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An Isolated Three-Port Bidirectional DC-DC Converter for Photovoltaic Systems with Energy Storage

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An Isolated Three-Port Bidirectional DC-DC Converter for Photovoltaic Systems with Energy Storage

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Abstract—This paper proposes a new isolated, three-port, bidirectional, DC-DC converter for simultaneous power management of multiple energy sources. The proposed converter has the advantage of using the least number of switches and soft switching for the main switch, which is realized by using a LCL-resonant circuit. The converter is capable of interfacing sources of different voltage-current characteristics with a load and/or a DC microgrid. The proposed converter is constructed for simultaneous power management of a photovoltaic (PV) panel, a rechargeable battery, and a load. Simulation and experimental results show that the proposed converter is capable of maximum power point tracking control for the PV panel when there is solar radiation and controlling the charge and discharge of the battery when there is surplus energy and power deficiency with respect to the load, respectively.

Index Terms—Battery, bidirectional DC-DC converter, isolated converter, multiport converter, photovoltaic (PV), soft switching, zero-current switching (ZCS).

I. INTRODUCTION

To integrate multiple DC energy sources of different types to a power grid, multiple independent DC-DC converters are commonly used to step up the time-variant, low-level source voltages to a constant high-level voltage that is required by a grid-tie inverter. Comparing to that solution, a multiport DC-DC converter is preferable owing to the advantages of using fewer components, lower cost, higher power density, and higher efficiency [1], [2].

The multiport converter topologies can be classified into two categories: nonisolated and isolated topologies [3]. Nonisolated multiport converters are usually used in the applications where a low voltage regulation ratio is required [4], [5]. In contrast, in the applications requiring a high voltage regulation ratio, isolated converters which contains a transformer is preferred [6]-[8].

The currently used isolated multiport topologies include the isolated full-bridge converter [6], which uses four controllable power switches for each source, the isolated half-bridge converter [9], which uses two switches for each source, and the isolated single-switch converter [10], which only uses one switch for each source. In some practical applications, energy storage, such as batteries, is commonly used to handle the intermittence of solar and wind energy sources. This requires that at least one port of the multiport converter is bidirectional. The aforementioned topologies are all unidirectional and cannot satisfy such applications [11].

Several bidirectional topologies, such as full-bridge [12], [13] and half-bridge [14], [15] topologies, have been proposed. These two topologies utilize many switches with complicated drive and control circuits. Recently, a three-port topology was proposed by adding one middle branch to the traditional half-bridge converter [16], [17]. It uses less controllable power switches than the half-bridge topology and can achieve zero-voltage switching (ZVS) for all main switches. However, the voltage of the primary source should be maintained at a high value to charge the battery and the battery is both charged and discharged within a switching period. Such a high-frequency charge/discharge has a negative effect on the battery lifetime.

This paper proposes a new isolated, three-port, bidirectional, DC-DC converter. It contains an inductor-capacitor-inductor (LCL)-resonant circuit to achieve zero-current switching (ZCS) for the main switch. Compared with the converter in [17] using five controllable switches, the proposed converter only use three switches; moreover, when using the same renewable energy source to charge a battery, the nominal voltage of the battery connected to the proposed converter can be higher than that connected to the converter in [17]. The proposed converter is applied for simultaneous power management of a photovoltaic (PV) system with a battery in this paper. The PV system and the battery are connected to the unidirectional port and the bidirectional port of the converter, respectively. A maximum power point tracking (MPPT) algorithm is designed for the PV panel to generate the maximum power when solar radiation is available. A charge and discharge controller is designed to control the battery to either absorb the surplus power generated by the PV panel or supply the deficient power required by the load. Simulation and experimental results are provided to validate the proposed converter.
II. TOPOLOGY AND OPERATING PRINCIPLE OF THE PROPOSED CONVERTER

A. Topology of the Proposed Converter

The circuit diagram of the proposed converter is shown in Fig. 1, which consists of a low-voltage-side (LVS) circuit and a high-voltage-side (HVS) circuit connected by a high-frequency transformer. The LVS consists of two ports, an energy storage capacitor \(C_r\), the primary winding of the transformer, and a capacitor \(C_n\), where \(L_p\) includes the added inductance \(L_{pl}\) and the leakage inductance of the transformer \(L_{pr}\). The HVS consists of the secondary winding of the transformer and a full-bridge rectifier implemented with the diodes \(D_{1s} - D_{4s}\). The transformer’s turn ratio is defined as: 
\[
n = \frac{N_p}{N_s},
\]
where \(N_p\) and \(N_s\) represent the numbers of turns of the primary and secondary windings, respectively. Among the switches, \(S_t\) is called the main switch because it not only controls the power generated by the source connected to Port 1 (P1), but also changes the direction of the current flowing through the transformer.

In this paper, the two ports on the LVS are connected to a PV panel and a battery. To simplify the analysis, the proposed converter is analyzed by two separate converters: one is a single-switch LCL-resonant converter [18], and the other is the converter is analyzed by two separate converters: one is a battery-related buck and boost converter consisting of single-switch LCL-resonant converter [18], and the other is the converter operation, define \(V_{dc}\) as the equivalent output voltage of the converter referred to the primary side of the transformer.

Mode 1: \(t \in [t_1, t_2]\) (see Fig. 3). Prior to Mode 1, \(S_t\) is off; the currents through \(L_r\) and \(L_p\) are zero and a positive value of \(I_{s1}\), respectively, i.e., \(i(t_1) = 0, i_p(t_1) = I_1\). When \(S_t\) is on, as shown in Fig. 4(a), \(L_r\) and \(L_p\) resonate with \(C_r\), the current of the inductor \(L_r\) increases and the voltage of the capacitor \(C_r\) decreases. Due to the existence of \(L_{pr}\), the current through the switch \(S_t\) increases slowly so that the switch is turned on under a low \(di/dt\) condition. The resonant frequency and the particular solution in this mode can be expressed as:
\[
\omega^{(i)} = \frac{1}{\sqrt{L_r / L_p}} \cdot C_r
\]

B. Single-Switch LCL-Resonant Converter for PV Panel

In a switching period, the voltages across \(C_1\) and \(C_i\) can be taken as constant values. Particularly, in the steady state, \(V_{C_1} = V_1\), where \(V_1\) is the output voltage of the PV panel. The converter has seven operating modes depending on the states of the switch \(S_t\) and the resonant circuit. Fig. 2 shows the equivalent resonant circuit in different modes. The differential equations of the resonant circuit in Mode \(k\) \((k = 1, ..., 7)\) are

\[
\begin{align*}
\dot{v} &= L_r^{(k)} \cdot \frac{di^{(k)}}{dt} \\
\dot{i} &= C_r \cdot \frac{dv}{dt} + L_r^{(k)}
\end{align*}
\]

where \(v\) represents the voltage of the capacitor \(C_r\); \(L_r^{(k)}\) and \(i^{(k)}\) represents the equivalent resonant inductance and the current through the equivalent resonant inductor in the \(k^{th}\) \((k = 1, ..., 7)\) operating mode, respectively. Then \(v\) can be solved from (1) and has the following form.

\[
v(t) = A^{(k)} \cos[(\omega^{(k)}(t - t_1))] + B^{(k)} \sin[(\omega^{(k)}(t - t_1))] + V^{(k)} (2)
\]

where
\[
\omega^{(k)} = \frac{1}{\sqrt{L_r^{(k)} \cdot C_r}} (3)
\]
is the resonant frequency in Mode \(k\); \(V^{(k)}\) is the particular solution of equation (1) in Mode \(k\), and \(A^{(k)}\) and \(B^{(k)}\) are coefficients, which can be expressed as:

\[
\begin{align*}
A^{(k)} &= v(t_1) - V^{(k)} \\
B^{(k)} &= \frac{I_1 - i_p(t_1) - i(t_1)}{\omega^{(k)} \cdot C_r}
\end{align*}
\]

where \(v(t_1), I_1, i_p(t_1), \) and \(i(t_1)\) represent the voltage across \(C_r\) and the currents of \(L_{1}\) because of a large \(L_1\), \(L_{pr}\), and \(L_r\) at time \(t_1\), respectively. Equations (4) and (5) can be solved from (1) and has the following form.
Fig. 3. Steady-state waveforms of the proposed converter.

\[ V^{(1)} = \frac{L_r}{L_r + L_p} \left( V_p + V_r \right) \]  

(7)

where // represents that \( L_r \) and \( L_p \) are connected in parallel. In this mode, the current through the primary magnetizing inductance \( L_m \) increases, the current \( i_{t2} \) through the secondary side of the transformer is positive, which indicates the conduction of the \( D_{s3} \) and \( D_{d3} \). At the end of Mode 1, \( i_p(t_2) = i_m(t_2) \), \( i \) achieves its maximum value \( I_{max} \), \( i_{t2}(t_2) = 0 \), and \( v_p(t_2) = 0 \), and \( v_p \) changes its polarity from positive to negative.

Mode 2: \( t \in [t_2, t_3] \), during which \( S_1 \) is on, \( i(t) > 0 \), and \( i_p(t) = i_m(t) \). \( I_{max} \) are reverse biased such that \( i_{t2} = 0 \). As shown in Fig. 4(b), \( L_m, L_p \), and \( L \) resonate with \( C_r \).

Since \( L_m >> L_p, L_m >> L_r \), then

\[ \omega^2 = \frac{1}{\sqrt{\left( \frac{1}{L_p} + \frac{1}{L_m} \right) L_r}} C_r = \frac{1}{\sqrt{L_r C_r}} \]  

(8)

\[ V^{(2)} = \frac{L_r}{L_r + L_p + L_m} V = \frac{L_r}{L_m} V_1 \]  

(9)

At the end of Mode 2, \( v_p(t_3) = -V_r, v(t_3) = V_1 - V_r \), the diodes \( D_{s2} \) and \( D_{d4} \) begin to conduct.

Mode 3: \( t \in [t_3, t_4] \), during which \( S_1 \) is on, \( i(t) > 0 \), \( v_p(t) = -V_r \), \( i_{t2} < 0 \). As shown in Fig. 4(c), \( L_r \) and \( L_p \) resonate with \( C_r \); the energy stored in \( L_r \) is released to charge the capacitor \( C_r \); \( v_p \) is clamped to \( -V_r \); and \( i_{t2} \) is negative, which indicates the conduction of \( D_{s2} \) and \( D_{d4} \). Compared to Mode 1, the only difference in the equivalent circuit in this mode is the sign of \( v_p \).

Thus, \( \omega^3 = \omega^1 \), and

\[ V^{(3)} = \frac{L_r}{L_r + L_p} \left( V_1 - V_r \right) \]  

(10)

This mode terminates at time \( t_4 \) when the current of \( L_r \) decreases to zero, i.e., \( i(t_4) = 0 \).

Mode 4: \( t \in [t_4, t_5] \), during which \( S_1 \) is on, \( i(t) < 0 \), \( v_p(t) = -V_r \), \( i_{t2} < 0 \), and \( D_{s2} \) and \( D_{d4} \) conduct. As shown in Fig. 4(d), a negative current flows through the internal diode of the switch \( S_1 \); the gate signal can be removed to turn off the switch, e.g., at time \( t_5 \), under the ZCS condition. The circuit equations are the same as those in Mode 3. Thus, \( \omega^4 = \omega^1, V^4 = V^3 \). At the end of Mode 4, \( i(t_5) = 0 \) and the voltage across the switch \( S_1 \) is the
same as that across the capacitor $C_r$, i.e., $v_{d1}(t_s) = v$.

Mode 5: $t \in [t_s, t_6]$, during which $S_1$ is off, $i(t) = 0$, $v_p = -V_T$, and $i_{r2} < 0$. As shown in Fig. 4(e), $L_m$ and the switch $S_1$ can be neglected in the circuit. The inductor $L_p$ resonates with $C_r$, and the direction of $i_p$ changes from negative to positive. The following can be obtained:

$$\omega^{(5)} = \frac{1}{\sqrt{L_p + L_m}} \cdot C_r$$
$$V^{(5)} = V_1 - V_T$$

At the end of Mode 5, $i_p(t_6) = i_m(t_6)$, $i_{r2}(t_6) = 0$, and $v_p$ changes its polarity from negative to positive.

Mode 6: $t \in [t_6, t_7]$, during which $S_1$ is off, $i(t) = 0$, $v_p(t) = v_m(t)$, $D_{s1}$ and $D_{s2}$ are reverse biased such that $i_{r2} = 0$. As shown in Fig. 4(f), $L_m$ and $L_p$ resonate with $C_r$, and $C_r$ is charged. The following can be obtained:

$$\omega^{(6)} = \frac{1}{\sqrt{L_p + L_m}} \cdot C_r$$
$$V^{(6)} = \frac{V_1}{2}$$

Once $S_1$ is turned on at time $t_6$, Mode 7 switches to Mode 1.

There are five inductances $L_1$, $L_2$, $L_m$, $L_p$, and $L_m$ in the proposed converter that need to be properly designed. $L_m$ is designed based on the following critical inductance $L_{mc}$ [18]:

$$L_{mc} = \frac{V_T \cdot T}{4 \cdot I_{ap, pk}}$$

where $T$ is the switching period of the switch $S_1$; $I_{ap, pk}$ is the peak current through the magnetizing inductor. In this paper, the root mean square (RMS) value of the magnetizing current is designed to be 2% of the RMS value of $i_p$. Then $L_m$ is designed to be larger than $L_{mc}$. Once the transformer is designed, the leakage inductance $L_p'$ of the transformer can be measured.

Given the load resistance $R_L$ and the transformer’s turn ratio $n$, the quality factor $Q$ of this LCL-resonant converter can be calculated [19]:

$$Q = \frac{8 \cdot n^2 \cdot R_L}{\pi^2 \cdot Z}$$

where $Z$ is the characteristic impedance of the resonant circuit defined below:

$$Z = \sqrt{\frac{L_p}{L_m + L_p}}$$

Given the desired value of $Q$ and the value of $R_L$, the value of $Z$ can be calculated from (16). In this paper, $Q$ is selected in the optimal range of [1.5, 5]. Specifically, the value of $Q$ is 3.7 when the nominal load is applied. Then, given the resonant frequency, $C_r$ can be calculated from (6) and (18). Considering the necessary condition $L_p > L_r$, to achieve ZCS [20], $L_r = L_{pr}$ is selected such that the currents through the switch $S_1$ and the transformer are close during the resonant stage. Then $L_r$ and $L_{pr}$ can be calculated from (18) with the measured value of $L_p'$.

The values of $L_1$ and $L_2$ are designed according to their desired current ripples [10]. In this paper, it is expected that the current ripples are within 5% of their nominal currents.

### III. POWER MANAGEMENT OF THE PROPOSED CONTROLLER

Two controllers are needed to manage the power in the LVS. Their objectives are to regulate the output DC-link voltage to a constant value and manage the power for the two sources, respectively. According to the availability of the solar power, there are three working scenarios of the converter, as illustrated in Fig. 5.

#### A. Three Working Scenarios

Scenario 1 ($p_1 > p_{out}$): the available solar power is more than the load demand. As shown in Fig. 5(a), the PV converter works in the MPPT mode; the battery is charged so that the DC-link voltage is controlled at a constant value.

Scenario 2 ($0 < p_1 < p_{out}$): there is solar radiation but the solar power is not sufficient to supply the load. As shown in Fig. 5(b), the PV panel is controlled in the MPPT mode by the MPPT algorithm described later. On the other hand, the deficient power is supplied by the battery, which is discharged by the boost converter, so that the DC-link voltage can be maintained at a constant value.

Scenario 3 ($p_1 = 0$): there is no solar power available and, thus, the battery is discharged to supply the load, as shown in Fig. 5(c). The active switches are $S_1$ and $S_3$.

Proper controllers are designed to manage the power of the system in different scenarios. Fig. 6 shows the overall system with controllers, which include a MPPT controller for the PV panel and charge and discharge controllers for the battery.

#### B. MPPT Controller for PV Panel

The proposed converter is applied for MPPT control of a PV panel using the perturbation and observation (P&O) MPPT algorithm [21] to maximize the PV panel’s output efficiency. Fig. 7 shows the flowchart of the MPPT algorithm. A ratio $r_c$ is defined to specify the relative power change (RPC) of the PV panel between two consecutive sampling steps.

$$r_c = \frac{|P_c(k) - P_c(k-1)|}{P_c(k-1)}$$

where $P_c(k)$ and $P_c(k-1)$ represent the measured output power of the PV panel in the $k^{th}$ and $(k-1)^{th}$ steps, respectively.
It can be seen that for the same power variation value, \( r_c \) is proportional to \( \frac{1}{P_1(k-1)} \). In this paper, the switching period \( T \) will not be changed if the RPC is lower than a predefined value (e.g., \( 10^{-4} \)). As shown in Fig. 7, the P&O MPPT algorithm is realized by the frequency modulation method [22], where the conduction time of \( S_1 \), \( t_{on} \), is fixed so that \( S_1 \) can achieve soft switching.

C. Charge and Discharge Controllers for Battery

Figs. 8(a) and (b) show the equivalent circuit of the battery and the converter when the battery works in the charge and discharge mode, respectively. To simplify the analysis, the battery is modeled as a capacitor \( C_b \) connected in series with its internal resistance \( r_b \). Since \( C_b \) is sufficiently large, the terminal voltage of the battery, \( v_{bat} \), can be calculated as \( V_{boc} - i_{bat} \cdot r_b \), where \( V_{boc} \) is the open-circuit voltage of the battery. Then the transfer function between the battery current \( i_{bat} \) and the duty cycle \( d_2 \) of the switch \( S_2 \) in the charge mode can be derived:

\[
G_c(s) = \frac{i_{bat}(s)}{d_2(s)} = \frac{(V_1 - V_{bat}) \cdot \left( s + \frac{1}{r_b \cdot C_b} \right)}{s^2 + \left( \frac{r_2}{L_2} + \frac{1}{r_b \cdot C_b} \right) s + \frac{r_2}{L_2} + \frac{D_2}{L_2 \cdot C_b}}
\]

where \( V_1 \) and \( V_{bat} \) represent the average voltages of the PV panel and the battery, respectively; \( r_2 \) and \( r_b \) represent the parasitic resistance of the inductor \( L_2 \) and the internal resistance of the battery, respectively; \( D_2 \) is the steady-state value of the duty cycle of the switch \( S_2 \).
Similarly, the transfer function between $i_{bat}$ and the duty cycle $d_1$ of the switch $S_1$ in the discharge mode is:

$$G_d(s) = \frac{i_{bat}(s)}{d_1(s)} = \frac{V_i \cdot \left( s + \frac{1}{r_i \cdot C_2} \right)}{s^2 + \left( \frac{r_2}{L_2} + \frac{1}{r_i \cdot C_2} \right)s + \frac{r_2}{L_2} \cdot \frac{r_i}{r_2} \cdot \frac{1}{L_2 \cdot C_2}}$$ (21)

To control the current of the battery, a proportional-integral (PI) controller, $K_p + K_i/s$, is used in the charge/discharge mode separately, as shown in Fig. 6. Each battery current PI controller takes the current error as the input to generate the duty cycle for $S_2$ or $S_3$ in the charge or discharge mode, respectively. When the reference current $I_0^{bat}$ is zero or negative, the charge PI controller is selected such that $d_2 \geq 0$ and $d_1 = 0$. Otherwise, when the reference current $I_0^{bat}$ is positive, the discharge PI controller is selected such that $d_3 > 0$ and $d_2 = 0$.

The bode plots of $i_{bat}(s)/d_2(s)$ and $i_{bat}(s)/d_3(s)$ without the PI compensations (i.e., the open-loop transfer functions) are shown in Fig. 9. The plots imply that the two open-loop systems have low gains and 0 dB/decade slopes in the low-frequency region. Therefore, the design objective is to increase the low-frequency gains and make them cross the 0 dB line with a −20 dB/decade slope, while maintaining a sufficiently large phase margin (> 45°) and a high crossover frequency. By setting the crossover frequency in the range of one to several hundred Hz with a phase margin of 70°, the charge and discharge PI controllers can be derived. The bode plots of $i_{bat}(s)/d_2(s)$ and $i_{bat}(s)/d_3(s)$ with the PI compensations (i.e., the closed-loop transfer functions) are shown in Fig. 9 as well. After the compensations, the low-frequency gains have been increased and the low-frequency slopes are changed to be −20 dB/decade. The crossover frequencies corresponding to the charge and the discharge controllers are 300 Hz and 400 Hz, respectively.

Once the current controllers are designed, the outer-loop DC-link voltage controller which has a lower cutoff frequency than the current loop is then designed. The PI parameters of the current and voltage controllers used in this paper are listed in Table I.

<table>
<thead>
<tr>
<th>Controller Parameters</th>
<th>$K_p$</th>
<th>$K_i$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current controller (charge)</td>
<td>2.5</td>
<td>5000</td>
</tr>
<tr>
<td>Current controller (discharge)</td>
<td>2.38</td>
<td>6600</td>
</tr>
<tr>
<td>Voltage controller</td>
<td>10</td>
<td>2</td>
</tr>
</tbody>
</table>

IV. SIMULATION RESULTS

Simulations are carried in MATLAB/Simulink to validate the proposed converter and the controllers. The parameters of the converter are as follows: transformer turn ratio $n = 5:14$, $L_r = 3.3\mu H$, $L_p = 3.5 \mu H$, and $C_r = 0.22 \mu F$. A SunWize SW-S55P PV panel is used, whose open-circuit voltage $V_{oc}$ and the short-circuit current $I_{sc}$ are 22 V and 3.15 A, respectively. The nominal voltage and internal resistance, $r_{in}$, of the battery are 7.5 V and 0.16 Ω, respectively. The on-time of the switch $S_1$, $t_{on}$, is 3 µs and the switching frequency varies in a range of 100 kHz to 170 kHz. The resistive load $R_L = 100 \Omega$. The desired DC-link voltage, $V'_{dc}$, and nominal power of the load are 50 V and 25 W, respectively.

To test the dynamic characteristic of the controllers, the solar radiation is step changed to examine the responses of the DC-link voltage and output power of the PV panel, as shown in Fig. 10. Fig. 10(a) shows that the initial solar radiation is zero and there is no power generated by the PV panel, as shown in Fig. 10(c). This indicates that the converter works in Scenario 1 and all of the power is supplied by discharging the battery. Fig. 10(b) shows that the DC-link voltage quickly reaches its reference value of 50 V.

Scenario 1 does not terminate until the solar radiation changes from zero to 400 W/m$^2$ at the 1st second. After that, the maximum power generated by the PV panel is 20 W, which is
less than the load demand of 25 W. Thus, the battery still works in the discharge mode to provide the deficient power required by the load and the variation of the DC-link voltage is negligible during the transition. From 1 second to 1.5 second, the converter works in Scenario 2. The PV panel generates the maximum power as indicated in Fig. 10(c). At 1.5 second, the solar radiation is changed from 400 W/m² to 600 W/m², which corresponds to 32-W maximum power. Then the battery stops discharging and starts to absorb the surplus power generated by the PV panel. It takes some time to change the direction of the battery current, which not only results in an approximately 2-V overshoot in the DC-link voltage, but also leads to the PV power generated less than the ideal maximum power during the transient period, as shown in Fig. 10(c). After 0.3 seconds, both the DC-link voltage and PV power reach the desired value and the ideal maximum power point (MPP), respectively.

To testify the effectiveness of the converter for MPPT control of the PV panel, real-world solar radiation data provided by the National Renewable Energy Laboratory (NREL) [23] was used. The data was collected from the South Table Mountain site in Golden, Colorado on Feb. 7, 2013. The duration of the dataset is 620 minutes with a resolution of one data point per minute. Fig. 11 shows the simulation results of the PV power and the DC-link voltage. As shown in Fig. 11(a), the energy extracted from the PV panel closely follows the ideal MPP by using the proposed converter and MPPT control algorithm. The DC-link voltage is well controlled at its desired value of 50 V. The maximum voltage error is approximately 1.2%, as shown in Fig. 10(b), which occurs around 10:10 am when the MPP increases to 25 W. At that time, the battery switches from the discharge mode to the charge mode. The steady-state voltage error is always less than 0.6%.

V. EXPERIMENTAL RESULTS

The system simulated is constructed in hardware to further validate the proposed converter and control algorithm via experimental studies. Fig. 12 shows the prototype of the whole system. It consists of the proposed DC-DC converter, which is connected to a SunWize SW-S55P PV panel, a battery, and an eZdsp F2812 control board. As shown in Fig. 12, the battery consists of four Samsung ICR18650-28A rechargeable lithium-ion cells, where two cells are connected in series to

Fig. 10. Step responses. (a) The profile of solar radiation, (b) DC-link voltage response, and (c) PV power response.

Fig. 11. Simulation results using the NREL data. (a) The generated PV power and (b) the DC-link voltage.
form a pack and two packs are connected in parallel. The nominal voltage, standard charging current, and capacity of the battery are 7.5 V, 2.8A, and 5600 mAh, respectively. The control algorithm is implemented in a TMS320F2812 DSP located on the control board. The parameters of the system are the same as those used in the simulation. Other parameters of the prototype converter are listed in Table II.

| $L_1$, $L_2$ | 320 μH | $C_1$, $C_2$, $C_3$ | 1000 μF |
| $L_o$ | 75 μH | $C_o$ | 0.22 μF (100VDC) |
| $L_c$ | 3.3 μH | $D_{S_1} - D_{S_2}$ | RHRP1540 |
| $L_{j1}$ | 3.3 μH | $S_1$ | IPB107N3 |
| $L_{j2}$ | 0.2 μH | $S_2$, $S_3$ | FDP5632 |

### A. Steady-State Waveforms

Fig. 13 shows the steady-state waveforms of the resonant components, the switch $S_1$, and the transformer when Port 1 is connected to a 16 V voltage source. As shown in Fig. 13(a), when the switch $S_1$ is turned on, the voltage across $C_o$, $v$, decreases and the current through $L_c$ and $S_1$, $i$, increases. The sinusoidal waveforms of $v$ and $i$ indicate the resonance between $L_c$ and $C_o$. The current through the switch $S_1$ increases slowly such that it is turned on under a low $di/dt$ (10 A/μs) condition. When the current drops to zero, the switch $S_1$ is fully turned off under the ZCS condition. Then the voltage across the switch $S_1$, $v_{ds1}$, is the same as that across the resonant capacitor $C_r$. The peak value of $v_{ds1}$ is 35.8 V, which is approximately twice the input voltage. Fig. 13(b) shows the voltage ($v_p$) and current ($i_p$ and $i_{T2}$) waveforms of the transformer. The direction of $v_p$ is the same as $i_{T2}$ regardless of the state of the switch. This indicates that the power is delivered to the secondary side of the transformer in the entire switching period.

### B. Voltage Regulation Ratio

Owing to the use of the LCL-resonant circuit, the frequency modulation method is used such that $S_1$ can achieve soft switching. The voltage regulation ratio, which is defined as $V_o/V_{in}$, is a function of the switching frequency when the frequency modulation method is used. Fig. 14 shows that the voltage regulation ratio increases almost linearly with the normalized switching frequency $f_s/f_r$, where $f_s$ and $f_r$ are the switching frequency and the resonant frequency, respectively.

### C. Three Scenarios

Fig. 15 shows the waveforms when the converter works in Scenario 1. As shown in Fig. 15(a), the power generated by the PV panel, $P_1$, in the steady state is 36.86 W, which is higher than the load power of 25 W. The negative battery current ($i_{bat}$ = -0.87 A) indicates that the battery works in the charge mode. The DC-link voltage ($v_{dc}$ = 50.18 V) is close to the reference value of 50 V, which demonstrates that the charge controller is effective to maintain a constant DC-link voltage. Fig. 15(b) shows the open-circuit transient responses of the voltage $v_1$, current $i_1$, and power $P_1$ of the PV panel when it is connected to Port 1. In the experiment, $C_1$ is initially fully discharged such that the initial value of $v_1$ is zero; the three switches $S_1$, $S_2$, and $S_3$ are off; and the controllers are deactivated. As shown in Fig. 15(b), $v_1$ increases from zero to its maximum value and $i_1$ decreases from a positive value to zero. As a result, $P_1$ first
increases from zero to its MPP (37.96 W) and then decreases to zero. Based on the results, the power-voltage (P-V) characteristic curve of the PV panel can be generated, as shown on the right hand side of Fig. 15(b). It is assumed that the operating point of the PV panel does not change within three minutes since it was a clear day (Dec. 2, 2013 in Lincoln, NE). As shown in Fig. 15(b), the maximum power of the PV panel is 37.96 W, which is slightly higher than the measured mean value of 36.86 W in Fig. 15(a). Such a small deviation is caused by the P&O MPPT algorithm in which the duty cycle of the main switch $S_1$ varies slightly around the optimal duty cycle from time to time.

Fig. 16 shows the waveforms when the proposed converter works in Scenario 2. (a) The steady-state waveforms; (b) the P-V characteristic curve of the PV panel.

Fig. 17. The waveforms when the converter works in Scenario 3.
works in Scenario 2. As shown in Fig. 16(a), the power \( p_1 \) generated by the PV panel is 18.76 W, which is less than that required by the load. The positive battery current \( (i_{bat} = 1.42 \text{ A}) \) indicates that the battery works in the discharge mode to supply the deficient power to the load. The DC-link voltage \( (v_{dc} = 50.4 \text{ V}) \) is close to the reference value of 50 V, which demonstrates the DC-link voltage is well controlled by the discharge controller. Similarly, the measured PV power is also close to its maximum power 19.5 W shown in Fig. 16(b). Therefore, the proposed converter is capable of MPPT control for the PV panel as long as the solar energy is available.

Fig. 17 shows the steady-state waveforms of the converter when it works in Scenario 3. In this scenario, no solar energy is available. In the experiment, the PV panel was disconnected from Port 1. As shown in Fig. 17, the duty cycle of the switch \( S_1 \) is fixed around 0.42 and \( S_1 \) is active to form the boost converter to discharge the battery. The battery discharges with a current of 3.83 A, increasing from 1.42 A in Scenario 2. Therefore, the battery provides more energy in Scenario 3 than it does in Scenario 2. Meanwhile, the DC-link voltage is also well controlled at the desired value.

D. Comparison with the Hard-Switched Converter

A novelty or advantage of the proposed converter is that the main switch \( S_1 \) is turned off under the ZCS condition by using the LCL-resonant circuit implemented in Port 1. To testify the benefit of using the LCL-resonant circuit, the voltage stress of \( S_1 \) of the proposed converter is compared to that of the corresponding hard-switched converter, which is obtained by removing the LCL-resonant circuit \( L_r \), \( C_r \), and \( L_p \) in the dash-line block of the proposed converter in Fig. 1. Fig. 18 shows the voltage and current waveforms of the switch \( S_1 \) of the hard-switched converter when \( V_p = 16 \text{ V} \). As shown in Fig. 18, the gate signal is removed when the current of the main switch \( S_1 \) is 4 A; therefore, the main switch is not turned off under the ZCS condition and the peak voltage across the main switch is 65.9 V. When turning on the main switch, the value of \( di/dt \) in the hard-switched converter is 118 A/μs. In comparison, in the proposed converter, the main switch is turned off under the ZCS condition and the voltage stress and \( di/dt \) are reduced to 35.8 V and 10 A/μs, respectively, as shown in Fig. 13(a).

The efficiencies of the two converters are measured when the load is supplied by a voltage source connected to Port 1. Fig. 19 compares the measured efficiencies of the proposed soft-switched converter and the hard-switched converter. As shown in Fig. 19, the efficiencies of both converters first increase with the load power, reach the peak values when the load power is around 60 W, and then decrease with the load power when it is higher than 70 W. The proposed soft-switched converter always has a higher efficiency than the hard-switched converter and has achieved the peak efficiency of 94.5%.

VI. CONCLUSIONS

This paper has proposed a new isolated, three-port, bidirectional, DC-DC converter which uses the minimum number of switches. The proposed converter has been used for simultaneous power management of multiple energy sources, i.e., a PV panel and a battery, in this paper. Simulation results have shown that the converter is not only capable of MPPT for the PV panel when there is solar radiation, but also can control the charge/discharge of the battery to maintain the DC-link voltage at a constant value. Moreover, the voltage stress and the value of \( di/dt \) of the main switch have been reduced compared with the corresponding hard-switched converter. The proposed converter is applicable to other types of renewable energy sources, such as wind turbine generators.

REFERENCES


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