

# High-Efficiency Bidirectional Buck–Boost Converter for Photovoltaic and Energy Storage Systems in a Smart Grid

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**Abstract**—This paper proposes a new bidirectional buck–boost converter, which is a key component in a photovoltaic and energy storage system (ESS). Conventional bidirectional buck–boost converters for ESSs operate in discontinuous conduction mode (DCM) to achieve zero-voltage switching turn-ON for switches. However, operation in DCM causes high ripples in the output voltage and current, as well as low power-conversion efficiency. To improve the performance of the conventional converter, the proposed converter has a new combined structure of a cascaded buck–boost converter and an auxiliary capacitor. The combined structure of the proposed converter reduces the output current ripple by providing a current path and the efficiency is increased. A prototype was built and tested to verify the effectiveness of the converter. The proposed converter has a maximum efficiency of 98%, less than 5.14 V<sub>p,p</sub> of output voltage ripple, and less than 7.12 A<sub>p,p</sub> of output current ripple. These results were obtained at an input voltage of 160 V, switching frequency of 45 kHz, output voltage of 80–320 V, and output power of 16–160 W. The experimental results show that the proposed converter has improved performance compared to the conventional converter.

**Index Terms**—DC–DC power conversion, energy storage, pulsedwidth modulated power converters.

## I. INTRODUCTION

MART grids (see Fig. 1) are future electric energy systems that have been studied to reduce mismatch between sources of electricity (such as renewable energy and power plants) and electricity consumers (homes, vehicles, factories, etc.). However, the energy production of renewable energy depends on environmental conditions. Therefore, an energy storage system (ESS) is needed in a smart grid to provide stability and efficiently manage the renewable energy [1]–[3].

An ESS consists of a battery that stores electric energy and a bidirectional dc–dc converter that transfers energy from the

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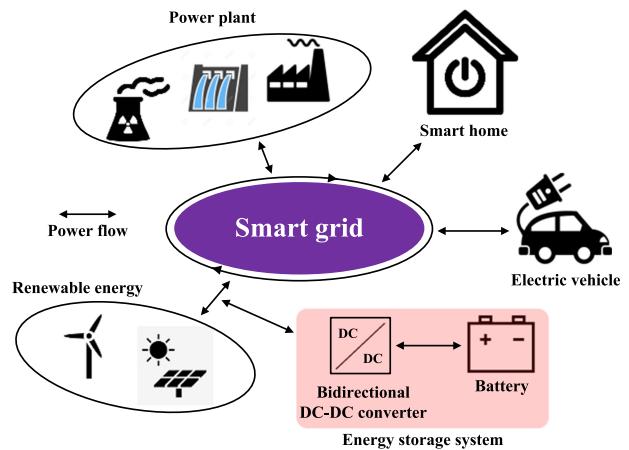


Fig. 1. Diagram of the smart grid.

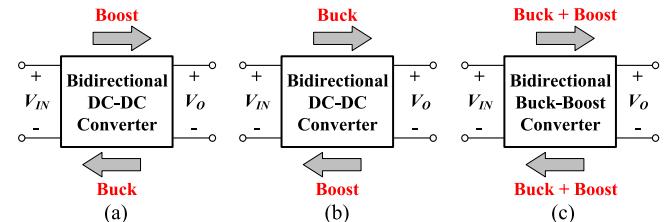


Fig. 2. Block diagrams in case of (a)  $V_{IN} < V_O$ , (b)  $V_{IN} > V_O$ , and (c)  $V_{IN} < V_O$  and  $V_{IN} > V_O$ .

battery and renewable energy source in both directions [4]–[9]. A conventional bidirectional dc–dc converter uses a half-bridge converter with two switches based on a buck or boost dc–dc converter. In the buck mode of the converter, electric energy is transferred from a high-voltage (HV) port to the low-voltage (LV) port. In boost mode, the electric energy is transferred from the LV port to the HV port. The conventional bidirectional converter has a limitation in that it can only be operated in buck mode in one direction and boost mode in the other direction [see Fig. 2(a) and (b)] [10]–[12]. Therefore, when the input is a photovoltaic (PV) module and the output is battery cells in a smart grid, a half-bridge converter based on a buck or boost converter cannot be used because of the following reasons.

- 1) The battery cells repeatedly perform charging and discharging operations, resulting in large voltage variation [13], [14].

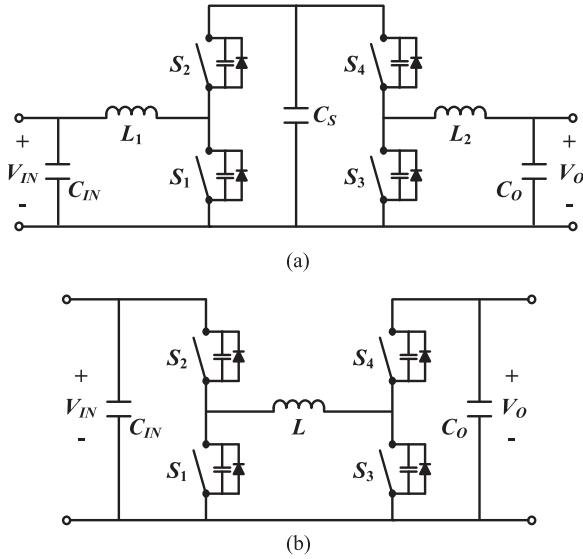


Fig. 3. Circuit structures of (a) CHB converter and (b) CBB converter.

- 2) The PV module has a large voltage variation that depends on the module temperature and the solar irradiance [15]–[17].

Thus, the ranges of the input voltage and output voltage can overlap [18], [19].

Bidirectional buck-boost converters [see Fig. 2(c)] were introduced for use in cases of overlapping input and output voltages [19]–[28]. They can operate in both buck and boost modes in both directions. A combined half-bridge (CHB) converter [see Fig. 3(a)] is the most basic bidirectional buck-boost converter and has a symmetric structure with respect to the storage capacitor \$C\_S\$ [20], [21]. There is one inductor at the input port and one at the output port, which results in LV ripples in the input and output. However, because the CHB converter uses two inductors of the same size, it is large and has a low power-conversion efficiency \$\eta\_e\$ due to the dc-offset current of each inductor.

Cascaded buck-boost (CBB) converter [see Fig. 3(b)], along with the CHB converter, has been commonly used in ESSs. Compared with the CHB converter, the CBB converter is smaller and has higher \$\eta\_e\$ because it uses only one inductor \$L\$ [19], [22]–[28]. Recently, research has been actively conducted on bidirectional buck-boost dc–dc converters in discontinuous conduction mode (DCM) because this mode can achieve zero-voltage switching (ZVS) turn-ON of the switches [19], [26]–[28]. However, operation in DCM increases the current ripple of \$L\$, which affects the output current ripple and increases the output voltage ripple.

In this paper, an enhanced CBB converter is proposed to improve the performance of the conventional CBB converter. The proposed converter is targeted to a PV-ESS system that uses a microinverter, which has been widely used in a smart grid [29], [30]. The converter has a new combined structure of a CBB converter and an auxiliary capacitor. This structure can reduce the output voltage ripple and increase \$\eta\_e\$ by effectively reducing the output current ripple. The circuit structure and operating principle of the proposed converter are described in

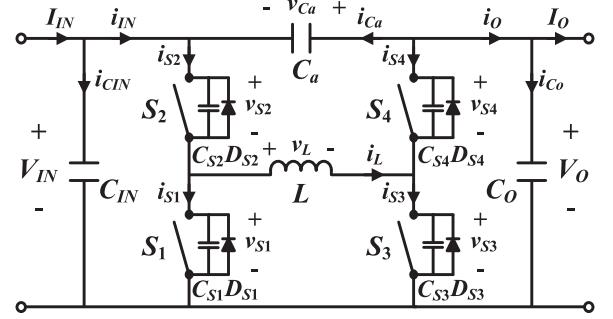


Fig. 4. Circuit structure of the proposed converter.

TABLE I  
STATES OF SWITCHES UNDER SIX OPERATING CONDITIONS

Six operating conditions		Switches			
Directions of energy transfer	Types of the operation	\$S_1\$	\$S_2\$	\$S_3\$	\$S_4\$
\$V_{IN} \rightarrow V_O\$	Buck	1 - \$D\$	\$D\$	0	1
	Buck-Boost	1 - \$D\$	\$D\$	\$D\$	1 - \$D\$
	Boost	0	1	\$D\$	1 - \$D\$
\$V_O \rightarrow V_{IN}\$	Buck	0	1	\$D\$	1 - \$D\$
	Buck-Boost	1 - \$D\$	\$D\$	\$D\$	1 - \$D\$
	Boost	1 - \$D\$	\$D\$	0	1

Section II, and design considerations are given in Section III. Experimental results are given in Section IV, and a conclusion is drawn in Section V.

## II. PROPOSED DC–DC CONVERTER

### A. Circuit Structure

The proposed converter (see Fig. 4) consists of a conventional CBB converter and an auxiliary capacitor (\$C\_a\$), and has a symmetric structure with respect to \$C\_a\$ and \$L\$. The CBB converter consists of two capacitors (\$C\_{IN}, C\_O\$), four switches (\$S\_1, S\_2, S\_3, S\_4\$), and an inductor (\$L\$). Four switches (\$S\_1, S\_2, S\_3, S\_4\$) and an inductor (\$L\$) control the direction of energy transfer and the ratio between the input voltage and output voltage. Four switches are turned ON under the ZVS condition by operating in DCM. Two capacitors (\$C\_{IN}, C\_O\$) reduce the output voltage ripple and noise, and an auxiliary capacitor (\$C\_a\$) reduces the output current ripple by providing a current path.

### B. Principle of Operation

The proposed converter operates with a fixed switching period \$T\_S\$ and controls the voltage gain by changing the duty ratio \$D\$ of the switches (\$S\_1, S\_2, S\_3, S\_4\$) from 0 to 1. Each switch has four states under six operating conditions created by the energy transfer directions between \$V\_{IN}\$ and \$V\_O\$ and the types of operation (buck, boost, and buck-boost), as shown in Table I. Due to the symmetric structure with respect to \$C\_a\$ and \$L\$, the operations are separated by only the types of operation in one direction of energy transfer (\$V\_{IN} \rightarrow V\_O\$).

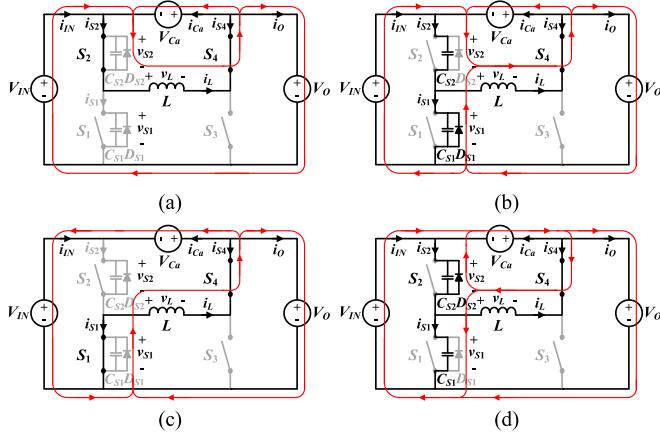


Fig. 5. Circuit diagrams for the operation of buck. (a) Mode 1. (b) Mode 2. (c) Mode 3. (d) Mode 4.

To simplify the analysis of the operation, the following assumptions are made: 1) the inductor and all capacitors are lossless; 2) the voltage ripples of  $C_{IN}$ ,  $C_a$ , and  $C_O$  are small enough to assume that  $V_{IN}$ ,  $V_{Ca}$ , and  $V_O$  are constant voltage sources; and 3) the converter operates in steady state.

*1) Buck Mode:* When the proposed converter operates in buck mode, it has four distinct operating modes (Modes 1–4). The equivalent circuits and operating waveforms are shown in Figs. 5 and 6.

*Mode 1 [see Fig. 5(a),  $t_0 \leq t \leq t_1$ ]:* starts when  $S_2$  is turned ON. At  $t = t_0$ ,  $S_2$  achieves ZVS turn-ON because the body diode  $D_{S2}$  of  $S_2$  is turned ON before  $t = t_0$ . Then, the voltage  $v_L$  of  $L$  becomes  $V_{IN} - V_O$ , and the current  $i_L$  of  $L$  is expressed as

$$i_L(t) = i_L(t_0) + \frac{V_{IN} - V_O}{L}(t - t_0). \quad (1)$$

The current  $i_{S2}$  of  $S_2$  is equal to  $i_L$ , so  $i_{S2}$  is expressed as

$$i_{S2}(t) = i_L(t_0) + \frac{V_{IN} - V_O}{L}(t - t_0).$$

In this mode, the current  $i_{Ca}$  of  $C_a$ , the current  $i_{Co}$  of  $C_O$ , the output current  $i_O$ , and the load current  $I_O$  have the following relations:  $i_O = i_{Co} + I_O$ ,  $i_{Co} = i_{Ca} \cdot C_O/C_a$ , and  $i_O = i_L - i_{Ca}$ . Therefore,  $i_{Ca}$  and  $i_O$  can be derived as

$$\begin{aligned} i_{Ca}(t) &= \frac{C_a}{C_a + C_O}[i_L(t) - I_O] \\ i_O(t) &= \frac{C_O}{C_a + C_O}i_L(t) + \frac{C_a}{C_a + C_O}I_O. \end{aligned} \quad (2)$$

The voltage  $v_{Ca}$  across the auxiliary capacitor  $C_a$  is expressed as

$$v_{Ca}(t) = V_{Ca} + \Delta v_{Ca,AC}(t)$$

where  $V_{Ca}$  and  $\Delta v_{Ca,AC}$  represent the dc voltage and the ac ripple voltage across the  $C_a$ , respectively. Because  $V_{Ca} \gg \Delta v_{Ca,AC}$ ,  $v_{Ca}$  can be approximated as

$$v_{Ca}(t) \approx V_{Ca} = V_O - V_{IN}.$$

*Mode 2 [see Fig. 5(b),  $t_1 \leq t \leq t_2$ ]:* starts when  $S_2$  is turned OFF. At this time,  $S_1$  remains in the OFF state to prevent a

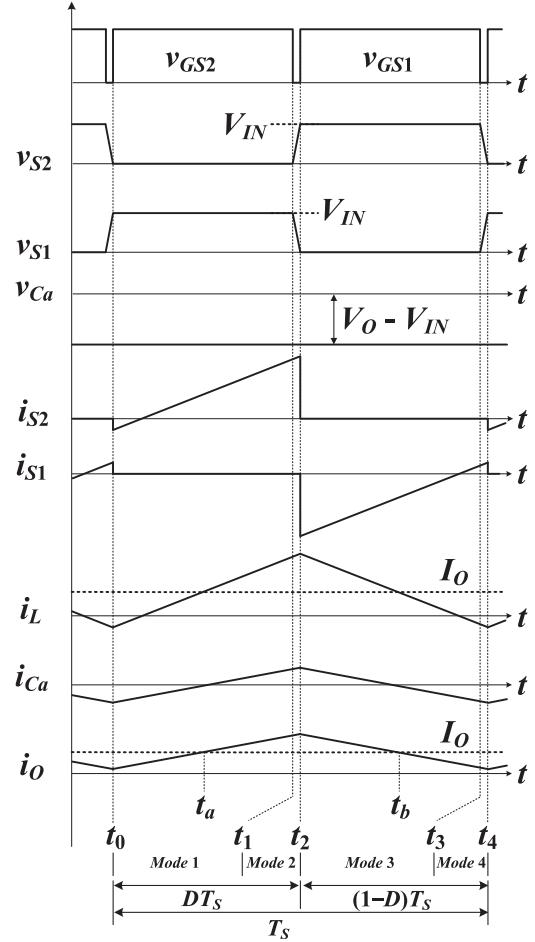


Fig. 6. Operational waveforms in the operation of buck.

shoot-through problem with  $S_1$  and  $S_2$ . In this mode, the output capacitor  $C_{S1}$  of  $S_1$  discharges from  $V_{IN}$  to 0, and the output capacitor  $C_{S2}$  of  $S_2$  charges from 0 to  $V_{IN}$ . Shortly after the discharging of  $C_{S1}$  and charging of  $C_{S2}$  are finished, the body diode  $D_{S1}$  of  $S_1$  is turned ON.

*Mode 3 [see Fig. 5(c),  $t_2 \leq t \leq t_3$ ]:* starts with the ZVS turn-on of  $S_1$  because  $D_{S1}$  is turned ON before  $t = t_2$ . Then,  $v_L$  becomes  $-V_O$ , and thereby  $i_L$  is expressed as

$$i_L(t) = i_L(t_2) - \frac{V_O}{L}(t - t_2). \quad (3)$$

The current  $i_{S1}$  of  $S_1$  is equal to  $-i_L$ , so  $i_{S1}$  is obtained as

$$i_{S1}(t) = -i_L(t_2) + \frac{V_O}{L}(t - t_2).$$

Because  $i_O = i_L - i_{Ca}$ ,  $i_O = i_{Co} + I_O$ , and  $i_{Co} = i_{Ca} \cdot C_O/C_a$ ,  $i_{Ca}$  and  $i_O$  are expressed as

$$i_{Ca}(t) = \frac{C_a}{C_a + C_O}[i_L(t) - I_O] \quad (4)$$

$$i_O(t) = \frac{C_O}{C_a + C_O}i_L(t) + \frac{C_a}{C_a + C_O}I_O. \quad (5)$$

*Mode 4 [see Fig. 5(d),  $t_3 \leq t \leq t_4$ ]:* starts when  $S_1$  is turned OFF and  $S_2$  remains in the OFF state. In this mode,  $C_{S1}$  charges from 0 to  $V_{IN}$ , and  $C_{S2}$  discharges from  $V_{IN}$  to 0. Shortly

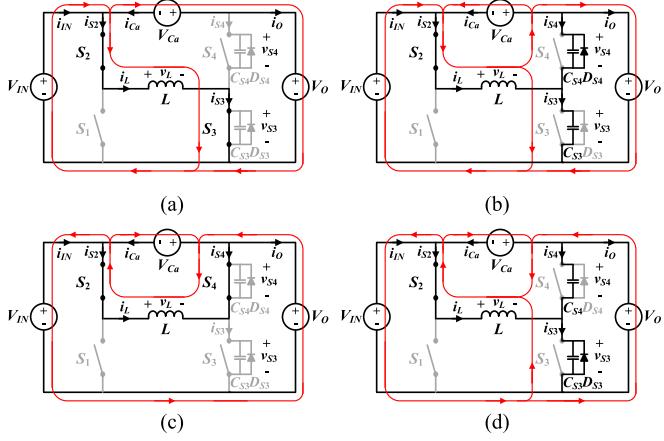


Fig. 7. Circuit diagrams for the operation of boost. (a) Mode 1. (b) Mode 2. (c) Mode 3. (d) Mode 4.

after the charging of  $C_{S1}$  and discharging of  $C_{S2}$  are finished,  $D_{S2}$  is turned ON.

At  $t = t_0$ ,  $i_L$  has an initial value of  $i_L(t_0)$ , and  $i_L(t_0)$  is obtained as follows: by inserting  $t = t_2$  into (1), the current ripple  $\Delta i_L$  of  $i_L$  is obtained as

$$\Delta i_L = i_L(t_2) - i_L(t_0) = \frac{V_{IN} - V_O}{L} DT_S \quad (6)$$

where  $DT_S = t_2 - t_0$ . The average current of  $L$  for one  $T_S$  is obtained as  $\langle i_L \rangle = I_O$  by applying the ampere-second balance law for capacitors to  $i_L(t) = i_{Ca}(t) + i_{Co}(t) + I_O$ . Then,  $i_L(t_0) = \langle i_L \rangle - \Delta i_L/2$  is represented as

$$i_L(t_0) = I_O - \frac{V_{IN} - V_O}{2L} DT_S \quad (7)$$

and  $i_L(t_2) = \langle i_L \rangle + \Delta i_L/2$  is expressed as

$$i_L(t_2) = I_O + \frac{V_{IN} - V_O}{2L} DT_S. \quad (8)$$

2) *Boost Mode*: The boost operation also has four distinct operating modes (Modes 1–4), and the equivalent circuits and operating waveforms are shown in Figs. 7 and 8, respectively.

*Mode 1* [see Fig. 7(a),  $t_0 \leq t \leq t_1$ ]: starts when  $S_3$  is turned ON. At  $t = t_0$ ,  $S_3$  achieves ZVS turn-ON because the body diode  $D_{S3}$  of  $S_3$  is turned ON before  $t = t_0$ . Then,  $v_L$  becomes  $V_{IN}$ , and  $i_L$  is expressed as

$$i_L(t) = i_L(t_0) + \frac{V_{IN}}{L} (t - t_0). \quad (9)$$

The current  $i_{S3}$  of  $S_3$  is the same as  $i_L$ , so  $i_{S3}$  is expressed as

$$i_{S3}(t) = i_L(t_0) + \frac{V_{IN}}{L} (t - t_0).$$

Because  $i_O = -i_{Ca}$ ,  $i_O = i_{Co} + I_O$ , and  $i_{Co} = i_{Ca} \cdot C_O/C_a$ ,  $i_{Ca}$  and  $i_O$  are expressed as

$$\begin{aligned} i_{Ca}(t) &= -\frac{C_a}{C_a + C_O} I_O \\ i_O(t) &= \frac{C_a}{C_a + C_O} I_O. \end{aligned} \quad (10)$$

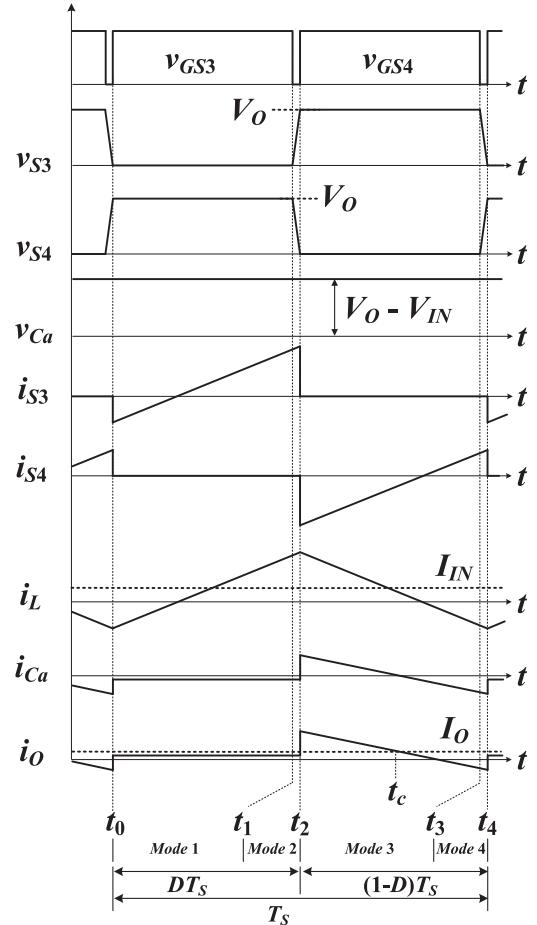


Fig. 8. Operational waveforms in the operation of boost.

The voltage  $v_{Ca}$  across the auxiliary capacitor  $C_a$  is expressed as

$$v_{Ca}(t) = V_{Ca} + \Delta v_{Ca,AC}(t).$$

Because  $V_{Ca} \gg \Delta v_{Ca,AC}$ , this expression can be approximated as

$$v_{Ca}(t) \approx V_{Ca} = V_O - V_{IN}.$$

*Mode 2* [see Fig. 7(b),  $t_1 \leq t \leq t_2$ ]: starts when  $S_3$  is turned OFF and  $S_4$  remains in the OFF state. In this mode, the output capacitor  $C_{S3}$  of  $S_3$  charges from 0 to  $V_O$ , and the output capacitor  $C_{S4}$  of  $S_4$  discharges from  $V_O$  to 0. Shortly after the charging of  $C_{S3}$  and discharging of  $C_{S4}$  are finished, the body diode  $D_{S4}$  of  $S_4$  is turned ON.

*Mode 3* [see Fig. 7(c),  $t_2 \leq t \leq t_3$ ]: starts with the ZVS turn-ON of  $S_4$  because  $D_{S4}$  is turned ON before  $t = t_2$ . Then,  $v_L$  becomes  $V_{IN} - V_O$ , and  $i_L$  is expressed as

$$i_L(t) = i_L(t_2) + \frac{V_{IN} - V_O}{L} (t - t_2). \quad (11)$$

The current  $i_{S4}$  of  $S_4$  is equal to  $-i_L$ , so  $i_{S4}$  is obtained as

$$i_{S4}(t) = -i_L(t_2) - \frac{V_{IN} - V_O}{L} (t - t_2).$$

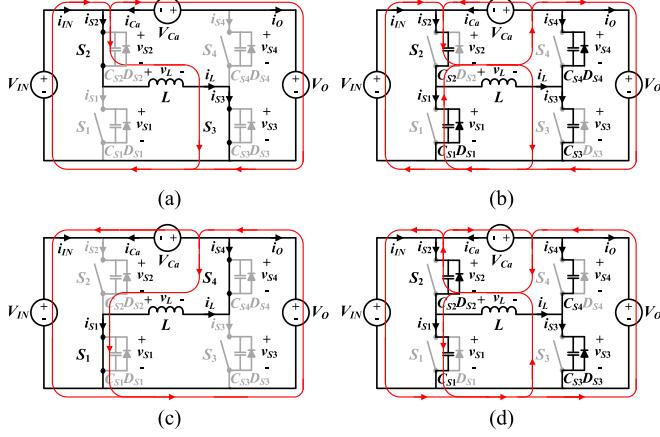


Fig. 9. Circuit diagrams for the operation of buck-boost. (a) Mode 1. (b) Mode 2. (c) Mode 3. (d) Mode 4.

Because  $i_O = i_L - i_{Ca}$ ,  $i_O = i_{Co} + I_O$ , and  $i_{Co} = i_{Ca} \cdot C_O / C_a$ ,  $i_{Ca}$  and  $i_O$  are expressed as

$$i_{Ca}(t) = \frac{C_a}{C_a + C_O} [i_L(t) - I_O] \quad (12)$$

$$i_O(t) = \frac{C_O}{C_a + C_O} i_L(t) + \frac{C_a}{C_a + C_O} I_O. \quad (13)$$

*Mode 4* [see Fig. 7(d),  $t_3 \leq t \leq t_4$ ]: starts when  $S_4$  is turned OFF and  $S_3$  remains in OFF state. In this mode,  $C_{S3}$  discharges from  $V_O$  to 0 and  $C_{S4}$  charges from 0 to  $V_O$ . Shortly after the discharging of  $C_{S3}$  and charging of  $C_{S4}$  are finished,  $D_{S3}$  is turned ON.

By inserting  $t = t_2$  into (9),  $\Delta i_L$  for boost operation is obtained as

$$\Delta i_L = i_L(t_2) - i_L(t_0) = \frac{V_{IN}}{L} DT_S \quad (14)$$

where  $DT_S = t_2 - t_0$ . Because  $< i_L > = I_{IN}$  and  $i_L(t_0) = < i_L > - \Delta i_L / 2$ ,  $i_L(t_0)$  is expressed as

$$i_L(t_0) = I_{IN} - \frac{V_{IN}}{2L} DT_S. \quad (15)$$

$i_L(t_2) = < i_L > + \Delta i_L / 2$  is expressed as

$$i_L(t_2) = I_{IN} + \frac{V_{IN}}{2L} DT_S. \quad (16)$$

*3) Buck-Boost Mode:* The buck-boost operation has four distinct operating modes (Modes 1–4), and the equivalent circuits and operating waveforms are shown in Figs. 9 and 10, respectively.

*Mode 1* [see Fig. 9(a),  $t_0 \leq t \leq t_1$ ]: starts when  $S_2$  and  $S_3$  are turned ON. At  $t = t_0$ ,  $S_2$  and  $S_3$  achieve ZVS turn-ON because  $D_{S2}$  and  $D_{S3}$  are turned ON before  $t = t_0$ . Then,  $v_L$  becomes  $V_{IN}$ , and  $i_L$  is expressed as

$$i_L(t) = i_L(t_0) + \frac{V_{IN}}{L} (t - t_0). \quad (17)$$

Both  $i_{S2}$  and  $i_{S3}$  are the same as  $i_L$ , so they are expressed as

$$i_{S2}(t) = i_{S3}(t) = i_L(t_0) + \frac{V_{IN}}{L} (t - t_0).$$

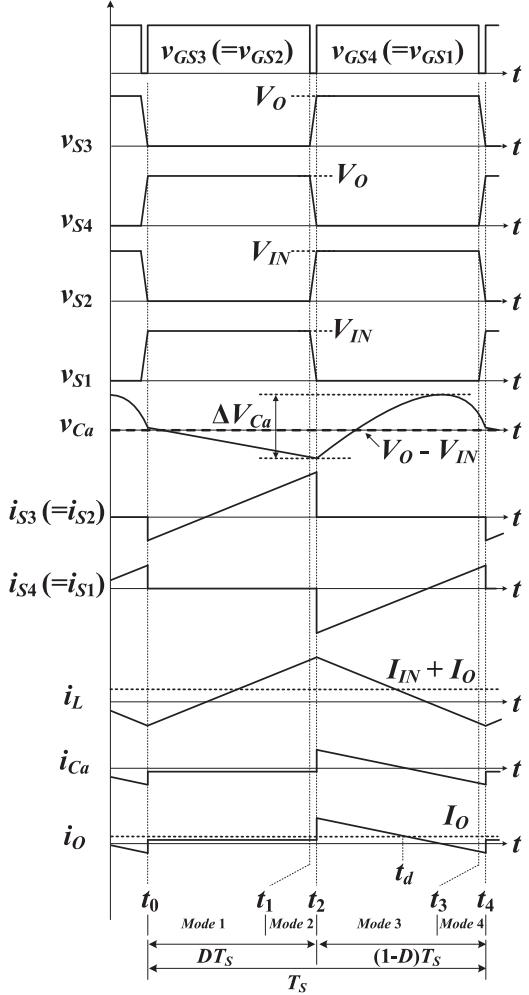


Fig. 10. Operational waveforms in the operation of buck-boost.

Because  $i_O = -i_{Ca}$ ,  $i_O = i_{Co} + I_O$ , and  $i_{Co} = i_{Ca} \cdot C_O / C_a$ ,  $i_{Ca}$  and  $i_O$  are expressed as

$$i_{Ca}(t) = -\frac{C_a}{C_a + C_O} I_O \quad (18)$$

$$i_O(t) = \frac{C_a}{C_a + C_O} I_O. \quad (19)$$

$v_{Ca}(t)$  is obtained using (18) as

$$\begin{aligned} v_{Ca}(t) &= v_{Ca}(t_0) + \frac{1}{C_a} \int_{t_0}^t i_{Ca}(t) dt \\ &\approx v_{Ca} + \frac{1}{C_a} \left[ \frac{-C_a I_O}{C_a + C_O} (t - t_0) \right] \\ &= V_O - \frac{I_O}{C_a + C_O} (t - t_0). \end{aligned}$$

*Mode 2* [see Fig. 9(b),  $t_1 \leq t \leq t_2$ ]: starts when  $S_2$  and  $S_3$  are turned OFF and  $S_1$  and  $S_4$  remain in the OFF state. In this mode,  $C_{S2}$  and  $C_{S3}$  charge from 0 to  $V_{IN}$  and from 0 to  $V_O$ , respectively.  $C_{S1}$  and  $C_{S4}$  discharge from  $V_{IN}$  to 0 and from  $V_O$  to 0, respectively. Shortly after the charging and discharging processes are finished,  $D_{S1}$  and  $D_{S4}$  are turned ON.

*Mode 3* [see Fig. 9(c),  $t_2 \leq t \leq t_3$ ]: starts with the ZVS turn-on of  $S_1$  and  $S_4$  because  $D_{S1}$  and  $D_{S4}$  are turned ON before  $t = t_2$ . Then,  $v_L$  becomes  $-V_O$ , and  $i_L$  is expressed as

$$i_L(t) = i_L(t_2) - \frac{V_O}{L}(t - t_2). \quad (20)$$

$i_{S1}$  and  $i_{S4}$  are equal to  $-i_L$ , so they are obtained as

$$i_{S1}(t) = i_{S4}(t) = -i_L(t_2) + \frac{V_O}{L}(t - t_2).$$

Because  $i_O = i_L - i_{Ca}$ ,  $i_O = i_{Co} + I_O$ , and  $i_{Co} = i_{Ca}$ .  $C_O/C_a$ ,  $i_{Ca}$  and  $i_O$  are expressed as

$$i_{Ca}(t) = \frac{C_a}{C_a + C_O}[i_L(t) - I_O] \quad (21)$$

$$i_O(t) = \frac{C_O}{C_a + C_O}i_L(t) + \frac{C_a}{C_a + C_O}I_O. \quad (22)$$

$v_{Ca}$  is obtained using (20) and (21) as

$$\begin{aligned} v_{Ca}(t) &= \frac{1}{C_a} \int_{t_2}^t i_{Ca}(t) dt + v_{Ca}(t_2) \\ &= \frac{1}{C_a + C_O} \left[ (i_L(t_2) - I_O)(t - t_2) - \frac{V_O}{2L}(t - t_2)^2 \right] \\ &\quad + v_{Ca}(t_2). \end{aligned}$$

*Mode 4* [see Fig. 9(d),  $t_3 \leq t \leq t_4$ ]: starts when  $S_1$  and  $S_4$  are turned OFF and  $S_2$  and  $S_3$  remain in the OFF state. In this mode,  $C_{S2}$  and  $C_{S3}$  discharge from  $V_{IN}$  to 0 and from  $V_O$  to 0, respectively.  $C_{S1}$  and  $C_{S4}$  charge from 0 to  $V_{IN}$  and from 0 to  $V_O$ , respectively. Shortly after the discharging and charging processes are finished,  $D_{S2}$  and  $D_{S3}$  are turned ON.

By inserting  $t = t_2$  into (17),  $\Delta i_L$  is obtained for buck-boost operation as

$$\Delta i_L = i_L(t_2) - i_L(t_0) = \frac{V_{IN}}{L}DT_S \quad (23)$$

where  $DT_S = t_2 - t_0$ . Because  $< i_L > = I_{IN} + I_O$  and  $i_L(t_0) = < i_L > - \Delta i_L/2$ ,  $i_L(t_0)$  is expressed as

$$i_L(t_0) = I_{IN} + I_O - \frac{V_{IN}}{2L}DT_S \quad (24)$$

and  $i_L(t_2) = < i_L > + \Delta i_L/2$  is expressed as

$$i_L(t_2) = I_{IN} + I_O + \frac{V_{IN}}{2L}DT_S. \quad (25)$$

### C. Voltage Gain

When the proposed converter operates in buck mode,  $v_L$  becomes  $V_{IN} - V_O$  for  $t_0 \leq t \leq t_2$  and  $-V_O$  for  $t_2 \leq t \leq t_4$ . Therefore, the volt-second balance law for  $L$  results in

$$\frac{V_O}{V_{IN}} = D \quad (26)$$

where  $DT_S = t_2 - t_0$  and  $(1 - D)T_S = t_4 - t_2$ .

In boost mode,  $v_L$  becomes  $V_{IN}$  for  $t_0 \leq t \leq t_2$  and  $V_{IN} - V_O$  for  $t_2 \leq t \leq t_4$ . Therefore, the volt-second balance law for  $L$  results in

$$\frac{V_O}{V_{IN}} = \frac{1}{1 - D} \quad (27)$$

where  $DT_S = t_2 - t_0$  and  $(1 - D)T_S = t_4 - t_2$ .

In buck-boost mode,  $v_L$  becomes  $V_{IN}$  for  $t_0 \leq t \leq t_2$  and  $-V_O$  for  $t_2 \leq t \leq t_4$ . Therefore, the volt-second balance law for  $L$  results in

$$\frac{V_O}{V_{IN}} = \frac{D}{1 - D} \quad (28)$$

where  $DT_S = t_2 - t_0$  and  $(1 - D)T_S = t_4 - t_2$ .

### D. Output Voltage Ripple

When the proposed converter operates in buck mode, the output voltage ripple  $\Delta v_o$  is expressed as

$$\Delta v_o = \frac{1}{C_O} \int_{t_a}^{t_b} i_{Co}(t) dt. \quad (29)$$

For  $t_0 \leq t \leq t_2$ ,  $i_{Co}$  is obtained using (1), (2),  $i_O = i_{Co} + I_O$ , and (7) as

$$i_{Co}(t) = \frac{C_O(V_{IN} - V_O)}{L(C_a + C_O)} \left[ (t - t_0) - \frac{DT_S}{2} \right]. \quad (30)$$

For  $t_2 \leq t < t_4$ , (3), (5),  $i_O = i_{Co} + I_O$ , and (8) result in

$$i_{Co}(t) = \frac{C_O}{C_a + C_O} \left[ \frac{-V_O}{L}(t - t_2) + \frac{V_{IN} - V_O}{2L}DT_S \right]. \quad (31)$$

As shown in Fig. 6,  $i_{Co}(t) = 0$  (or  $i_O(t) = I_O$ ) occurs at  $t = t_a$  and  $t = t_b$ . By inserting  $t = t_a$  into (30),  $t_a$  is obtained as

$$t_a = t_0 + \frac{DT_S}{2} \quad (32)$$

and  $t_b$  is obtained by inserting  $t = t_b$  into (31) as follows

$$t_b = t_2 + \frac{V_{IN} - V_O}{2V_O}DT_S.$$

Using (26), this equation can be represented as

$$t_b = t_2 + \frac{(1 - D)T_S}{2}. \quad (33)$$

Then,  $\Delta v_o$  is obtained using (26), (29), (31)–(33) as

$$\Delta v_o = \frac{V_O(1 - D)T_S^2}{8L(C_a + C_O)}. \quad (34)$$

When the proposed converter operates in boost mode,  $\Delta v_o$  is expressed as

$$\Delta v_o = \frac{1}{C_O} \int_{t_2}^{t_c} i_{Co}(t) dt. \quad (35)$$

For  $t_2 \leq t < t_4$ , (11), (13),  $i_O = i_{Co} + I_O$ , and (16) result in

$$\begin{aligned} i_{Co}(t) &= \frac{C_O}{C_a + C_O} \left[ I_{IN} + \frac{V_{IN} - V_O}{L}(t - t_2) \right. \\ &\quad \left. + \frac{V_{IN}}{2L}DT_S - I_O \right]. \end{aligned} \quad (36)$$

As shown in Fig. 8,  $i_{Co}(t)$  becomes 0 at  $t = t_2$  and  $t = t_c$ . By inserting  $t = t_c$  into (36),  $t_c$  is obtained as

$$t_c = t_2 + \frac{(I_{IN} - I_O)L + V_{IN}DT_S/2}{V_O - V_{IN}}. \quad (37)$$

Then,  $\Delta v_o$  is obtained using (35)–(37) as

$$\Delta v_o = \frac{[(I_{IN} - I_O)L + V_{IN}DT_S/2]^2}{2L(C_a + C_O)(V_O - V_{IN})}. \quad (38)$$

In buck-boost operation,  $\Delta v_o$  is expressed as

$$\Delta v_o = \frac{1}{C_O} \int_{t_2}^{t_d} i_{CO}(t) dt. \quad (39)$$

For  $t_2 \leq t < t_4$ , (20), (22),  $i_O = i_{CO} + I_O$ , and (25) result in

$$i_{CO}(t) = \frac{C_O}{C_a + C_O} \left[ I_{IN} - \frac{V_O}{L}(t - t_2) + \frac{V_{IN}}{2L}DT_S \right]. \quad (40)$$

As shown in Fig. 10,  $i_{CO}(t)$  becomes 0 at  $t = t_2$  and  $t = t_d$ . By inserting  $t = t_d$  into (40),  $t_d$  is obtained as

$$t_d = t_2 + \frac{LI_{IN} + V_{IN}DT_S/2}{V_O}. \quad (41)$$

Then,  $\Delta v_o$  is obtained using (39)–(41) as

$$\Delta v_o = \frac{(LI_{IN} + V_{IN}DT_S/2)^2}{2LV_O(C_a + C_O)}. \quad (42)$$

### E. Loss Analysis

The five main causes of power dissipation in the proposed converter include the switching and conduction losses of switches, the winding and core losses of inductor  $L$ , and the equivalent series resistance (ESR) losses of the capacitors [31]–[33].

The switching loss of a switch ( $P_{SW,swit}$ ) usually includes turn-ON loss, turn-OFF loss, reverse recovery loss of the body diode, and output capacitance loss. However, the proposed converter has only the turn-OFF and output capacitance losses because it achieves ZVS turn-ON for the switches. The turn-OFF loss of switch is expressed as

$$P_{SW,turn-off} = \frac{1}{2}V_{SW,turn-off}I_{SW,turn-off}(T_1 + T_2)f_S \quad (43)$$

where  $T_1 = (V_{SW,turn-off}C_{gd}R_g)/(V_t + I_{SW,turn-off}/g)$ ,  $T_2 = R_gC_g \ln(I_{SW,turn-off}/(gV_t) + 1)$ ,  $V_{SW,turn-off}$  and  $I_{SW,turn-off}$  are the switch voltage and current at the instance of turn-OFF transition,  $C_{gd}$  is the gate–drain capacitance,  $R_g$  is the gate resistance,  $C_g$  is the gate capacitance,  $g$  is the transconductance, and  $V_t$  is the threshold voltage of a switch. The output capacitance loss of a switch is expressed as

$$P_{Cds} = \frac{1}{2}C_{ds}V_{SW,turn-off}^2f_S \quad (44)$$

where  $C_{ds}$  is the output capacitance of the switch. The switching loss of the switch is expressed as

$$P_{SW,swit} = P_{SW,turn-off} + P_{Cds}. \quad (45)$$

The conduction loss of a switch ( $P_{SW,cond}$ ) is expressed as

$$P_{SW,cond} = I_{SW,rms}^2R_{on} \quad (46)$$

where  $I_{SW,rms}$  is the RMS value of the switch current, and  $R_{on}$  is the ON-resistance of the switch.

The winding loss of inductor ( $P_{L,wind}$ ) is expressed as

$$P_{L,wind} = I_{L,rms}^2R_L \quad (47)$$

TABLE II  
VOLTAGE AND CURRENT OF THE KEY POWER COMPONENTS FOR CALCULATING FIVE MAIN CAUSES OF POWER DISSIPATION

Five main causes of power dissipation	Operating modes		
	Buck mode	Boost mode	Buck-Boost mode
Switching loss of switch	$V_{S1,turn-off}$ $= V_{S2,turn-off} = V_{IN}$ $I_{S1,turn-off}$ $= -I_O + \frac{V_{IN} - V_O}{2L}DT_S$ $I_{S2,turn-off}$ $= I_O + \frac{V_{IN} - V_O}{2L}DT_S$	$V_{S3,turn-off}$ $= V_{S4,turn-off} = V_O$ $I_{S3,turn-off}$ $= I_{IN} + \frac{V_{IN}}{2L}DT_S$ $I_{S4,turn-off}$ $= -I_{IN} + \frac{V_{IN}}{2L}DT_S$	$V_{S1,turn-off}$ $= V_{S2,turn-off} = V_{IN}$ $V_{S3,turn-off}$ $= V_{S4,turn-off} = V_O$ $I_{S2,turn-off}$ $= I_{S3,turn-off}$ $= I_{IN} + I_O + \frac{V_{IN}}{2L}DT_S$ $I_{S1,turn-off} = I_{S4,turn-off}$ $= -I_{IN} - I_O + \frac{V_{IN}}{2L}DT_S$
	$I_{S1,rms} = \sqrt{1-D}I_{L,rms}$ $I_{S2,rms} = \sqrt{D}I_{L,rms}$ $I_{S4,rms} = I_{L,rms}$	$I_{S3,rms} = \sqrt{D}I_{L,rms}$ $I_{S4,rms} = \sqrt{1-D}I_{L,rms}$ $I_{S1,rms} = I_{L,rms}$	$I_{S1,rms} = \sqrt{1-D}I_{L,rms}$ $I_{S2,rms} = \sqrt{D}I_{L,rms}$ $I_{S3,rms} = \sqrt{D}I_{L,rms}$ $I_{S4,rms} = \sqrt{1-D}I_{L,rms}$
	$I_{L,rms}$ $= \sqrt{I_O^2 + \frac{V_{IN} - V_O}{12L}DT_S}$	$I_{L,rms}$ $= \sqrt{I_{IN}^2 + \frac{V_{IN}}{12L}DT_S}$	$I_{L,rms}$ $= \sqrt{(I_{IN} + I_O)^2 + \frac{V_{IN}}{12L}DT_S}$
	$\Delta i_L = \frac{V_{IN} - V_O}{L}DT_S$	$\Delta i_L = \frac{V_{IN}}{L}DT_S$	$\Delta i_L = \frac{V_{IN}}{L}DT_S$
Conduction loss of inductor	$I_{Ca,rms} = I_{Co,rms}$ $= \frac{V_{IN} - V_O}{4\sqrt{3}L}DT_S$	$I_{Ca,rms} = I_{Co,rms}$	$I_{Ca,rms} = I_{Co,rms}$ $\approx \frac{1}{2}\sqrt{1-D}I_{L,rms}$
Core loss of inductor			
ESR loss of capacitor			

where  $R_L$  is the winding resistance of the inductor.

The core loss of inductor ( $P_{L,core}$ ) is expressed as

$$P_{L,core} = k f_S^x B_{max}^y (l_e A_e) \quad (48)$$

where  $k$ ,  $x$ , and  $y$  are the coefficient of core loss, maximum flux density  $B_{max}$  is  $B_{max} = L\Delta i_L/(NA_e)$ ,  $N$  is the number of turns,  $l_e$  is the effective magnetic path length of the core, and  $A_e$  is the effective cross-sectional area of the core.

The ESR loss of capacitor ( $P_{C,ESR}$ ) is expressed as

$$P_{C,ESR} = I_{C,rms}^2 R_C \quad (49)$$

where  $I_{C,rms}$  is the RMS value of the capacitor current and  $R_C$  is the ESR of the capacitor.

The voltages and currents of the key power components for calculating these power losses have different values for the different operating modes (see Table II). The theoretical efficiency of the proposed converter can be calculated by inserting voltage and current values into (43)–(49) related to the power loss calculations.

### F. Operation of the Controller

The proposed converter is controlled by pulsedwidth modulation (PWM) signals ( $v_{g1} - v_{g4}$ ), which are generated by the voltage-mode control [see Fig. 11(a)]. Two voltages ( $V_{IN}$  and  $V_O$ ) are sensed to implement the voltage-mode control and protect the overvoltage.  $V_O$  is used as an output voltage for generating the PWM control signal of the main switch in the energy transfer direction from  $V_{IN}$  to  $V_O$ , and  $V_{IN}$  is used as an output voltage in the opposite direction. The current of inductor is sensed to protect against over current.

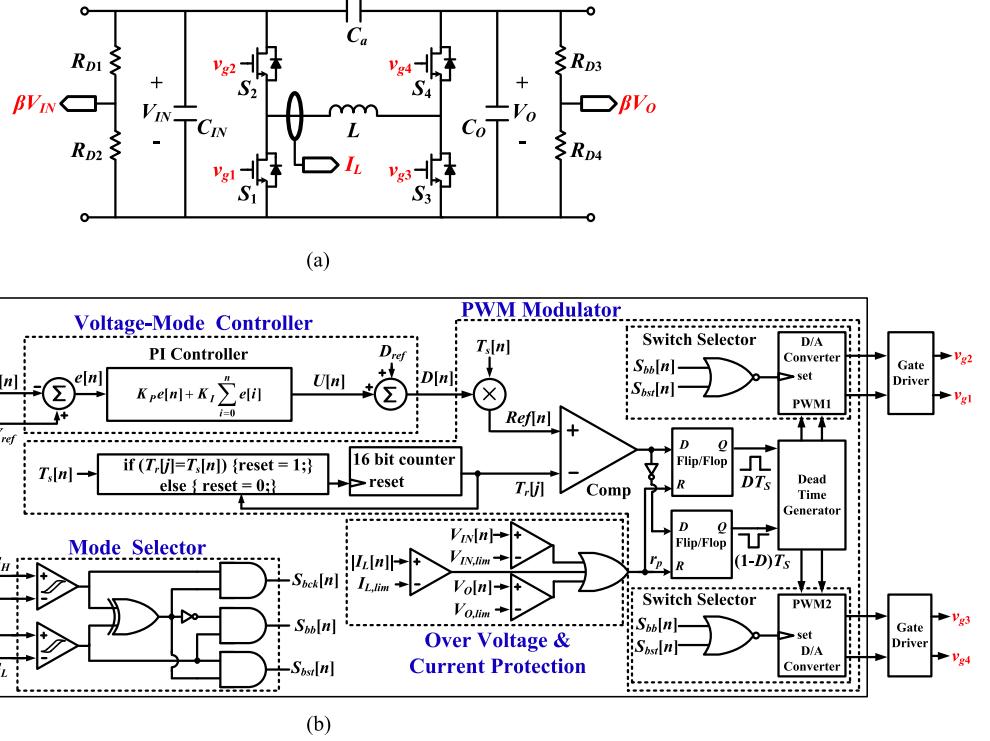


Fig. 11. (a) Circuit structure of the proposed converter with the voltage and current sensing ( $\beta V_{IN}$ ,  $\beta V_O$ , and  $I_L$ ). (b) Block diagram of the digital controller for the proposed converter.

Fig. 11(b) represents a block diagram of the digital controller for the proposed converter. When the energy transfer direction is expressed by the demand of the system, the direction selector determines the sensing output voltage ( $V_{IN}$  or  $V_O$ ) needed for the voltage-mode control. The voltage-mode PI controller then generates the duty ratio  $D$  of the main switch by comparing the sensed voltage with the reference voltage. The mode selector informs the PWM modulator of the operating mode of the proposed converter determined by the sensed  $V_{IN}$  and  $V_O$ . Finally, using the information from the mode selector and voltage-mode controller, the PWM modulator and the switch selector generate four gate signals ( $v_{g1}$ – $v_{g4}$ ), which control the proposed converter.

The proposed converter has six operating conditions that depend on the energy transfer direction and operating modes (see Table I). In each energy transfer direction, one of three operating modes is used (buck, boost, or buck-boost), which depends on the relationship between  $V_{IN}$  and  $V_O$ . However, the operation of the converter can be unstable because abrupt transitions occur between the modes (buck  $\leftrightarrow$  buck-boost or buck-boost  $\leftrightarrow$  boost). Hysteresis control and the switch selector are used to solve this problem. The hysteresis control uses different gains ( $G_{L1}$ ,  $G_{L2}$ ,  $G_{H1}$ ,  $G_{H2}$ ) for smooth transition between modes, and the switch selector determines the main switch that controls the operation and voltage gain  $G$  ( $= V_O / V_{IN}$ ) of the converter in each mode (see Fig. 12). The hysteresis control is explained in the following for operating mode transitions.

1) *Transition From Buck Mode to Buck-Boost Mode:* The transition from buck mode to buck-boost mode occurs at

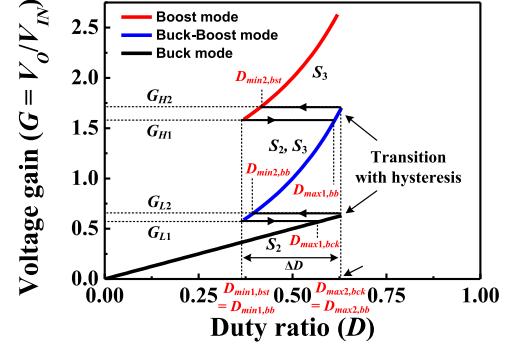


Fig. 12. Voltage gain versus duty ratio of the proposed converter.

$G = G_{L2}$  (@  $D = D_{max2,bck}$ ). For buck-boost mode,  $S_1$ ,  $S_2$ ,  $S_3$ , and  $S_4$  have duty ratios of  $1-D$ ,  $D$ ,  $D$ , and  $1-D$ , respectively, and  $S_2$  becomes the main switch. The duty ratio  $D$  of  $S_2$  is decreased from  $D_{max2,bck}$  to  $D_{min2,bb}$  so that the proposed converter has a voltage gain  $G$  of  $G_{L2}$  in buck-boost mode.

2) *Transition From Buck-Boost Mode to Boost Mode:* The transition from buck-boost mode to boost mode occurs at  $G = G_{H2}$  (@  $D = D_{max2,bb}$ ). For boost mode,  $S_1$ ,  $S_2$ ,  $S_3$ , and  $S_4$  have duty ratios of 0, 1,  $D$ , and  $1-D$ , respectively, and  $S_3$  becomes the main switch. The duty ratio  $D$  of  $S_3$  is decreased from  $D_{max2,bb}$  to  $D_{min1,bst}$  so that the proposed converter has a voltage gain  $G$  of  $G_{H2}$  in the boost mode.

3) *Transition From Boost Mode to Buck-Boost Mode:* The transition from the boost mode to the buck-boost mode

occurs at  $G = G_{H1}$  (@  $D = D_{min1,bst}$ ). For buck-boost mode,  $S_1, S_2, S_3$ , and  $S_4$  have duty ratios of  $1-D, D, D$ , and  $1-D$ , respectively, and  $S_3$  becomes the main switch. The duty ratio  $D$  of  $S_3$  is increased from  $D_{min1,bst}$  to  $D_{max1,bb}$  so that the proposed converter has a voltage gain  $G$  of  $G_{H1}$  in the buck-boost mode.

*4) Transition From Buck-Boost Mode to Buck Mode:* The transition from the buck-boost mode to the buck mode occurs at  $G = G_{L1}$  (@  $D = D_{min1,bb}$ ). For buck mode,  $S_1, S_2, S_3$ , and  $S_4$  have duty ratios of  $1-D, D, 0$ , and  $1$ , respectively, and  $S_2$  becomes the main switch. The duty ratio  $D$  of  $S_2$  is increased from  $D_{min1,bb}$  to  $D_{max1,bck}$  so that the proposed converter has a voltage gain  $G$  of  $G_{L1}$  in the buck mode.

In the opposite energy transfer direction ( $V_O \rightarrow V_{IN}$ ), the transitions between different operating modes are controlled by the same method.

### III. DESIGN CONSIDERATIONS

#### A. Switches $S_1, S_2, S_3$ , and $S_4$

The voltage stress for each switch is  $V_O$  or  $V_{IN}$ , depending on the position of the switch and operating modes. The proposed converter uses four identical power semiconductors as switches. Therefore, the voltage stress  $v_{S,max}$  for a switch is determined as a large value of  $V_{IN}$  and  $V_O$ .

The current stresses for the switches change according to the energy transfer directions and operating modes. The maximum current  $i_{S,max}$  of a switch is equal to  $i_L(t_2)$ , and is expressed in (8), (16), and (25) for buck, boost, and buck-boost operations.

The switch component should be chosen to satisfy  $v_{S,max} < V_{Switch}$  and  $i_{S,max} < I_{Switch}$ , where  $V_{Switch}$  and  $I_{Switch}$  are the maximum voltage and current ratings of the switch component.

#### B. Inductor $L$

The proposed converter operates in DCM to achieve ZVS turn-ON of the switches. To operate in DCM, the minimum value  $i_L(t_0)$  of  $i_L$  should be smaller than 0. For buck operation, the condition  $i_L(t_0) < 0$  and (7) result in

$$L < \frac{(V_{IN} - V_O) DT_S}{2I_{O,max}}. \quad (50)$$

For boost operation, the condition for  $L$  is derived by inserting the condition  $i_L(t_0) < 0$  into (15) as

$$L < \frac{V_{IN} DT_S}{2I_{IN,max}}. \quad (51)$$

In buck-boost operation, the condition for  $L$  is derived by inserting the condition  $i_L(t_0) < 0$  into (24) as

$$L < \frac{V_{IN} DT_S}{2(I_{IN,max} + I_{O,max})}. \quad (52)$$

$L$  should be chosen to satisfy conditions (50)–(52).

#### C. Auxiliary Capacitor $C_a$

For a given allowed voltage ripple  $\Delta v_{o,allow}$ , the condition  $\Delta v_{o,allow} > \Delta v_o$  and (34) result in the following condition in

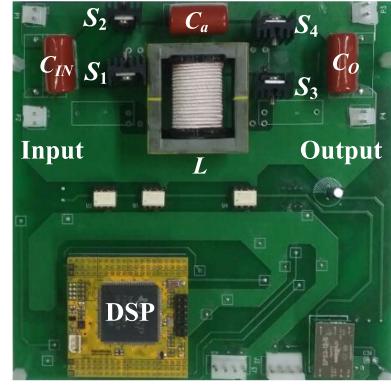


Fig. 13. Photograph of the proposed converter.

buck operation:

$$C_a > \frac{V_O (1 - D) T_S^2}{8L\Delta v_{o,allow}} - C_O. \quad (53)$$

For boost operation, the condition for  $C_a$  is obtained using (38) under the condition  $\Delta v_{o,allow} > \Delta v_o$

$$C_a > \frac{(I_{IN} - I_O + \frac{V_{IN} DT_S}{2L}) [(I_{IN} - I_O) L + \frac{V_{IN} DT_S}{2}]}{2\Delta v_{o,allow} (V_O - V_{IN})} - C_O. \quad (54)$$

In buck-boost operation, the condition for  $C_a$  is obtained using (42) under the condition  $\Delta v_{o,allow} > \Delta v_o$

$$C_a > \frac{(LI_{IN} + V_{IN} DT_S/2)^2}{2L\Delta v_{o,allow} V_O} - C_O. \quad (55)$$

$C_a$  should satisfy conditions (53)–(55) to obtain low  $\Delta v_o$  for all operating ranges.

$C_a$  has different current stresses in different operating modes. The current stresses for  $C_a$  in buck, boost, and buck-boost modes are obtained by inserting  $t = t_2$  into (4), (12), and (21), respectively. The voltage stress  $v_{Ca,max}$  is  $|V_O - V_{IN}|$ .

### IV. EXPERIMENTAL RESULTS

The proposed converter (see Fig. 13) with the digital controller in the digital signal processor was designed to operate at  $V_{IN} = 160$  V,  $V_O = 80 - 320$  V,  $P_O = 16-160$  W, and  $f_S = 45$  kHz. From conditions (50)–(52),  $L = 184$   $\mu$ H was determined to operate in DCM.  $C_a = 3.3$   $\mu$ F was determined by inserting  $\Delta v_{o,allow} = 0.04$   $V_{O,min} = 3.2$  V and  $C_O = 3.3$   $\mu$ F into conditions (53)–(55). The proposed converter was built using the components and circuit parameters in Table III. In addition, a conventional CBB converter of [28] was built with the same specifications for comparison.

The waveforms of  $i_L, V_{Ca}, V_{IN}$ , and  $V_O$  in two energy transfer directions ( $V_{IN} \rightarrow V_O$  and  $V_O \rightarrow V_{IN}$ ) were measured at  $V_{IN} = 160$  V,  $V_O = 80 - 320$  V, and  $P_O = 160$  W (see Fig. 14). Regardless of the energy transfer directions, the value of  $V_{Ca}$  was measured as  $V_O - V_{IN}$  in buck and boost modes and was measured as the ripple voltage of  $C_a$  in buck-boost mode.

The voltage and current waveforms of the switches for the proposed converter were measured at  $V_{IN} = 160$  V,

TABLE III  
VALUES OF COMPONENTS FOR THE EXPERIMENTAL CONVERTERS

Components	Proposed converter	Conventional CBB converter
Core of inductor	EER 4042	EER 4042
Inductor ( $L$ )	184 $\mu\text{H}$	184 $\mu\text{H}$
Input capacitor ( $C_{IN}$ )	3.3 $\mu\text{F}$	3.3 $\mu\text{F}$
Output capacitor ( $C_O$ )	3.3 $\mu\text{F}$	3.3 $\mu\text{F}$
Auxiliary capacitor ( $C_a$ )	3.3 $\mu\text{F}$	
( $S_1$ )	STP28NM60	STP28NM60
( $S_2$ )	STP28NM60	STP28NM60
( $S_3$ )	STP28NM60	STP28NM60
( $S_4$ )	STP28NM60	STP28NM60

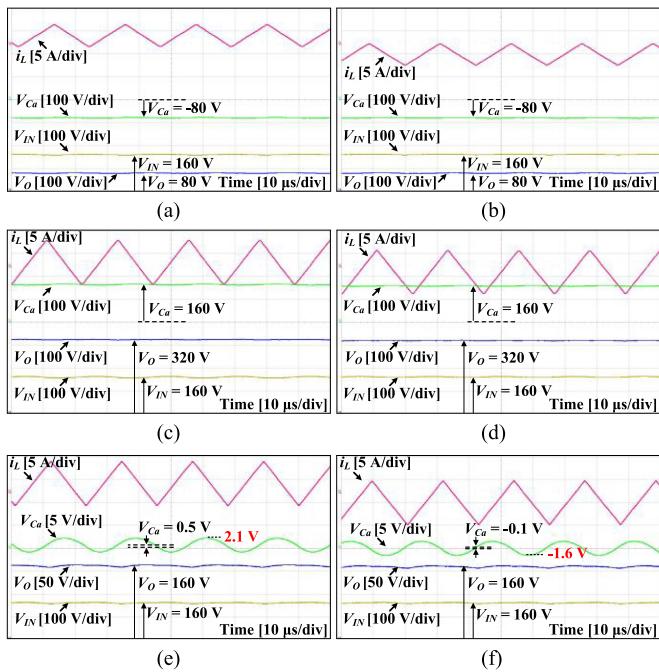


Fig. 14. Waveforms for  $i_L$ ,  $V_{IN}$ ,  $V_O$ , and  $V_{Ca}$  for the proposed converter measured at  $P_O = 160 \text{ W}$  and the energy transfer directions from (a)  $V_{IN} = 160 \text{ V}$  to  $V_O = 80 \text{ V}$ , from (b)  $V_O = 80 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ , from (c)  $V_{IN} = 160 \text{ V}$  to  $V_O = 320 \text{ V}$ , from (d)  $V_O = 320 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ , from (e)  $V_{IN} = 160 \text{ V}$  to  $V_O = 160 \text{ V}$ , and from (f)  $V_O = 160 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ .

$V_O = 80 - 320 \text{ V}$ , and  $P_O = 16 \text{ W}$  (see Fig. 15). Fig. 15(a), (c), and (e) presents the switch waveforms in the energy transfer direction from  $V_{IN}$  to  $V_O$ . Fig. 15(b), (d), and (f) presents the switch waveforms in the energy transfer direction from  $V_O$  to  $V_{IN}$ . At  $V_O = 80 \text{ V}$ , the proposed converter operated in buck mode, and the voltage stresses of the switches were 160 V [see Fig. 15(a)]. At  $V_O = 160 \text{ V}$ , the voltage stresses of the switches were 160 V because the proposed converter operated in buck-boost mode [see Fig. 15(e)]. At  $V_O = 320 \text{ V}$ , the voltage stresses of the switches were 320 V because the proposed converter operated in boost mode [see Fig. 15(c)]. In all operating modes, ZVS turn-ON of the switches was achieved because the proposed converter is operated in DCM. In the opposite direction, the switches were also operated with ZVS turn-ON, and the volt-

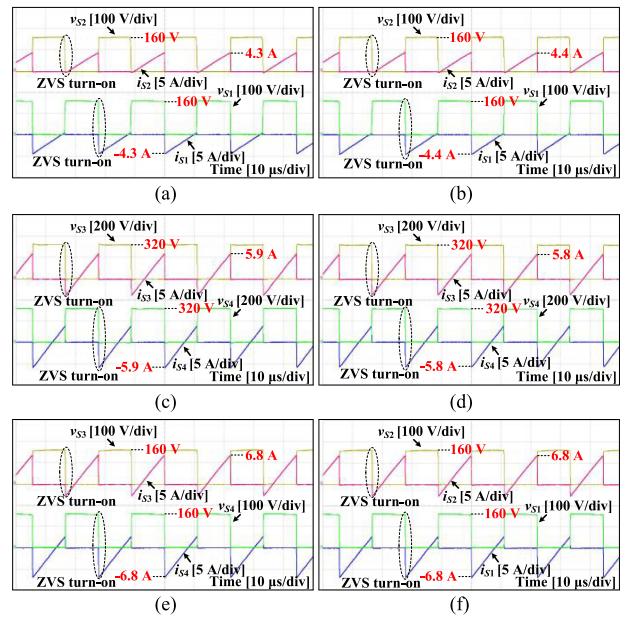


Fig. 15. Voltage and current waveforms for switches of the proposed converter measured at  $P_O = 160 \text{ W}$  and the energy transfer directions from (a)  $V_{IN} = 160 \text{ V}$  to  $V_O = 80 \text{ V}$ , from (b)  $V_O = 80 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ , from (c)  $V_{IN} = 160 \text{ V}$  to  $V_O = 320 \text{ V}$ , from (d)  $V_O = 320 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ , from (e)  $V_{IN} = 160 \text{ V}$  to  $V_O = 160 \text{ V}$ , and (f)  $V_O = 160 \text{ V}$  to  $V_{IN} = 160 \text{ V}$ .

age stresses of the switches became  $V_O$  or  $V_{IN}$  depending on the operating mode [see Fig. 15(b), (d), and (f)].

The current  $i_L$  of  $L$ , the output voltage ripple  $\Delta v_o$ , and the output current ripple  $\Delta i_o$  of the proposed and conventional converters were measured at  $V_{IN} = 160 \text{ V}$ ,  $V_O = 80 \text{ V}$  and  $320 \text{ V}$ ,  $P_O = 160 \text{ W}$ . The energy transfer direction was from  $V_{IN}$  to  $V_O$  (see Fig. 16). Both the proposed and conventional converters had the same  $i_L$  because they used the same  $L$ . However, the proposed converter had lower  $\Delta v_o$  and  $\Delta i_o$  because it distributes  $i_L$  to  $C_a$  and  $C_O$ . When the energy transfer direction was from  $V_O$  to  $V_{IN}$  (see Fig. 17), the proposed converter also had lower  $\Delta v_{in}$  and  $\Delta i_{in}$  than those of the conventional converter.

The power-conversion efficiency  $\eta_e$  of the proposed and conventional converters was measured at  $V_{IN} = 160 \text{ V}$ ,  $V_O = 80 - 320 \text{ V}$ , and  $P_O = 16 - 160 \text{ W}$  (see Fig. 18). When operating in buck and boost mode [see Fig. 18(a) and (b)],  $\eta_e$  of the proposed converter was higher than that of the conventional converter because the proposed converter had lower  $\Delta i_o$  than the conventional converter as a result of using  $C_a$ . However, when operating in buck-boost mode [see Fig. 18(c)],  $\eta_e$  of the proposed converter was almost equal to that of the conventional converter because the proposed converter has similar  $\Delta i_o$  to that of the conventional converter in this mode.

$\eta_e$  of the proposed converter was measured at  $V_{IN} = 160 \text{ V}$ ,  $V_O = 80 - 320 \text{ V}$ ,  $P_O = 16$  and  $160 \text{ W}$ , and  $f_S = 45 - 135 \text{ kHz}$  (see Fig. 19).  $\eta_e$  decreases as  $f_S$  increases, and the maximum difference between the efficiencies measured at the operating frequencies of 45 and 135 kHz is about 5% at  $P_O = 16 \text{ W}$ . These results show that the proposed converter can achieve high  $\eta_e$  by operating with low  $f_S$ .

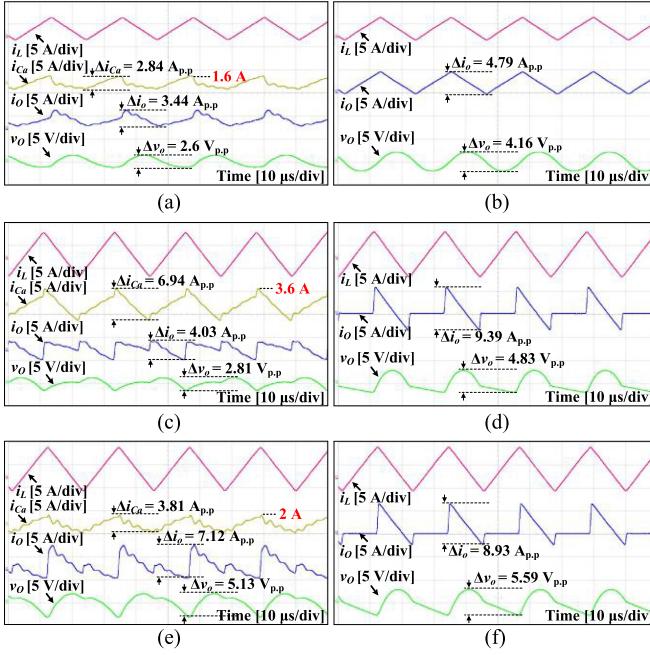


Fig. 16. Waveforms for  $i_L$ ,  $\Delta v_o$ , and  $\Delta i_o$  for the proposed converter measured at  $P_O = 160$  W and the energy transfer directions from (a)  $V_{IN} = 160$  V to  $V_O = 80$  V, from (c)  $V_{IN} = 160$  V to  $V_O = 320$  V, and from (e)  $V_{IN} = 160$  V to  $V_O = 160$  V. For the conventional converter measured at  $P_O = 160$  W and the energy transfer directions from (b)  $V_{IN} = 160$  V to  $V_O = 80$  V, from (d)  $V_{IN} = 160$  V to  $V_O = 320$  V, and from (f)  $V_{IN} = 160$  V to  $V_O = 160$  V.

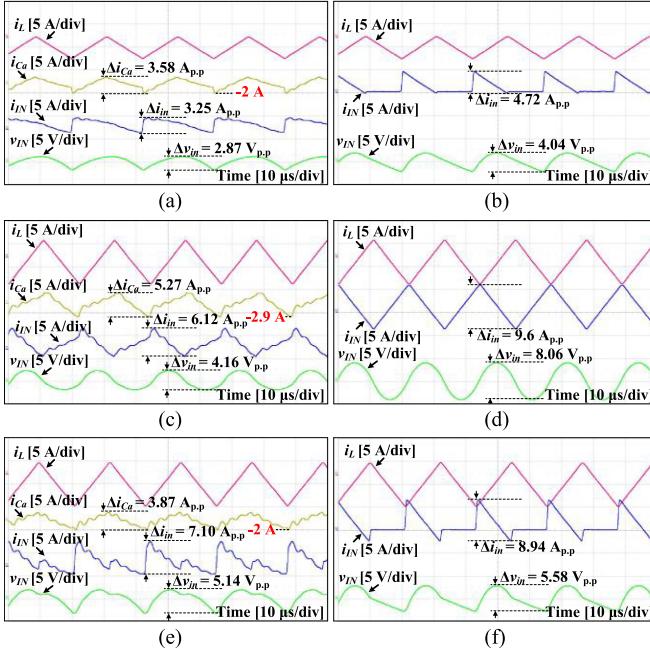


Fig. 17. Waveforms for  $i_L$ ,  $\Delta v_{in}$ , and  $\Delta i_{in}$  for the proposed converter measured at  $P_O = 160$  W and the energy transfer directions from (a)  $V_O = 80$  V to  $V_{IN} = 160$  V, from (c)  $V_O = 320$  V to  $V_{IN} = 160$  V, and from (e)  $V_O = 160$  V to  $V_{IN} = 160$  V. For the conventional converter measured at  $P_O = 160$  W and the energy transfer directions from (b)  $V_O = 80$  V to  $V_{IN} = 160$  V, from (d)  $V_O = 320$  V to  $V_{IN} = 160$  V, and from (f)  $V_O = 160$  V to  $V_{IN} = 160$  V.

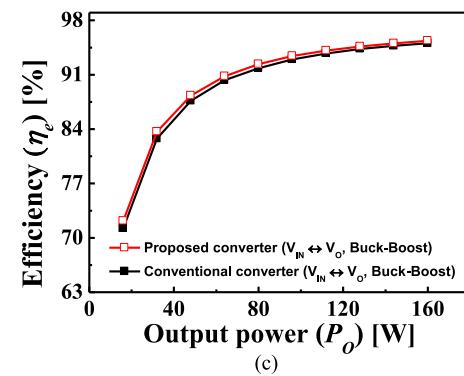
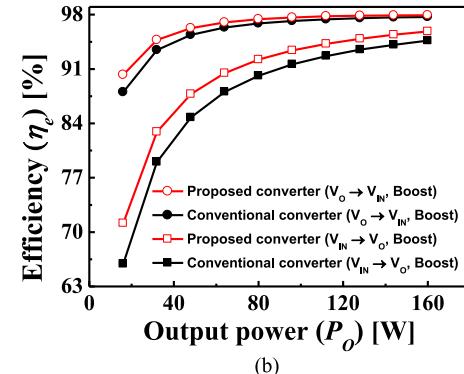
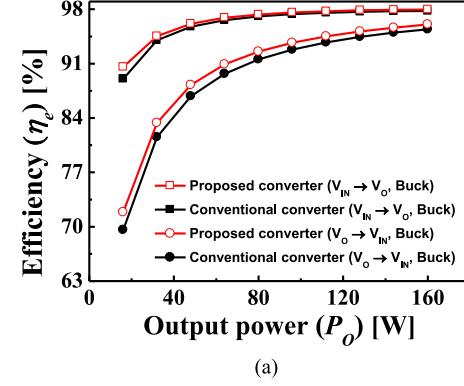


Fig. 18. Power-conversion efficiency versus output power for the proposed and conventional converters when operating in (a) buck mode, (b) boost mode, and (c) buck-boost mode.

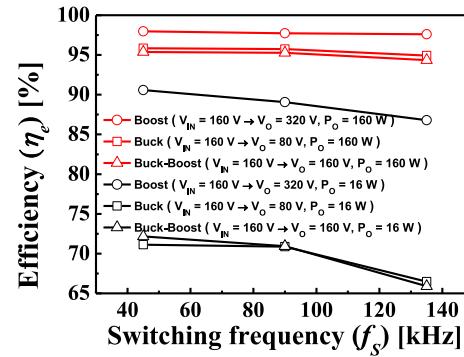


Fig. 19. Power-conversion efficiency versus switching frequency in the proposed converter.

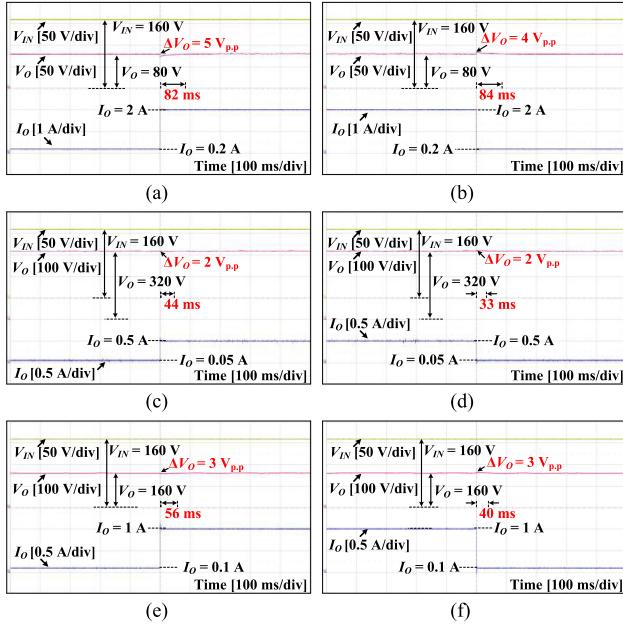


Fig. 20. Transient responses for the proposed converter while changing  $I_O$  (a) from 0.2 to 2 A and (b) from 2 to 0.2 A at the energy transfer direction from  $V_{IN} = 160$  V to  $V_O = 80$  V; (c) from 0.05 to 0.5 A and (d) from 0.5 to 0.05 A at the energy transfer direction from  $V_{IN} = 160$  V to  $V_O = 320$  V; (e) from 0.1 to 1 A and (f) from 1 to 0.1 A at the energy transfer direction from  $V_{IN} = 160$  V to  $V_O = 160$  V.

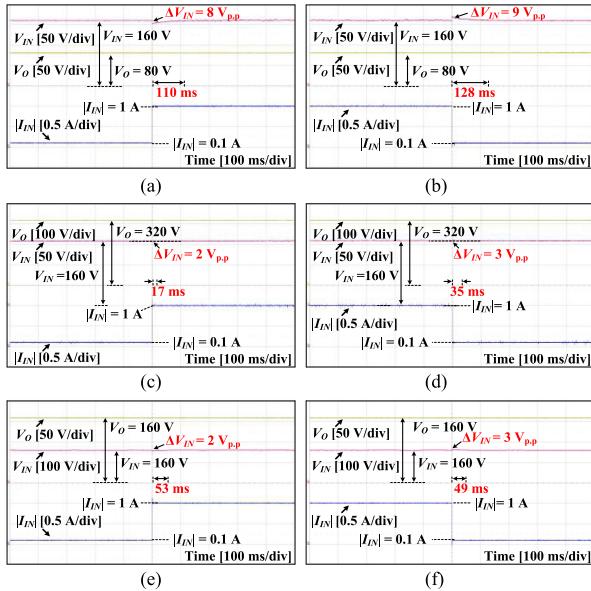


Fig. 21. Transient responses for the proposed converter while changing  $|I_{IN}|$  (a) from 0.1 to 1 A and (b) from 1 to 0.1 A at the energy transfer direction from  $V_O = 80$  V to  $V_{IN} = 160$  V; (c) from 0.1 to 1 A and (d) from 1 to 0.1 A at the energy transfer direction from  $V_O = 320$  V to  $V_{IN} = 160$  V; (e) from 0.1 to 1 A and (f) from 1 to 0.1 A at the energy transfer direction from  $V_O = 160$  V to  $V_{IN} = 160$  V.

The experimental transient responses (see Figs. 20 and 21) to load changes of 10–100% for the proposed converter were measured depending with different energy transfer directions ( $V_{IN} \rightarrow V_O$  or  $V_O \rightarrow V_{IN}$ ) and three operating modes (buck, boost, and buck-boost modes). At the load transition, the

maximum measured voltage spike was 9 V<sub>p,p</sub>, and  $V_O$  returned to steady state within 17–128 ms after the load change. These results show that the proposed converter operated well for a sudden change under load conditions.

## V. CONCLUSION

A new bidirectional buck-boost converter was proposed in this paper. The proposed converter effectively had lower output current ripple than the conventional CBB converter, which was achieved by providing a bypass path for the output current. The reduced output current ripple enabled lower output voltage ripple and higher power-conversion efficiency compared to the conventional converter. The proposed converter had a maximum efficiency of 98% at  $V_{IN} = 160$  V,  $V_O = 80$ –320 V,  $P_O = 16$ –160 W, and  $f_S = 45$  kHz, and the output voltage ripple was less than 5.14 V<sub>p,p</sub>. These results show that the proposed converter is suitable for PV-ESS in a smart grid, which requires a bidirectional buck-boost converter with high efficiency and low ripples in the output voltage and current.

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