A NOVEL IMPROVING ENERGY STORAGE SYSTEM USING BIDIRECTIONAL ISOLATED CONVERTER APPLICATION TO GRID CONNECTED SYSTEM

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Abstract— Bidirectional power flow capability is a key feature of dc–dc converters, permitting flexible interfacing to energy storage devices. Although the proposed converter has an inherent soft-switching attribute, it is limited to a reduced operating range depending on voltage conversion ratio and output current as in conventional. Here proposed novel bidirectional dc-dc converters (BDC) have recently received a lot of attention due to the increasing need to systems with the capability of bidirectional energy transfer between two dc buses. A wide input range bi-directional dc-dc converter is described along with the phase-shift modulation scheme and phase-shift with duty cycle control for photovoltaic cell, in different modes. The delivered power and peak current are analyzed and calculated. The key parameters of the bi-directional dc-dc converter, the relationships between the input voltage, phase-shift angle, ratio of the transformer and leakage inductance are analyzed and optimized. The proposed scheme is simulated using MATLAB/SIMULINK software.

Index Terms—Bidirectional dc–dc converter, current-fed, fuel cell (FC), phase shift, super capacitor (SC).

I. INTRODUCTION

The essential part of the renewable energy system is a storage element. The storage element gathers the energy fluctuations and enables to improve the system dynamic properties. To charge and discharge the storage element, the bidirectional DC-DC converter is used. The DC-DC converter often ensures an electrical isolation between low voltage and high voltage parts of the system, and then the transformer is used. In order to feed the transformer a DC power must be converted into AC power and next rectified to DC power. To minimize the transformer size, weight and cost, the frequency of the AC power should be as high as possible. The frequency increase is limited by the transistor conduction and switching losses. It should be noticed that the main source of the power dissipation is the low voltage side converter because it conducts a high current. So, the main effort of the research is directed to the low voltage converter efficiency.

Bidirectional DC–DC converters are used to transfer the power between two DC sources in either direction. These converters are widely used in applications, such as hybrid electric vehicle energy systems, uninterrupted power supplies, fuel-cell hybrid power systems [1-4], PV hybrid power systems, and battery chargers. Many bidirectional DC-DC converters have been researched. Some literatures research the isolated bidirectional DCDC converters, which include the half-bridge types and full-bridge types. These converters can provide high step-up and step-down voltage gain by adjusting the turns ratio of the transformer. The integration of photovoltaic (PV) power systems and energy storage schemes is one of the most significant issues in renewable power generation technology. The rising number of PV installations due to increasingly attractive economics, substantial environmental advantages and supportive energy policies require enhanced strategies for their operation in order to improve the power supply stability and reliability.

Energy storage system for the PV power generation is addressed in the literature. Some concentrates on the system configurations as well as control strategies whereas some others focus solely on the converter topologies or on the control techniques. In the conventional PV system architecture, the PV power is transferred to the load through a unidirectional and a bidirectional converter where a considerable amount of power loss occurs in each conversion stage. Hence, the system efficiency deteriorates with the increasing number of power conversions. These disadvantages arise from the fact that both of the converters in the conventional system process the PV array output power. In some previous applications, the battery-bank is directly connected to a dc bus without a bidirectional converter. This configuration requires more battery stacks and reduces the system efficiency. Also, the battery life is degraded without proper control of charging and discharging of the battery.

Though series strings of storage batteries provide high voltage, a slight mismatch or temperature difference can cause charge imbalance if the series string is charged as a unit. Such high voltage batteries are expensive and produce more arcing on the switches than the low voltage batteries. Another problem with higher voltage batteries is the possibility of one cell failing. A faulty cell would produce lower voltage, however, in an extreme case,
one open cell could break the current flow. A modified DC-DC boost converter is presented. The voltage gain of this converter is higher than the conventional DC-DC boost converter. Based on this converter, a bidirectional DC-DC converter is proposed. The proposed converter employs a coupled inductor with same winding turns in the primary and secondary sides. Comparing to the proposed converter and the conventional bidirectional boost/buck converter, the proposed converter has the following advantages: 1) higher step-up and step-down voltage gains; 2) lower average value of the switch-current under same electric specifications.

II. OPERATION PRINCIPLES OF THE HYBRID BIDIRECTIONAL DC–DC CONVERTER

As shown in Fig. 1, a BHB [5],[6] structure locates on the primary side of the transformer T1 and it associates with the switches S1 and S2 that are operated at 50% duty cycle. The SC bank as an auxiliary energy source is connected to the variable low voltage (LV) dc bus across the dividing capacitors, C1 and C2.

Fig. 1. Proposed hybrid bidirectional dc–dc converter topology

Bidirectional operation can be realized between the SC bank and the high-voltage (HV) dc bus. Switches S3 and S4 are controlled by the duty cycle to reduce the current stress and ac RMS value when input voltage VFC or VSC are variable over a wide range. The transformers T1 and T2 with independent primary windings as well as series-connected secondary windings are employed to realize galvanic isolation and boost a low input voltage to the HV dc bus. A dc blocking capacitor Cb is added in series with the primary winding of T2 to avoid transformer saturation caused by asymmetrical operation in full-bridge circuit. The voltage doublers circuit utilized on the secondary side is to increase voltage conversion ratio further. The inductor L2 on the secondary side is utilized as a power delivering interface element between the LV side and the HV side. According to the direction of power flow, the proposed converter has three operation modes that can be defined as boost mode, SC power mode, and SC recharge mode. In the SC power mode, only the SCs are connected to provide the required load power. When the dc bus charges the SCs, the power flow direction is reversed which means the energy is transferred from the HV side to the LV side, and thereby the converter is operated under the SC recharge mode.

A. Boost Mode

In the boost mode, the timing diagram and typical waveforms are shown in Fig. 2, where n1 and n2 are the turn ratios of the transformers. The current flowing in each power switch on the primary side is presented, but the voltage and current resonant slopes during the switching transitions are not shown here for simplicity. To analyze the operation principles clearly, the following assumptions are given: 1) all the switches are ideal with anti parallel body diodes and parasitic capacitors; 2) the inductance L1 is large enough to be treated as a constant source; 3) the output voltage is controlled well as a constant; 4) the leakage inductance of the transformers, parasitic inductance, and extra inductance can be lumped together as L2 on the secondary side.

Fig. 2. Timing diagram and typical waveforms in the boost mode

1) Stage 1 (t0–t1): It can be seen that at any time, the voltage across L2 is always V T1b + V T2b − V CO, but V T1b, V T2b, and V CO have different values in different operating intervals. In (t0–t1), S1, S4, and S6 are gated, so V T1b = n1VFC, V T2b = 2n2VFC, and V CO = −Vo/2, and thereby iL2 will increase linearly. Because i T1a+i T2a are negative and iL1 is positive, the current flows through the diode of switch S1. The current paths during this interval are shown in Fig. 3(a).
2) Stage 2 (t1–t2): From t1, the value of \( iT_{1a} + iT_{2a} \) starts to be positive, and thus S4 conducts to carry the current, but S1 may conduct until the value of \( IL_1 \) is smaller than that of \( iT_{1a} + iT_{2a} \). The equivalent circuit is shown in Fig. 3(b).

3) Stage 3 (t2–t3): At t2, S6 is turned OFF. The inductor L2 begins to resonate with the stray capacitors CS5 and CS6. When the voltage across CS5 reduces to zero, the body diode of S5 starts to conduct, so the voltage VCO equals \( V_o/2 \).

4) Stage 4(t3–t4): At t3, S5 is turned ON under zero voltage switching (ZVS). The current paths are illustrated in Fig. 3(c).

5) Stage 5 (t4–t5): At t4, S4 is turned OFF. The inductor L2 begins to resonate with the stray capacitors CS3 and CS4. When the voltage across CS3 reduces to zero, DS3 is, therefore, forward biased. The voltage across the primary winding of T2 is clamped to zero, i.e., \( VT_{2b} = 0 \). Hence, VL2 equals \( VT_{1b} – VCO \) and the current paths are shown in Fig. 3(d).

6) Stage 6(t5–t6): At t5, S1 is turned OFF. The inductor L2 begins to resonate with the stray capacitors of the switches CS1 and CS2. CS1 is charged from approximately 0 V, while CS2 is discharged from 2VFC. The rate of change on voltage depends on the magnitude \( iT_{1a} + iT_{2a} – IL_1 \). At t5, VCS2 attempts to overshoot the negative rail and then DS2 is forward biased. After that, S2 can be turned ON under ZVS.

7) Stage 7(t6–t7): During this interval, \( VT_{1b} = -n_{1} VFC \), \( VT_{2b} = -2n_{2} VFC \), and \( VCO = V_o/2 \), so the primary current decays. UntiLL1 is bigger than \( iT_{1a} + iT_{2a} \), the current starts to flow through the switch S2, and thus the equivalent circuit is shown in Fig. 3(e).

8) Stage 8 (t7–t8): From t7, both \( iT_{1a} \) and \( iT_{2a} \) are to be negative, which makes S3 and S5 conduct. The equivalent circuit is shown in Fig. 3(f). After t8, the second half cycle starts.

The power delivered by this converter can be calculated, referring to the typical waveforms shown in the Appendix, as follows:
Where $\delta$ is the phase-shift angle; $\omega$ is the switching angular frequency; $V_L = n_1 V_F$ and $V_H = V_o/2$, respectively; the duty cycle $d$ is defined as

$$d = \frac{T_{onS3}}{T_s} = \frac{T_{onS4}}{T_s}. \quad (2)$$

When $d = \pi$, $v_{T1b} + v_{T2b}$ will be the waveform with only two voltage levels, and then (1) will be

$$P_o = \frac{V_L V_H}{2 \pi \omega L_2} \cdot \delta (\pi - \delta). \quad (3)$$

When $\delta = 0$, the output power is calculated by

$$P_o = \frac{V_L V_H}{2 \pi \omega L_2} \cdot d (\pi - d). \quad (4)$$

In order to limit the reactive power in the converter, the phase shift angle normally is smaller than $\pi/4$ and thereby the first Sub equation in (1) is more practical to analyze the average power. Based on that, the output power, which is with respect to the base $V_L V_H/2 \pi \omega L_2$, is plotted in Fig. 4. It can be seen that when the duty-cycle control is utilized together with the phase-shift control, at the same input and output voltages, the average power delivered is increased, because the duty-cycle control can limit the required reactive power. But with the duty cycle reducing, the output power increasing is not significant. When the phase-shift angle is larger than 0.6, the delivered average power is decreased, because in fact the duty-cycle control reduces the average voltage across the secondary windings. A close study reveals that because of the BHB configuration the average current stress of $S_2$ is much higher than that of $S_1$, whereas the current stresses of $S_3$ and $S_4$ are kept the same.

![Fig. 4. Relationship between the output power (p.u.) and phase-shift angle/duty cycle.](image)

Referring to the definition in Fig. 2, the ON-time conducting current of each main device is given by

$$i_{S1,ON}(t) = i_{T1a}(t) + i_{T2a}(t) - i_{L1}(t)$$

$$i_{S2,ON}(t) = i_{L1}(t) - i_{T1a}(t) - i_{T2a}(t)$$

$$i_{S3,ON}(t) = i_{T2a}(t)$$

$$i_{S4,ON}(t) = -i_{T2a}(t). \quad (5)$$

From (5), obviously, $S_2$ carries more current than $S_1$, so that devices with different current ratings can be chosen for $S_1$ and $S_2$. Thus, the peak current values of the primary side switches are

$$I_{S1,peak} = \frac{P_o}{\eta V_F} + (n_1 + n_2) \cdot I_{peak}$$

$$I_{S2,peak} = \frac{P_o}{\eta V_F} + (n_1 + n_2) \cdot I_{peak}$$

$$I_{S3,peak} = I_{S4,peak} = n_2 \cdot I_{peak} \quad (6)$$

Where $\eta$ is the efficiency of the converter and $I_{peak}$=max (I1, I2, I3), and I1, I2, and I3 are calculated as follows:

$$I_1 = I_{L2}(t_2) = \frac{\pi V_H + (4\delta - d - \pi) V_L}{2 \omega L_2}$$

$$I_2 = I_{L2}(t_4) = \frac{(\pi + 2\delta - 2d) V_H + (3d - \pi) V_L}{2 \omega L_2}$$

$$I_3 = I_{L2}(t_5) = \frac{(2\delta - \pi) V_H + (\pi + d) V_L}{2 \omega L_2} \quad (7)$$

The ZVS condition can be deduced on the precondition that the anti-parallel diode of switch must conduct before the switch is triggered. Then, the soft-switching conditions for switches $S_1$ and $S_2$, switches $S_3$ and $S_4$, and switches $S_5$ and $S_6$ are related to the magnitude of $i_{T1a} + i_{T2a} - i_{L1}$, $i_{T2a}$, and $i_{L2}$, respectively, i.e., the main devices are turned OFF with a positive current flowing and then the current diverts to the opposite diode which allows the incoming MOSFET to be switched on under zero voltage. Thus, in order to achieve ZVS turn ON, the currents must obey

$$i_{T1a}(t_0) + i_{T2a}(t_0) - I_{L1} < 0; \quad (\text{for } S_1)$$

$$i_{T2a}(t_0) < 0; \quad (\text{for } S_3)$$

$$i_{L2}(t_2) > 0; \quad (\text{for } S_4)$$

$$i_{L2}(t_4) + i_{T2a}(t_5) - I_{L1} > 0; \quad (\text{for } S_6)$$

$$i_{T2a}(t_5) > 0; \quad (\text{for } S_3)$$

$$i_{L2}(t_6) < 0; \quad (\text{for } S_5). \quad (8)$$

Hence, substituting (7) into (8), ZVS constraints with respect to circuit parameters and control variables are deduced as
For a short period of utility power failure in UPS system that can be handled by SCs or during the fuel-cell warming-up stage, the converter will be operated under the SC power mode and the power flows from SC bank to the dc voltage bus as shown in Fig. 5. The timing diagram and typical waveforms in this mode are illustrated in Fig. 6. It can be seen that the typical waveforms are similar with those in the boost mode, but because there is no \( i_L \), the current stresses of \( S_1 \) and \( S_2 \) are completely the same.

Comparing to (8), the ZVS constraints for \( S_1 \) and \( S_2 \) are different in this mode and thereby the ZVS condition can be expressed as

\[
\begin{align*}
V_o &> 2n_1(n_1 + n_2)(\pi + d)\frac{V_d}{Z_\omega} \quad \text{(for } S_1) \\
V_o &< 2n_1(n_1 + n_2)(\pi - 2d)\frac{V_d}{Z_\omega} \quad \text{(for } S_2) \\
\frac{V_o}{V_{dc}} &< 2n_1\frac{\pi + d}{\pi} \quad \text{(for } S_3, S_4) \\
\frac{V_o}{V_{dc}} &> 2n_1\frac{\pi + d - 4d}{\pi} \quad \text{(for } S_5, S_6). 
\end{align*}
\]

(B. SC Power Mode)

The peak current can be expressed by

\[
I_{S1,\text{peak}} = I_{S2,\text{peak}} = (n_1 + n_2) \cdot I_d.
\]

With the same method used in the boost mode, to achieve ZVS turn-ON, the currents must obey

\[
\begin{align*}
ir_{L1}(t_0) + ir_{L2}(t_0) &< 0; \quad \text{(for } S_1) \\
ir_{L2}(t_0) &< 0; \quad \text{(for } S_4) \\
ir_{L1}(t_5) + ir_{L2}(t_5) &> 0; \quad \text{(for } S_2) \\
ir_{L2}(t_5) &> 0; \quad \text{(for } S_3) \\
i_{L2}(t_5) &< 0; \quad \text{(for } S_6). 
\end{align*}
\]

(C. SC Recharge Mode)

As shown in Fig. 5, in the SC recharge mode, the SC will be charged by the HV dc bus which means that the power flows from the HV side to the LV side. The timing diagram and typical waveforms are illustrated in Fig. 8, where the gate drive signal of S5 is leading to that of S1 due to the reversed power-flow direction.

### III. QUASI-OPTIMAL DESIGN METHOD

To increase the conversion efficiency, generally based on the precise mathematic model of the power loss of each component and the converter switching times, the phase-shift angle, and the duty cycle can be calculated to control the converter and make the total power losses minimal [7]. But this method has two critical limitations in practice: 1) performance will suffer when the loss models employed in the circuit and the switching times are not available or not precise; and 2) the controller with the needed phase-shift angle and duty cycle depending on the variable input voltage and output power is complex to design. Hence, a quasi-optimal design is proposed here which includes two design criteria.

1) Minimize the RMS value of \( i_L \) by the phase-shift and duty-cycle control to reduce the conduction losses.
2) Keep the ZVS operation for HV-side switches to reduce the switching losses.

IV. PHOTO VOLTAIC SYSTEM (PV)

A PV system directly converts sunlight into electricity. The main device of a PV system is a solar cell. Cells may be grouped to form panels or arrays. Power electronic converters are usually required to process the electricity from the PV device. These converters may be used to regulate the voltage and current at the load, to control the power flow in grid-connected systems, and for the maximum power point tracking (MPPT) of the device. The solar cell is basically a semiconductor diode exposed to light. Solar cells are made of several types of semiconductors using different manufacturing processes. PV cells can be modeled as a current source in parallel with a diode. When there is no light present to generate any current, the PV cell behaves like a diode, which is shown in Fig 7.

![Fig.7- Simplified Equivalent Circuit Model for a Photovoltaic Cell](image)

V. MATLAB MODELEING AND SIMULATION RESULTS

Here simulation is carried out in two different cases, in that 1). Implementation of Proposed High Efficiency Bidirectional Converter. 2). Implementation of Proposed High Efficiency Bidirectional Converter with PV Source Applications to grid.

Case 1: Implementation of Proposed High Efficiency Bidirectional Converter

![Fig.8 Matlab/Simulink Model of Proposed High Efficiency Bidirectional Converter](image)

Fig.8 shows the Matlab/Simulink Model of Proposed High Efficiency Bidirectional Converter using Matlab/Simulink Platform.

Case 2: Implementation of Proposed High Efficiency Bidirectional Converter with PV Source Applications to Grid.

![Fig.13 Matlab/Simulink Model of Proposed High Efficiency Bidirectional Converter with PV Source with grid.](image)
Fig. 13 shows the Matlab/Simulink Model of Proposed High Efficiency Bidirectional Converter with PV Source using Matlab/Simulink Platform.

Fig. 14 Input Current & Primary Side Current

Fig. 14 shows the Input Current & Primary Side Current of Proposed High Efficiency Bidirectional Converter with PV Source.

Fig. 15 Secondary Side Current

Fig. 15 shows the Secondary Side Current of Proposed High Efficiency Bidirectional Converter with PV Source.

Fig. 16 Secondary Side Voltage

Fig. 16 shows the Secondary Side Voltage of Proposed High Efficiency Bidirectional Converter with PV Source.

Fig. 17 Output Voltage

Fig. 17 shows the Output Voltage of Proposed High Efficiency Bidirectional Converter with PV Source.

Fig. 18 Shows The Output Voltage Waveform of Three Level Inverter.

Fig. 19 Grid Voltage

Above Fig. 19 Shows the Grid Voltage of High step-up DC-DC converter with grid connected system.

VI. CONCLUSION

This paper investigates a bidirectional DC–DC converter, which has simple circuit configuration with non-conventional renewable energy sources. Renewable energy resources (RES) are being increasingly applications to many more systems with help of power electronic conversion technology, by using this technology we achieve high reliability to support the grid connected system as well as standalone system. In general, the energy source in a distributed power scheme is a fuel cell, a micro turbine, or a photo-voltaic cell as well as super capacitors. These energy conversion devices produce a dc voltage, which must be converted to an ac voltage for residential or industrial application. A novel hybrid bidirectional dc–dc converter consisting of a current-fed input port and a voltage-fed input port was proposed and studied. Using the steady-state analysis, the relationship between the voltage gains of the proposed converter was presented to analyze the power flows. The simple quasi-optimal design method was investigated to reduce the current ac RMS current and extend the ZVS range. So, we can conclude that the proposed converter is a promising candidate circuit for the FC and SC applications.

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