which results in increasing the converter costs. In order to mitigate this problem, a new topology with two active switches and aforementioned advantage is reported in [23] at the expense of higher component stresses and reduced efficiency. The two-switch tri-state buck–boost proposed in [31] also eliminates the RHP zero, but efficiency is still quite low. In addition, this method has never been applied to the cascading buck–boost topology.

Adding a damping network besides the magnetic coupling between input and output inductors proposed by authors of this paper will not only make the dynamic responses fast, but also result in a unique continuous control-to-output transfer function in the boundary of operating mode change leading to a smooth transition between operation modes.

In this part of paper fast dynamic response and continuous control-to-output transfer function in the boundary of operating mode change are mathematically studied and experimentally proved. Issues relating to steady state analysis, circuit design procedure, low-ripple I/O current and high efficiency discussion are studied in the first part of this paper [32].

After reviewing the principle of converter operation very briefly, state space equations of the converter are given in the following section. Dynamic analysis and small signal modeling of the proposed converter model are delivered in section 'Dynamic and small signal model of the converter'. Sections 'Modelling verification' and 'Impacts of damping network presence' deal with modeling verification and the impacts of presence of damping network in root locus map respectively. In Section 'Frequency domain analysis', the stability issues are discussed in the frequency domain. Section 'Experimental results' focuses on the experimental results and for demonstrating the impacts of presence of damping network, time domain responses are thoroughly studied for all operation modes. The last section presents a conclusion for this study.

Principle of operation

Fig. 1 demonstrates the proposed converter. Equivalent topology in both boost and buck operation modes is illustrated in Fig. 2(a) and (b). Turning on the switches 3 and 4 (Q_3 and Q_4) permanently and firing the switches 1 and 2 (Q_1 and Q_2) in PWM operating result in boost operation mode Similarly, in buck mode, while Q_1 and Q_2 are permanently kept turned off, Q_3 and Q_4 operate in PWM. The steady state operation waveforms of this converter for both boost and buck modes of operation are illustrated in [32]. Their duty cycles, $d_{12}(t)$ in boost mode and $d_{34}(t)$ in buck mode, are controlled in such a way that output voltage be regulated around a desired value (here is 36 V) in both operation modes. Please refer to the first part of this paper for details of steady state operation [32].

Including a capacitor and resistor connecting in series, damping network plays the role of decaying element for input voltage oscillations. The benefit of using this element and its impacts on the dynamic behavior of proposed converter would be explained completely in the following sections.

The two main methods used for obtaining the differential equations describing the dynamic behavior of interleaved converters are state space averaging (SSA) methods [16] and signal flow graph (SFG) [33].

The final differential equations modeling the converter behavior are given by (1) assuming the operation of continuous conduction mode (CCM) and a switching frequency much higher than the converter natural frequencies (for additional details and steady state analysis of the converter based on state space averaging method in details, readers are referred to the part I [32]).

$$\begin{cases} L_{m1}\left\langle \frac{di_{Lm1}(t)}{dt} \right\rangle_{T_s} = V_g(t) - V_{C1}(t)(1 - d_{12}(t)) \\ L_{m2}\left\langle \frac{di_{Lm2}(t)}{dt} \right\rangle_{T_s} = V_g(t) - V_{C1}(t)(1 - d_{12}(t)) \\ L_1\left\langle \frac{di_{L1}(t)}{dt} \right\rangle_{T_s} = V_g(t) - V_{out}(t) + V_{C1}(t)(d_{34}(t) + d_{12}(t) - 1) \\ L_2\left\langle \frac{di_{L2}(t)}{dt} \right\rangle_{T_s} = V_g(t) - V_{out}(t) + V_{C1}(t)(d_{34}(t) + d_{12}(t) - 1) \\ C_1\left\langle \frac{dV_{C1}(t)}{dt} \right\rangle_{T_s} = (i_{Lm1}(t) + i_{Lm2}(t))(1 - d_{12}(t)) \\ -\frac{V_{C1}(t) - V_{cd}(t)}{R_d} + (i_{L1}(t) + i_{L2}(t))(1 - d_{12}(t) - d_{34}(t)) \\ C_2\left\langle \frac{dV_{out}(t)}{dt} \right\rangle_{T_s} = (i_{L1}(t) + i_{L2}(t)) - \frac{V_{out}}{R} \\ C_d\left\langle \frac{dV_{Cd}(t)}{dt} \right\rangle_{T_s} = \frac{V_{C1}(t) - V_{cd}(t)}{R_d} \end{cases} \tag{1}$$