Bi-Directional High-Conversion-Ratio CLLC Resonant Converter with a New Synchronous Rectification Scheme for Low Conduction Loss

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This paper proposes a new synchronous rectification (SR) scheme for frequency-modulated bi-directional CLLC resonant converters. By introducing a small-phase-shift between the two full-bridges and utilizing the switching transient, the proposed scheme effectively avoids the magnetizing current being provided from the low voltage side. This leads to considerable reduction in the conduction losses and current stress of the switches in a high-step-down design. Furthermore, the output voltage characteristic is analyzed using a novel mathematical model, which, comparing to the fundamental harmonic approximation method, can not only acquire a more accurate result of steady-state characteristic, but also calculate an accurate set of initial values for the passive components, which is crucial to determine the proper dead-time and phase-shift. Based on the new model, the boundaries of dead-time and phase-shift are discussed in detail. Finally, a prototype is built, and experiments are conducted to verify the principles of the proposed scheme.

Keywords: CLLC resonant converter, phase shift, synchronous rectification, high-step-down

1. Introduction

Resonant full-bridge converter is a popular choice for high-step-down dc-dc conversion applications, due to its simplicity to achieve high conversion ratio and zero voltage switching (ZVS). Comparing to the common LLC converters, CLLC converter, as shown in Fig. 1(a), has a symmetrical feature desirable for bi-directional operations^{(1)–(4)}, and is hence preferred in various applications, especially the emerging energy storage and dc distribution industries^{(5)–(8)}.

In the conventional control scheme of CLLC resonant converters, the output rectifier switches are turned off and the anti-parallel diodes are utilized⁽¹⁾⁽⁹⁾⁽¹⁰⁾. In this scheme, however, the bi-directional power transition speed is inevitably slow due to the transition of the drive signals. Moreover, in many cases, the diodes produce high conduction loss due to forward voltage drop and hence limit the converter efficiency. Synchronous rectification (SR) is an effective way to solve these issues, which has been widely applied in LLC resonant converters^{(11)–(17)}. In literature⁽¹¹⁾, a synchronous rectification method is proposed to make the on/off behavior of MOSFETs the same as the diode rectifier. In literature⁽¹²⁾, the synchronous rectification is used to deal with problems caused by parasitic parameters of the half bridge. In literature⁽¹³⁾, cutting down the cost of the synchronous rectification is the main target. In literature⁽¹⁴⁾, the synchronous rectification is realized by sending the same PWM signals to the primary and the secondary side. In literature⁽¹⁵⁾, a synchronous rectification method is proposed where PWM signals of the

Fig. 1. Circuit of full-bridge CLLC resonant converter

secondary side lags behind that of the primary side. In literature⁽¹⁶⁾, only the dead time of one of the two half bridges in secondary H bridge is extended. In literature⁽¹⁷⁾, PWM signals of the secondary side lags behind that of the primary side and the dead time of secondary side is larger. But still relevant discussions on the CLLC converters are limited.

Furthermore, the magnetizing current in the transformer must be kept at a certain amplitude for achieving ZVS in resonant converters⁽¹⁸⁾. But in a high-step-down design, as the transformer turns ratio is large, the magnetizing current will be significantly higher if the transformer is magnetized by the low voltage side than by the high voltage side, leading to increased copper loss and conduction loss.

This paper proposes a new SR scheme that has a promising capability of fast bi-directional power transition as well as ensuring the high voltage supply provides the magnetizing current. The target application is the two-stage energy storage system, where The CLLC converter is mainly used to

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Fig. 2. Conventional and proposed SR schemes for CLLC converter

guarantee the high efficiency power conversation. Comparing to the traditional SR scheme, which shows inconsistency in the switching transient process of bi-directional operations, the proposed scheme utilizes the switching transient process. Specifically, the proposed scheme introduces a small phase-shift between the two full-bridges, as shown in Fig. 2. The word “small” refers to the condition that the phase-shift t_p is less than the dead-time t_d . To guarantee the magnetizing current provided by the high voltage side in a high-step-down design, the drives on the low-voltage side are always leading.

This paper is an extension of (19), where the simulation and experimental results show disagreement with the first harmonics analysis-based results, and consequently the phase-shift and dead-time determination is not accurate enough. In addition, a similar idea to introduce phase-shift in the synchronous rectification for the CLLC converter is introduced in (20), but it is generally for half-bridge topology, and again, the gain analysis is based on first harmonics analysis method. In this paper, the principle of the switching transient under the proposed SR scheme is further analyzed. Then, a time domain-based analysis method for steady-state characteristics of the converter is introduced. Based on the improved model, the limitations of phase-shift and dead time are discussed. Furthermore, a 1 kW prototype is built, and the theoretical results are verified through experiments.

2. Principle of Switching Transient

The normalized equivalent circuit of a CLLC converter is illustrated in Fig. 1(b), where L_r , C_{r1} , C_{r2} compose a resonant tank with a resonant frequency of f_r , and L_m is usually more than three times larger than L_r . For the analysis of the switching transient, the following assumptions are made:

- As L_m is significant than L_r , the voltage on L_m is close to a square wave, and therefore, i_m is regarded as a triangular wave in a switching period, and is constant during switching transient;
- As the resonant period is far longer than the switching transient, the voltage variation of C_{r1} , C_{r2} are neglected during the switching transient;
- Before the switching transients, the changing rates of i_1 and i'_2 are generally low, and their influence is ignored in the switching transient analysis.

The switching transient when u_1 , u'_2 change from negative to positive is taken here as an example. According to the initial direction of i'_2 , two circumstances are discussed separately. The key waveforms are presented in Fig. 3, and the corresponding circuit diagrams are demonstrated in Fig. 4.

2.1 $i'_2 < 0$ before Switching Transient Right before time t_0 , the following conditions will be true: $u_1 = -u_{dc1}$, $u'_2 = -u'_{dc2}$, $i_m < 0$, u_L is close to zero, and S_2 , S_3 , S_6 and S_7 are conducting.

(a) $i'_2 < 0$ before switching transient

(b) $i'_2 > 0$ before switching transient

Fig. 3. Key waveforms of switching transient

Stage 1 [t_0, t_1]: At time t_0 , S_6 & S_7 are turned off. With the presence of L_m and L_r , i'_2 cannot have a step change and is hence negative, so it begins to charge the parasitic capacitance of S_6 & S_7 and discharge that of S_5 & S_8 .

Stage 3 [t_2, t_3]: At time t_2 , i'_2 reaches zero, so D_5 & D_8 are turned off, allowing C'_{s2} and L_r to resonate. In this interval, i'_2 will continue rising, and u'_2 starts to decrease.

Stage 2 [t_1, t_2]: At time t_1 , $u'_2 = u'_{dc2}$, and then diodes D_5 & D_8 are turned on. During this time period, i'_2 increases at a high and constant rate, which is mainly decided by the voltage drop on L_s :

$$\frac{di'_2}{dt} = \frac{u_{dc1} + u_{dc'}}{L_r} \dots \dots \dots (1)$$

Stage 3 [t_2, t_3]: At time t_2 , i'_2 reaches zero, so D_5 & D_8 are turned off, allowing C'_{s2} and L_r to resonate. In this interval, i'_2 will continue rising, and u'_2 starts to decrease.

Stage 4 [t_3, t_4]: At time t_3 , $u'_2 = -u'_{dc2}$, so u_L becomes small again, and i_1 , i'_2 will only change at a very slow rate. Moreover, as $i_m < 0$ and $i'_2 > 0$, there should be $i_1 < 0$, meaning that S_2 & S_3 are conducting.

Stage 5 [t_4, t_5]: At time t_4 , S_2 & S_3 are turned off by the controller. Consequently, i_1 begins to charge/discharge the parasitic capacitance C_{s1} of S_1 - S_4 .

Stage 6 [t_5, t_6]: At time t_5 , $u_1 = u_{dc1}$, diodes D_1 & D_4 are turned on. According to the high and constant voltage added

Fig. 4. Key current paths and circuit diagrams during a switching transient for initial value of i'_2 is negative. Directions of currents are marked by red arrows, and directions of voltage drops are marked by blue arrow

Fig. 5. Key current paths and circuit diagrams during a switching transient during t_0-t_4 for initial value of i'_2 is positive

on L_r , i_1 will decrease and i'_2 will increase at a high and constant rate, at a value opposite to that of (1).

Stage 7 [t_6, t_7]: At time t_6 , i'_2 reaches zero, so D_6 & D_7 are turned off and, after another charging process, D_5 & D_8 are turned on. The resonant process of C'_{s2} and L_s decides the final current:

$$i'_2 = -2u_{dd} \sqrt{C'_{s2}/L_r} \dots \dots \dots (2)$$

Stage 8 [t_7, t_8]: During this stage, i_1, i'_2 will be changing at a very slow rate, and ideally i'_2 will still be negative at the end of this period.

Stage 9 [t_8, t_9]: At time t_8 , since D_5 & D_8 are on right before time t_8 , S_5 & S_8 can be turned on under ZVS condition.

Stage 10: At time t_9 , ZVS turning-on can be achieved for S_1 & S_4 .

2.2 $i'_2 > 0$ before Switching Transient If $i'_2 > 0$, then at time t_0 the current will switch from S_6 & S_7 to D_6 & D_7 , and u'_2 will not change. Before S_2 & S_3 turn off, the status is the same to that of Stage 4 in part A, as shown in Fig. 5. Then, the system will respond as [t_4, t_9] described in the previous part.

In summary, for both cases, the switching transient ends at time t_7 , when i'_2 equals to the value specified in (2), which can usually be considered as a very small value. Therefore, when the converter is modulated at a switching frequency f_s close to f_r , which means the current waveforms of the inverter and the rectifier are approximately sinusoidal, the switching transient analyzed above will guarantee that the ac current of the low voltage side is in phase with the corresponding ac voltage. Consequently, the low voltage side will provide almost zero reactive power, and the transformer is fully magnetized by the high voltage side.

3. Steady-State Characteristic

In previous work on resonant converters, the currents in the resonant tank are approximated as simply sinusoidal waves at the resonant frequency, and a fundamental frequency simplification is applied to the analysis of steady-state characteristic⁽¹⁹⁾. However, it is very common that L_m is less than ten times larger than L_r , which will also result in a visible, lower-frequency sinusoidal term in the currents. In fact, when the switching frequency is significantly different to the resonant frequency, the solution acquired in (19) shows that there would be a notable phase-shift between the two ac voltages u_{dc1} and u_{dc2} , which is incorrect since the phase-shift in the synchronous rectification scheme is very small.

It is desired to develop an improved analytical model for

$$\begin{bmatrix} u_{Cr1}(t) \\ u'_{Cr2}(t) \\ i_{Lm}(t) \\ i'_2(t) \end{bmatrix} = \mathbf{A}(t) \begin{bmatrix} u_{dc1}(0) \\ u_{dc2}(0) \\ u_{Cr1}(0) \\ u_{Cr2}(0) \\ i_{Lm}(0) \\ i'_2(0) \end{bmatrix} + \begin{bmatrix} a_{11}(t) & a_{12}(t) & a_{13}(t) & a_{14}(t) & a_{15}(t) & a_{16}(t) \\ a_{21}(t) & a_{22}(t) & a_{23}(t) & a_{24}(t) & a_{25}(t) & a_{26}(t) \\ a_{31}(t) & a_{32}(t) & a_{33}(t) & a_{34}(t) & a_{35}(t) & a_{36}(t) \\ a_{41}(t) & a_{42}(t) & a_{43}(t) & a_{44}(t) & a_{45}(t) & a_{46}(t) \end{bmatrix} \begin{bmatrix} u_{dc1}(0) \\ u_{dc2}(0) \\ u_{Cr1}(0) \\ u_{Cr2}(0) \\ i_{Lm}(0) \\ i'_2(0) \end{bmatrix} \dots \dots \dots (6)$$

$$\begin{bmatrix} u_{Cr1}(0) \\ u'_{Cr2}(0) \\ i_{Lm}(0) \end{bmatrix} = - \begin{bmatrix} u_{Cr1}(T/2) \\ u'_{Cr2}(T/2) \\ i_{Lm}(T/2) \end{bmatrix} = - \begin{bmatrix} a_{12}(T/2) & a_{14}(T/2) & a_{15}(T/2) \\ a_{22}(T/2) & a_{24}(T/2) & a_{25}(T/2) \\ a_{32}(T/2) & a_{34}(T/2) & a_{35}(T/2) \end{bmatrix} \begin{bmatrix} u_{dc2}(0) \\ u'_{Cr2}(0) \\ i_{Lm}(0) \end{bmatrix} - \begin{bmatrix} a_{11}(T/2) & a_{13}(T/2) & a_{16}(T/2) \\ a_{21}(T/2) & a_{23}(T/2) & a_{26}(T/2) \\ a_{31}(T/2) & a_{33}(T/2) & a_{36}(T/2) \end{bmatrix} \begin{bmatrix} u'_{dc1}(0) \\ u_{Cr1}(0) \\ i'_2(0) \end{bmatrix} \dots \dots \dots (8)$$

the proposed SR scheme, not only for a more accurate steady-state characteristics, but also for an accurate set of initial values for the passive components, which is crucial to determine the proper dead-time and phase-shift.

Therefore, in this study, a more accurate model based on the Laplace domain is proposed, where all the harmonics are included in the analytical result. In half of the switching period, S_1 , S_4 , S_5 and S_8 are constantly turned on, and the circuit becomes a simple network of impedances and capacitances. In this system, there are four state variables i_{Lm} , i_{Lr} , u_{Cr1} , u'_{Cr2} , as well as two constant inputs u_{dc1} and u'_{dc2} . The characteristic equation of this circuit is

$$(s^2 - \omega_1^2) \cdot (s^2 - \omega_2^2) = 0 \dots \dots \dots (3)$$

where

$$\omega_1, \omega_2 = \left[\frac{C'_{r2}L_m + C_{r1}(L_m + L_r)}{2} \pm \sqrt{\left(\frac{C'_{12}L_m + C_{r1}(L_m + L_r)}{2} \right) - C_{r1}C'_{r2}L_mL_r} \right]^{0.5} \dots \dots \dots (4)$$

The characteristic equation indicates that the step response r of any input u (either input u_{dc1} , u'_{dc2} or the initial value of i_{Lm} , i_{Lr} , u_{Cr1} , u'_{Cr2}) has only a dc term and two sinusoidal terms with a frequency of ω_1 and ω_2 , which could be written as

$$r(t) = [a_1 \sin(\omega_1 t + \varphi_1) + a_2 \sin(\omega_2 t + \varphi_2) + a_3]u(t) \dots \dots \dots (5)$$

where a_1 , a_2 , a_3 , ϕ_1 and ϕ_2 can be determined either analytically, or numerically by simulation and curve fitting. In this particular problem, the time domain equation can be written as

where

$$A(t) = A_1 \sin(\omega_1 t) + A_2 \cos(\omega_1 t) + A_3 \sin(\omega_2 t) + A_4 \cos(\omega_2 t) + A_5 \dots \dots \dots (7)$$

Then, by performing an analysis on the half switching period when u_1 and u'_2 are constants and positive, the steady-state output voltage u'_{dc2} characteristic can be acquired. Specifically, a set of linear equations about the initial values $i_{Lm,0}$, $i_{Lr,0}$, $u_{Cr1,0}$, $u'_{Cr2,0}$ as well as u'_{dc2} can be listed based on the following conditions:

Fig.6. Output voltage characteristic of the prototype converter

a. The final values (when $t = T_s/2$) of i_{Lm} , u_{Cr1} , u'_{Cr2} are exactly opposite to their initial values;

b. $i_{Lr}(0) = i_{Lm}(0) - i'_{2(0)}$, where $i'_{2(0)}$ is specified in (2);

The integral of i_{Lr} times u_{dc1} in the half switching period equals to the input energy, which is $0.5Pf_s$.

c. The integral of i_{Lr} times u_{dc1} in the half switching period equals to the input energy, which is $0.5Pf_s$.

In other words, the initial values of $u_{dc2}(0)$, $u'_{Cr2}(0)$ and $i_{Lm}(0)$ are acquired by solving the following equation:

where

$$u_{dc1}(0) = U_{dc1} \dots \dots \dots (9)$$

$$u_{Cr1}(0) = \frac{PU_{dc1}}{2f_s C_{r1}} \dots \dots \dots (10)$$

$$i'_{2(0)} = -2u_{dc} \sqrt{C'_{s2}/L_r} \dots \dots \dots (11)$$

To demonstrate the relationship between the output voltage characteristic and circuit parameters, the following variables are introduced: λ is the ratio of the magnetizing inductance to the resonant inductance, i.e., $\lambda = L_m/L_r$, and k_c is defined as $k_c = C_{r1}/(C_{r1} + C_{r2}/n^2)$, which is ratio of the two capacitor impedances. Based on the full harmonic model, the frequency-gain curves of some sample circumstances are acquired, as demonstrated in Fig.6, where λ is set to 4 in all cases and f_N is the normalized frequency. f_N can be calculated as f_s/f_r . f_s is the switching frequency and f_r is the resonant frequency. The result is generally showing agreement with the fundamental harmonics analysis in (19). However, it can also be observed that the output voltage characteristic is more consistent regarding to power when k_c is high,

Table 1. Parameters of CLLC Converter Prototype

Symbol	QUANTITY	Value
U_{dc1}	nominal input voltage	520 V
U_{dc2}	nominal output voltage	70 V
P_0	nominal power	± 1.0 kW
S_1 - S_4	high voltage switches	Cree C2M0080120D
S_5 - S_8	low voltage switches	Infineon IRFP4127
n	transformer turn ratio	34:5
L_r	leakage inductance	150 μ H
L_m	magnetizing inductance	550 μ H
C_{r1}	resonant capacitor 1	0.1 μ F
C_{r2}	resonant capacitor 2	8.0 μ F
C_{s1}	Primary side switch C_{oss}	58 pF
C_{s2}	Secondary side switch C_{oss}	410 pF
f_r	resonant frequency	52 kHz

Fig. 7. Output voltage characteristic of the prototype converter

whereas it is highly dependent to load when k_c is low. Moreover, at low k_c it would be impossible to regulate the voltage under bi-directional power conditions, because the maximum voltage at $P = +1.0$ pu is equal to the minimum voltage at $P = -1.0$ pu. Therefore, it would be desired to keep k_c relatively large, so the output voltage regulation capability will be ensured. Furthermore, a CLLC converter with parameters given in Table 1 is taken as an example. The calculated output voltage u'_{dc2} versus f_s at several given P is plotted in Fig. 7, which shows that for both power directions, the output voltage characteristics are generally close, and u_{dc2} decreases as f_s increases.

4. Phase-Shift and Dead-Time Restrictions

A major feature of the proposed SR scheme is that the claimed benefit of having magnetizing current provided from the high voltage side can be achieved automatically, which does not require any high-accuracy sampling nor any complicated calculation. Still, there are certain restriction boundaries for the phase-shift and dead-time to ensure the proposed SR scheme works properly. According to the principle of operation discussed in Section II, the following two conditions must be satisfied:

- i'_2 must cross over zero during a switching transient, otherwise the reactive current will be supplied by the low voltage side.
- i'_2 must be negative right before S_5 and S_8 turn on to achieve ZVS.

To clearly demonstrate the restriction conditions, the following variables are defined as shown in Fig. 2: t_d is the dead time, t_p is the phase-shift time, and $t_o = t_d - t_p$ is the overlap time. Then, based on the above two requirements, the time intervals of the switching transient must satisfy the following rules:

- If the initial value of i'_2 is negative, then $[t_0, t_2]$ should be sufficiently long to make i'_2 reach zero. In the calculation, the time interval $[t_0, t_1]$ is neglected, and assume the current changes linearly during $[t_1, t_2]$, then the minimum t_p is solved:

$$t_{p,min} = -\frac{L_r \cdot i'_2(T/2)}{u_{dc1} + u'_{dc2}}, \text{ for } i'_2(T/2) < 0 \dots \dots \dots (12)$$

- $[t_4, t_6]$ should be sufficiently long to make i'_2 reach zero. Here, since the current at the beginning of t_4 is low, the interval $[t_4, t_5]$ cannot be neglected, so firstly this interval needs to be calculated. By assuming the current charging and discharging the parasitic capacitance is equal to the magnetizing current, the time length and the variation of the current is acquired:

$$t_{45} = t_5 - t_4 = -\frac{C_{s1}(u_{dc1} + u'_{dc2})}{i_{Lm(0)}} \dots \dots \dots (13)$$

$$\Delta i'_{2,45} = -\frac{t_{45}(u_{dc1} + u'_{dc2})}{L_r} \dots \dots \dots (14)$$

Then, according to the sign of $i'_2(T/2)$, the minimum value of t_o is acquired:

$$t_{0,min} = t_{45} - \frac{L_r(i'_2(T/2) + \Delta i'_{2,45})}{u_{dc1} + u'_{dc2}}, \quad i'_2(T/2) > 0 \dots \dots \dots (15)$$

$$t_{0,min} = t_{45} - \frac{L_r[(u_{dc1} + u'_{dc2}) \sqrt{C'_{s2}/L_r} + \Delta i'_{2,45}]}{u_{dc1} + u'_{dc2}}, \text{ for } i'_2(T/2) < 0 \dots \dots \dots (16)$$

- If $P < 0$, $[t_7, t_8]$ should be short enough so at $t = t_8$ there still has $i'_2 < 0$. In other words, the maximum value of t_o is limited by:

$$t_{0,max} = \frac{\pi}{2} \sqrt{C_{s2}L_x} + \frac{(u_{dc1} + u'_{dc2}) \sqrt{C_{s2}/L_r}}{di'_2(T/2)/dt}, \text{ for } \frac{di'_2(T/2)}{dt} > 0 \dots \dots \dots (17)$$

Furthermore, rules d-e are preferred to be satisfied:

- $[t_3, t_4]$ should be as short as possible to shorten the transient process.
- The times t_p and t_o should be close to rule a and b, e.g., in 100 ns, otherwise the lossy diodes will be conducting for a long time.

By applying rules a-e, the feasible boundaries of dead-time and phase-shift can be determined. As an example, the range of t_p and t_o versus f_s for the prototype converter (Table 1) at full power conditions are calculated, and the results are plotted in Fig. 8. The feasible range of t_p and t_o is highly dependent to f_s and P , but still it is possible to introduce a segmented function to determine a proper set of t_p and t_o .

5. Experiment Verification

A CLLC converter prototype is built, as shown in Fig. 9, with parameters given in Table 1. The experimental waveforms of u_1 , u_2 and i_2 are shown in Fig. 10, which clearly show that under the proposed SR scheme, the low-voltage side current i_2 is reset to zero after every switching transient, and consequently the reactive part of the current is significantly suppressed. The typical switching transient waveforms are provided in Fig. 11, where t_p and t_o are equal. The result shows consistency to the theoretical analysis in section II and the key waveforms in Fig. 3.

Furthermore, a set of constant-load, variable- f_s experiments under constant $u_{dc1} = 520$ V condition were implemented, where u_{dc2} was measured, as plotted in Fig. 12. The result generally is observed in the experimental result, which

Fig. 8. Feasible phase-shift time t_p and overlap time t_o of the prototype converter. Solid color area: preferred range; shaded area: acceptable range; dotted line: one of the feasible t_p and t_o versus f_s segmented functions

Fig. 9. Prototype of CLLC resonant converter

is mainly caused by the voltage drop on MOSFETs and the inductor resistance. Moreover, there is a significant and important difference of the output-voltage characteristic between the two SR schemes: under the traditional scheme, u_{dc2} can hardly reach the nominal 70 V value in all cases. The reason is that the source of i_m differs under $f_s > f_r$ and $f_s < f_r$

Fig. 10. Experimental waveforms of conventional and proposed SR scheme. Cyan line: u_1 ; blue line: u_2 ; magenta line: i_2

Fig. 11. Detail of the switching transient

Fig. 12. Experiment result of output voltage u_{dc2} versus switching frequency f_s

Fig. 13. Experiment result of efficiency versus switching frequency

f_r conditions, leading to different f_s - u_{dc2} characteristic with a limit of u_{dc2} reached near $f_s = f_r$. The proposed scheme, however, has a unified f_s - u_{dc2} characteristic due to the constant source of magnetizing current, and consequently achieving an extended range of u_{dc2} , so the nominal output voltage can be achieved in all cases.

The measured efficiency versus switching frequency is plotted in Fig. 13. Furthermore, the efficiency-output voltage plot is presented in Fig. 14. In practical experimental implementation, we use different constant resistors as the load. The variation of “Approx. Power” is within 50 W which is relatively small and should not have big influence on efficiency comparison. For the measurement in Fig. 13 and Fig. 14, we use a multimeter to measure the voltage and current. The accuracy of the multimeter is 0.01 V and 0.01 A. For the experimental prototype, the value of the input and output capacitors are large enough. So that the ripple of both current and voltage is small and should not have a big influence on efficiency. As a result, We just measure the DC component of the voltage and current and then calculate the efficiency. The figures indicate that at the same switching frequency and power, the proposed SR scheme achieves higher efficiency due to the reduced reactive current on the low voltage side. Consequently, given the same output voltage and power, the proposed SR scheme can generally achieve higher efficiency

Fig. 14. Experiment result of efficiency versus output voltage

than the traditional SR scheme.

6. Conclusion

This paper has proposed a new SR scheme for bi-directional high-step-down CLLC converters. Utilizing the transient of a small phase shift between the full-bridges, this scheme successfully keeps the magnetizing current being constantly provided from the high voltage side and thus reduces conduction loss of the low voltage side. The principle of the switching transient has been analyzed and further verified by experiments. The steady-state voltage gain of the converter is generally consistent under different load and power direction, and thus it is possible to tune the output voltage to nominal value under bi-directional conditions. Still, the output voltage range is narrower than CLLC converters without SR, and the transient process is highly dependent to the circuit parameters, including the non-linear output capacitance of MOSFETs, bringing complexity to the selection of dead-time and phase shift.

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