New Control Strategy for Bidirectional LLC Resonant Converter in Energy Storage Systems

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Abstract—this paper proposes a topology which is based on the traditional LLC resonant converter and has a symmetrical circuit structure to achieve the bidirectional power flow capability. With the advantage of ZVS and ZCS, reverse energy is eliminated, so high efficiency is achieved of the proposed topology, which can be above 97% at full load. In order to be adoptable in bidirectional applications like batteries and super-capacitors, a new control scheme is also proposed in the paper, by which the power flow direction can be regulated automatically. Simulation and experimental results are given to verify the analysis.

I. INTRODUCTION

In order to save energy and protect the environment, distributed generation and smart grid technologies have become more and more popular. In these applications energy storage systems (ESSs) are needed to keep voltage constant and make the power supply stable in different load conditions. ESSs should have the ability of bidirectional power flow to absorb energy when the power supply is excess, and release it when the power supply is insufficient [1], so bidirectional DC-DC converter is an indispensable component in the energy storage systems for the purpose of bidirectional power flow [2].

Bidirectional DC-DC converters for ESSs should have the characters like high power density, high efficiency and high reliability, and isolated topologies are more attractive for it can achieve high voltage ration and less battery cells have to be connected in series [3]-[5]. Among all the isolated topologies for bidirectional applications, dual active bridge (DAB) converter is the most popular [6]. Though DAB converter has many advantages in the bidirectional applications, the phase shift operating mode of DAB converter leads to high circulating energy, and decreases the efficiency. Various kinds of methods have been proposed to reduce the circulating energy and improve the efficiency. An improved DAB topology using diodes to limit the reverse current was proposed to reduce the circulating energy, but it loses the bidirectional power flow capability [7]. Several control methods with two or more phase shift angles as control variables were proposed to minimize the circulating energy and increase the efficiency in [8]-[10], but the control methods are complex and the turn off loss is still high.

The turn off current can be significantly reduced in LLC resonant converter for its ZCS and ZVS capability, so the turn off loss is reduced and the efficiency is high. A bidirectional LLC resonant topology for vehicular applications was proposed in [11]. However, the topology is still a conventional series resonant converter during backward mode, which is not preferred for wide voltage range application. In [12], a bidirectional CLLC resonant converter with two resonant tanks in the transformer primary side and secondary side respectively was proposed. The extra resonant tank increases the cost and volume of the converter, and the voltage gain is reduced compared to the traditional LLC converter. Besides that, it uses the dead-band control of output voltage to decide the power flow direction, so the voltage variation within the borderline won’t be detected. Furthermore current in the output side has to flow through the body diodes of switches which may cause high conduction loss.

This paper proposes an improved bidirectional LLC resonant converter for ESSs. Switching frequency modulation is used to regulate the output power. All the MOSFETs in the proposed topology can achieve ZVS, and ZCS is achieved for MOSFETs in the output side when the switching frequency is below the resonant frequency. In order to achieve bidirectional power flow and buck/boost operation capability in any mode, an extra inductor is added between the midpoints of two switches legs in transformer primary side, so the proposed topology is symmetrical in forward operation and backward operation. MOSFETs are used in the secondary side instead of diodes, and they are also operated at the same switching frequency with...
MOSFETs in the input side to prevent current from flowing through their body diodes in order to reduce conduction loss and reverse recovery current. High conversion efficiency is achieved by the proposed topology. New control methods are also proposed in the paper to achieve the desired voltage gain, and also to regulate the power flow direction.

II. PRINCIPLE OF UNIDIRECTIONAL OPERATION

The unidirectional operation means that the converter is operated in forward mode (charging mode) or backward mode (discharging mode) separately with energy source in one side and load in the other side.

The proposed LLC resonant converter is shown in Fig. 1. In order to achieve the bidirectional operation capability and remain the advantages of traditional LLC resonant converter in both operating mode, the primary side and secondary side are both full-bridge structures consisting 4 MOSFETs. An extra inductor $L_{m2}$ is added between point A and point B, which is separated from the transformer and the resonant inductor. In forward mode (charging mode), the resonant inductor $L_r$, resonant capacitor $C_r$ and transformer magnetizing inductor $L_{m1}$ form the LLC resonant tank, and the extra inductor $L_{m2}$ is used to help achieve ZVS of primary side switches. In backward mode (discharging mode), $L_{m1}$ is coupled into the transformer. So $L_{m2}$ is served as the resonant inductor to achieve the ZCS for MOSFETs in the output side (primary side), and the transformer magnetizing inductor $L_{m1}$ is used to help achieve ZVS of MOSFETs in the input side (secondary side) correspondingly. If $L_{m2}$ is equal to $L_{m1}$, the parameters of resonant component in forward and backward operation will be same, and the converter is symmetrical. So only operating principles in forward mode is analyzed in detail.

In a unidirectional LLC resonant converter with full bridge in the input side and diodes rectifier in the output side, the waveforms of resonant current and magnetizing current when $f_s < f_r$ is shown in Fig. 2. The resonant current from $t_0$ to $t_1$ can be expressed as follows:

$$i_r(t) = \sqrt{2}I_{rms} \sin(\omega t - \theta) \quad (1)$$

$\theta$ is the initial phase angle of $i_r(t)$ at $t_0$, so

$$i_r(t_0) = -\sqrt{2}I_{rms} \sin \theta \quad (2)$$

For $i_r$ is symmetrical in the two half switching cycle, value of $i_r$ at $t_0$ and $t_2$ is opposite, and during the DCM mode $i_r$ is assumed to be constant approximately, so value of $i_r$ at $t_1$ can be expressed as in (5):

$$i_r(t_1) = \sqrt{2}I_{rms} \sin(\omega t_1 - \theta) = \sqrt{2}I_{rms} \sin \theta \quad (3)$$

At last $t_1$ is solved in (6):

$$t_1 = \frac{1}{2f_r} \quad (4)$$

According to the analysis, the conduction time for MOSFETs in the secondary side is also equal to $1/2f_r$, and then they will go into discontinuous conduction mode.

In order to achieve the bidirectional operation capability of LLC resonant converter, a synchronous control method is proposed in [13]. The gate drive signals of M5, M6, M7 and M8 are same with M1 to M4 respectively with 50% duty cycle, and the current in the output side will always in continuous conduction mode (CCM). The voltage gain with the synchronous control method is shown in (1) and the voltage gain of traditional unidirectional LLC resonant converter is shown in (5):

$$G = \frac{1}{\sqrt{(k+1)\frac{x^2}{x^2} - 2Q\tan \varphi (x^2 - 1) + Q^2 (x^2 - 1)^2}} \quad (5)$$

$$G = \frac{1}{\sqrt{\frac{(k+1)\frac{x^2}{x^2} - Q^2 (x^2 - 1)^2}} + Q^2 (x^2 - 1)^2}} \quad (6)$$

Where $k = \frac{L_{m1}}{L_r}$, $Q = \frac{\pi^2 Z_r}{8n^2 R_n}$, $x = \frac{f_s}{f_r}$.

If the switching frequency is much lower than the resonant frequency, the phase difference between the output voltage and output current will be larger, which will result in a much lower voltage gain. Besides that, the zero current switching is lost, so the switching loss and reverse energy increases. As a conclusion, the synchronous control
method made the bidirectional operation possible, but it also led to some disadvantages.

Based on the analysis before, a new control method is proposed, in which the gate drive signals for MOSFETs in the secondary side in a half switching is set to 1/2\(f_r\) when the switching frequency is below the resonant frequency, then they can turn off when current through them decreases to zero. With this unsynchronized control strategy, MOSFETs in the secondary side can turn off with ZCS, reverse energy is limited, and the voltage gain is same with traditional LLC resonant converter. Besides that, current in the secondary side won’t flow through the body diodes of MOSFETs which will keep the conducting loss low.

Fig. 3 and Fig. 4 show the gate drive signals, waveforms and equivalent circuits when the switching frequency is below, equal and above the resonant frequency \(f_r\) respectively, and arrows in dotted line in Fig. 4 show the actual current flow direction.

A. \(f_s < f_r\)

Waveforms when \(f_s < f_r\) is shown in Fig. 3 (a), and there are three operating modes in a half switching, which are shown in Fig. 4 (a), (b), and (d).

Mode 1 \((t_0 - t_1)\): The equivalent circuit is shown in Fig. 4 (a). This mode is same with the first operating mode with the synchronous 50% duty cycle control scheme described before.

Mode 2 \((t_1 - t_2)\):

The equivalent circuit is shown in Fig. 4 (b). M5 and M8 turn off at \(t_1\) and \(i_s\) is zero at the moment, so M5 and M8 can turn off with ZCS. Then the secondary side is separated from the primary side, \(i_s\) will keep zero in this mode, and energy to keep the output voltage constant is supplied by the output capacitor. In the primary side, \(C_r\), \(L_r\) plus \(L_{m1}\) form...
the resonant period is quite long, current will increase very slowly, so it is assumed to be constant.

Mode 3 (t2 – t3):

The equivalent circuit is shown in Fig. 4 (d). M1 and M4 turn off at t1. \( i_L \) plus the auxiliary inductor current \( i_{L_{aux}} \) will charge the parasitic capacitors of M1 and M4, and discharge the parasitic capacitors of M2 and M3 until the voltage across the parasitic capacitors of M1 and M4 equal to the input voltage, then current in the primary side begin to flow through the body diodes of M2 and M3. Then current in the secondary side begin to flow through the body diodes of M6 and M7, so they will turn on with ZVS.

B. \( f_s = f_r \)

Gate drive signals and waveforms when \( f_s = f_r \) is shown in Fig. 3 (b), it is seen that synchronous control method is adopted. There are only two operating modes in half switching cycle, which are shown in Fig. 4 (a) and (d). When M1, M4, M5 and M8 turn off at t1, Mode 1 ends, \( i_s \) is equal to \( i_{L_m} \), so the “Mode 2” described above is not exist anymore.

C. \( f_s > f_r \)

When the switching frequency is higher than the resonant frequency, gate drive signals for MOSFETs in the secondary side are also same with M1 to M4, which is shown in Fig. 3 (c). There are three operating modes in a half switching cycle, and the equivalent circuits are shown in Fig. 4 (a), (c) and (d).

Mode 1 (t0 – t1):

This operating mode is similar to “Mode 1” when \( f_s \) is below or equal to \( f_r \). It ends when M1, M4, M5 and M8 turn off at t1, and \( i_s \) is higher than \( i_{L_m} \) at that moment.

Mode 2 (t1 – t2):

The equivalent circuit is shown in Fig 4 (c). The parasitic capacitors of M1 and M4 are charged by \( i_{L_{aux}} \) and \( i_s \), so M2 and M3 will turn on with ZVS. Since \( i_s \) is still above zero, current in the secondary side will flow through the body diode of M5 and M8, which will quickly drop to zero at t2, then this mode ends.

Mode 3 (t2 – t3):

\( i_s \) is equal to \( i_{L_m} \) at t2, and then \( i_s \) will change its direction. Since M5 and M8 are already off, current will flow through the body diodes of M6 and M7. Then M6 and M7 can be turned on with ZVS.

As described above, the pulse width of gate drive signals for MOSFETs in the secondary side is related to the switching frequency: when \( f_s < f_r \), the pulse width is equal to \( 1/2f_s \) in a half switching cycle; when \( f_s > f_r \), M5 to M8 will turn on and off synchronously with M1 to M4. With this improved control strategy, no matter what the switching frequency is, all the switches in the topology can achieve ZVS, and ZCS is achieved for MOSFETs in the secondary side when \( f_s < f_r \).

When the converter is operated in backward operation, the primary side is the output side and the secondary side turns out to be the input side. So the gate drive signals for MOSFETs in secondary side and primary side are swapped when \( f_s < f_r \), when \( f_s \geq f_r \), gate drive signals for all the MOSFETs are still with 50% duty cycle. In backward operation, the auxiliary inductor will operate as part of the resonant tank, the converter is also a LLC resonant converter, and MOSFETs in the primary side (output side) can turn off with ZCS.

III. PRINCIPLE OF BIDIRECTIONAL OPERATION

The main function for ESSs in DG is to regulate the output power and keep the DC bus voltage constant. For energy generated in DG like photovoltaic or wind power can’t always keep constant, and the load condition may vary with time, the bidirectional converter in ESS has to regulate the output power in a continuous way, and should automatically change the power flow direction in order to satisfy the power demand in both light load and heavy load.

A new control method is proposed for this bidirectional application.

It is known from the analysis in unidirectional operation when \( f_s < f_r \), with the unsynchronized control method, the voltage gain will be higher than 1. If the input side MOSFETs conduct \( 1/2f_{s} \) second in a half switching cycle, and the output side MOSFETs are operated with 50% duty cycle, the voltage gain will be lower than 1 for the topology is symmetrical.

When the proposed converter is operated in bidirectional mode, G is regarded as \( \frac{V_o}{V_{in}} \), in which \( V_o \) is the in the high voltage side of the converter and \( V_{in} \) is in the low voltage side. When we talk about “charging mode”, it means power flow from high voltage side to low voltage side, while “discharging mode” means that power flow from low voltage side to high voltage side.

![Figure 5. Gate drive signals for MOSFETs with different voltage](image)
Assuming the converter is operated in discharging mode, the gate drive signals for different voltage gain are shown in Fig. 5. If the voltage gain is going to decrease from above 1, the pulse width of gate drive signals for M1 to M4 will keep constant with 50% duty cycle, and gate drive signals for M5 to M8 will become shorter until they are same with the pulse width of gate drive signals for M1 to M4, at that moment the switching frequency \( f_s \) is same with the resonant frequency \( f_r \) and the voltage gain is equal to 1. When the voltage gain needs to keep decreasing to below 1, the switching frequency will begin to decrease, and the pulse width of gate drive signals for M5 to M8 will keep constant with 50% duty cycle, and pulse width of gate drive signals for M1 to M4 will become wider. In close loop operation, \( V_o \) is set to be constant, if \( G > 1 \), increase the switching frequency will reduce the output current, if \( G < 1 \), increase the switching frequency will increase the output current, and if \( G = 1 \), the switching frequency is always equal to the resonant frequency.

The application of the proposed converter in a DG is shown in Fig. 6, in which \( V_d \) is the DC bus voltage, \( V_o \) is the output voltage which is also the load voltage, \( r_0 \) is the internal resistance of ESS, and \( r_1 \) is regarded as the line resistance between DC bus and load. DC bus and the ESS are connected by the proposed converter. Assuming the converter is operated in discharging mode and the voltage gain is above 1 (\( V_o > nV_d \)), if load becomes heavier, current in DC bus will increase, and voltage across \( r_1 \) will also increase, so \( V_o \) decreases. As soon as decrease of \( V_o \) is detected, the switching frequency of the converter decreases, and the output current of ESS will increase, so \( V_o \) will increase again until it reaches the initial value. When load becomes lighter, current and voltage across \( r_1 \) decreases, and \( V_o \) will increase, so increase of \( V_o \) is detected, and the switching frequency begin to increase, then the converter will automatically decrease the output current of ESS until \( V_o \) decreases to the initial value. If \( V_o \) can’t go back to the initial value even when the output current has decreased to zero, which means there is excess energy generated by the renewable resource, so the converter will change to charging operation, and ESS begin to absorb the excess energy from. When \( V_o \leq nV_{dc} \) the operating principle is similar, and will not be described in detail.

IV. EXPERIMENTAL RESULTS

The main circuit is shown in Fig. 7. In order to make a voltage difference between the DC bus and the output of converter to simulate the mismatch in power, a line resistor \( r_1 \) is connected in series with \( V_o \). Which should be known that there is not such a resistor in realistic like DG; it is only used to verify the dynamic process of bidirectional operation in simulation and experiments. Voltage sources are connected with resistors in parallel, so each side of the converter can generate energy and also absorb energy, which is similar to the interaction between ESS and DG. In order not to connect two sources in parallel directly, a clamping diode is connected in series with each voltage source.

The parameters of the converter are shown in table 1. \( k \) is the ratio of magnetizing inductor and resonant inductor, which is as large as possible to reduce the conduction loss, but a larger \( k \) will narrow the output voltage range, in order to meet the voltage range demand, \( k = 6 \) is designed. For the maximum output power in discharging mode is 1 kW, \( R_{w1} = 160 \Omega \) is designed, and \( R_{w2} \) is set to 5 \( \Omega \).

A. Unidirectional experimental results

In the experiments of unidirectional operation, a resistor is adopted as load in the output side. The maximum output power is 1 kW, and the primary side voltage is 400 V, and the secondary side voltage range from 80 V to 130 V. Fig. 8 shows the voltage and current waveforms of MOSFETs, resonant tank and the auxiliary inductor when the converter is operated in forward mode. It is seen that: no matter what the voltage gain is, ZVS can always be achieved; when the voltage gain is above or equal to 1, the secondary MOSFETs turn off with ZCS; the average current of \( L_{w2} \) is zero, and its peak value is relatively small as the inductance is large.
Fig. 9 shows the waveforms in backward operation, it is seen that current is opposite with which in forward mode, and ZVS and ZCS can also be achieved. When the switching frequency is below the resonant frequency, voltage across MOSFETs in the primary side will oscillate in the discontinuous conduction mode as shown in Fig. 9 (a).

B. Bidirectional experimental results

Fig. 10 shows the dynamic process of bidirectional power flow when the voltage gain is above 1, and the control scheme is as shown in Fig. 5 (a). In stage 1, the converter is in steady state operation in discharging mode, so $i_o$ and $i_{m8}$ are above 0. Then $V_d$ in the output side increase to 450 V, and the converter goes into stage 2: $V_o$ will increase with $V_d$ at the beginning, after that the switching frequency is regulated and $V_o$ decreases to its initial value very fast, $i_o$ and $i_{m8}$ are negative and the converter is operated in the charging mode in stage 2. When this 450 V voltage across $V_d$ is removed, the converter goes into stage 3: $V_o$ will decrease first, then get back to 400V against, and $i_o$ and $i_{m8}$ goes back to positive, in this stage the converter is in discharging mode again. As a result, by regulate the switching frequency to regulate the power rate; the change of $V_d$ will not influence the output voltage of the converter. The experimental results are same with the simulation model, and the theoretical analysis is verified.

Fig. 11 shows the waveforms in discharging mode in bidirectional operation, when the voltage gain $G$ is below 1
and the control scheme is shown in Fig. 5 (b). The power flow direction and operating principle of this mode is same with the stage 2 in Fig. 10. It is seen that waveforms of input side and output side in discharging mode with $G < 1$ correspond to the waveforms of output side and input side in charging mode with $G > 1$, for these two operation modes are symmetrical.

When $V_d$ increases gradually from below 400 V to above 400 V, output current $i_o$ will also decreases from positive to negative, and Fig. 12 shows the relationship between the switching frequency and output current, in which the dotted line shows the simulation results and the solid line is the experimental results.

Fig. 13 compares the efficiency of the proposed topology, LLC resonant converter with MOSFETs’ body diodes rectify and DAB converter. The efficiency of these three converters is tested with the same low voltage side voltage (90 V) and high voltage side voltage (400 V). It is obvious that the
efficiency of the proposed topology is much higher than the other two; ZCS is achieved in the output side so it doesn’t have the high circulating energy like in DAB converter, also the conduction loss by using the proposed control scheme is much lower than which using diodes rectifier method.

CONCLUSION

This paper proposes a LLC resonant topology with improved structure, in which MOSFETs are used in the output side instead of diodes to achieve the bidirectional power flow capability, and an auxiliary inductor is added in the primary side to make the topology symmetrical in backward operation. In order to reduce the conduction loss in the output side, MOSFETs in the secondary side also turn on and off at the switching frequency. An unsynchronized control scheme is adopted to achieve ZCS for MOSFETs in the secondary side. With zero current switching in the output side, there will be little reverse energy and turn off loss will also decrease. These characters help to achieve higher efficiency of the proposed topology compared with the traditional isolated bidirectional topologies.

With the improved topology and a new control scheme, the converter can automatically regulate the amount of energy transferred and the power flow direction. If it is used to connect an ESS to DG, no matter load is heavy or light, energy supplied and absorbed is always match, and the load voltage will keep constant, so the proposed topology will be popular in energy storage systems. Simulation model and a 1kW prototype are built and verify the theoretical analysis well.

REFERENCE


