New Model Dc –Dc Converter With Soft Switching in Terms of Zero Voltage and Reducing Current Switching

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***Abstract*—In this paper one new model bidirectional, non-isolated dc-dc converter is introduced. This new model introduced can operate in terms of zero voltage (ZVS) and fixed switching frequency without consideration of direction of flow. To provide condition of ZVS for switches in this model, we use a simple auxiliary circuit included one inductor coil coupled to the main coil and auxiliary coil. Because of the operation of ZVS switches, the diode reverse recovery problem does not occur. In addition, to provide a soft switching in this circuit, any additional switches are not used. Therefore, circuit becomes easier to implement and control. In order to confirm the theory analysis of bidirectional dc-dc converter offered, the experimental and simulation result are presented.**

***Keywords*—non-isolated, zero voltage switching (ZVS), bidirectional dc-dc converter, pulse width modulation (PWM), Buck-Boost converter.**

I. INTRODUCTION

Today, bidirectional dc-dc converter has become an important issue in power electronic. This converter to transfer power between two dc sources is used in two directions. Though this converter are able to reverse flow direction current and power, their voltage polarity of two dc source remain without change. Considering the points mentioned, this converter is used widely in many industrial applications such as hybrid electric vehicle energy system, uninterruptible power supplies, fuel cell hybrid power systems, photovoltaic hybrid power systems and battery chargers[1]-[4].Bidirectional dc-dc converter can be divided into two groups non-isolated, isolated. Many bidirectional, isolated dc-dc converter is presented for example full bridge [4]-[6] and half bridge [7]-[9]. When both sides of circuit cannot be connected to earth and required high voltage gain, bidirectional, isolated converter is used [15]. If high voltage ratio and electrical isolation are not required, bidirectional, non-isolated converters are more favorable because of the simple structure and control, higher efficiency and power density [10]-[12]. In addition, because of the number of active switch and fewer passive components, the cost of this converter is less than isolated converters [11]. In order to reduce the size and weight of bidirectional dc-dc converters, switching frequency must be high. But increasing switching frequency in hard switching converters causes increasing electromagnetic interference (EMI) noise and switching losses. Many techniques have been proposed to solve this problem. Some circuits use the technique of resonant and quasi-resonant [13]-[14]. In these converters, the condition of soft switching is provided for switches. But, high resonance causes current and voltage stresses. And, the possibility of the optimum design of the magnetic elements is not provided due to variable frequency control. Active Clamp technique is used in some circuits [11]-[12]. Also, the condition of soft switching is provided for switches in Active Clamp dc-dc converters. But, the stress of the semiconductor elements and losses is limited the duty cycle using of this technique for bidirectional applications. Therefore, an auxiliary circuit is used to provide the condition of soft switching in many converters [10]-[15]. This auxiliary circuit added has one or more switches. Bidirectional, non-isolated dc-dc converters need at least two main switches for bidirectional power transmission. With the addition of an auxiliary circuit which is comprised multiple switches, the condition of the soft switching is provided. But, the complexity of the control circuit, the conduction losses and overall costs are increased.

In this paper, a usual buck-boost bidirectional dc-dc converter is proposed. To provide the condition of soft switching in this converter uses a simple auxiliary circuit without any additional switches that can solve the reverse recovery body diode problem of switches. This auxiliary circuit includes an inductor coil coupled to the main coil and auxiliary coil. The converter switches are controlled by pulse modulation (PWM). Both switches are switching complement each other, therefore, it is very simple to implement its control circuit.

In this paper, a new bidirectional dc-dc converter in chapter 2 and theoretical analysis is presented in chapter3. The transducer design considerations are presented in section4. The simulation and experimental results of the converter is given in section5.

II. DESCRIBE THE PROPOSED BIDIRECTIONAL DC-DC CONVERTER

The bidirectional dc-dc converter proposed is shown in fig 1. This converter is very similar to conventional converter, except that the primary inductor coil L1 to L2 coupled and auxiliary inductor L3 is added. This auxiliary circuit provides the condition of ZVS irrespective of the direction of power flow converter. Coupled inductor L1 is modeled as magnetic inductance L_m and an ideal transformer having turn ratio equal $N_2/N_1 = n$. Leakage inductance of the coupled inductor L1 is concluded in the auxiliary inductor L3. Diodes D1 and D2 are intrinsic body diodes of the main switches S1 and S2. Capacitors C1 and C2 are parasitic capacitors of S1 and S2.

III. PERFORMANCE OF THE PROPOSED BI-DIRECTIONAL CONVERTER

Buck- Boost bidirectional converter can act as Buck-Boost both the direct and inverse. But, this circuit is assumed that converter operates the Boost in the direct and the Buck in the reverse. In this case, the duty cycle and the drive circuit switches are the same in both situations. The main waveforms of the proposed converter for boost mode are shown in fig 2. Also, each state equivalent circuit for this mode is shown in fig 3.

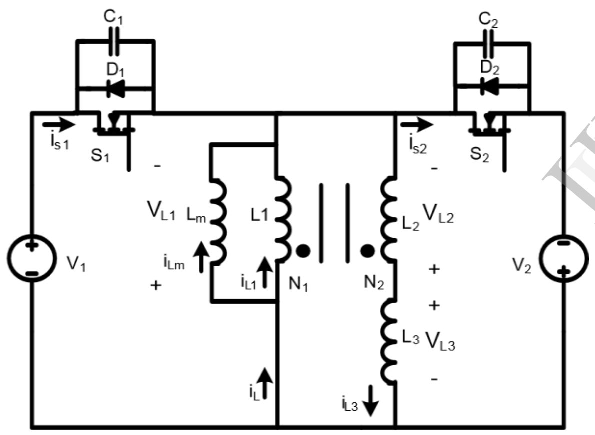

Fig1. The proposed bi-directional dc-dc converter

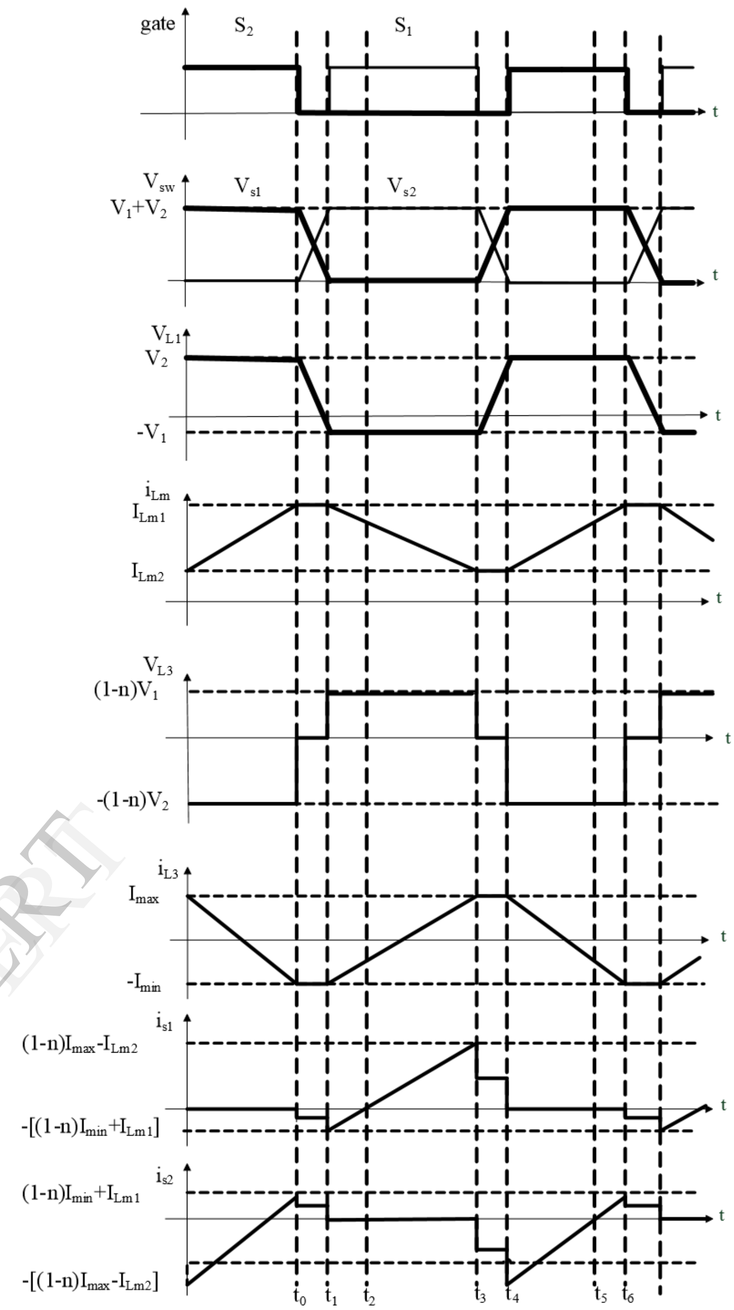

Fig 2.The main waveforms of the proposed converter in boost mode

A. performance of the proposed converter in boost mode

The proposed circuit in boost mode has six different functions in a switching period (T_s=t_6-t_0). As fig 2. shows, it is assumed that the switch $_{S2}$ is conducting before the time t_0. Therefore, the current of magnetic inductance L_m increases linearly and the current of auxiliary inductor L3 decreases linearly. These currents have the maximum and minimum values in I_{Lm1} and -I_{min} at t_0, respectively.

Step1 [t_0,t_1] : This situation begins when the S2 switches turn off. Since the coupled inductor L1 is equal to niL3. Current of is2 at t0 is equal to:

$$i_{s2}(t_0) = i_{s1}(t_0) + i_L(t_0) + i_{L3}(t_0) = (1-n)I_{min} + I_{Lm1} \qquad (1)$$

It is assumed that capacitors C1 and C2 are very small, duration of this situation is very small and all currents are considered constant. Therefore, V_{S2} and V_{S1} voltages change linearly in this situation. And, this time can be calculated as follow:

$$\Delta t_1 = \frac{(C_1 + C_2)(V_1 + V_2)}{(1-n)I_{min} + I_{Lm1}} \qquad (2)$$

Step 2 [t_1,t2]: the V_{S1} voltage reaches zero at the time t1 and the D1 body diode begins to conduct. Form this moment the pulse gate can be applied. Since the V_{L1} voltage is equal to $-v1$, current of Lm increases linearly.

$$i_{Lm}(t) = I_{Lm1} - \frac{V_1}{L_m}t \qquad (3)$$

Because auxiliary coil voltage L3 is equal to (1-n1)V, current iL3 increases linearly.

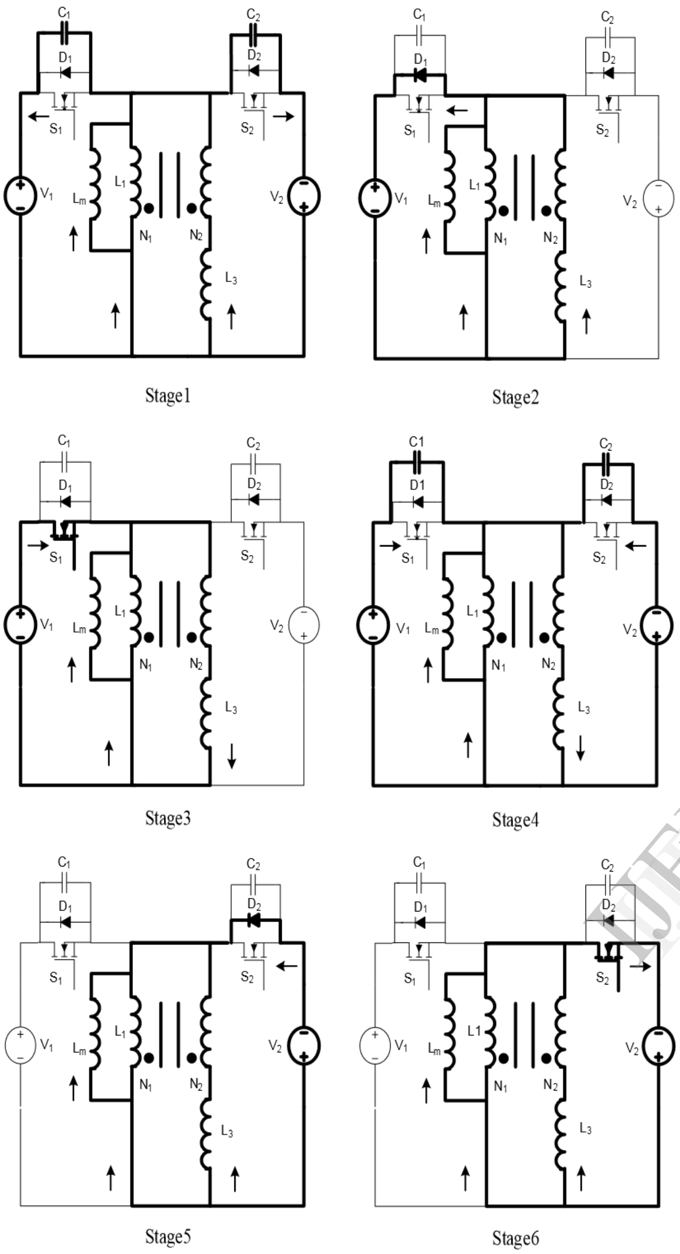

Fig 3. The status of the proposed converter in boost mode

$$i_{L3}(t) = -I_{min} + \frac{(1-n)V_1}{L_3}t \qquad (4)$$

Also, current i_L can be obtained from (3) and (4) as follows:

$$i_L(t) = I_{Lm1} - \frac{V_1}{L_m}t - nI_{min} + \frac{n(1-n)V_1}{L_3}t \qquad (5)$$

And switch current S1 is equal to:

$$i_{s1}(t) = -I_{Lm1} - (1-n)I_{min} + \left(\frac{(1-n)^2V_1}{L_3} + \frac{V_1}{L_m}\right)t \qquad (6)$$

This mode ends when the current of switch S1 reaches zero, therefore, the time of this mode is obtained from the following equation.

$$\Delta_{t2} = \frac{I_{Lm1} + (1-n)I_{min}}{\left(\frac{(1-n)^2V_1}{L_3} + \frac{V_1}{L_m}\right)} \qquad (7)$$

Step3 [t_2,t3]: In this mode, D1 body diode current is transferred to the switch S1 and the same slope of previous mode is continuing. Since, the body diode has turned on until the S1 is on, switch S1 is turned on in terms of ZVS. At the end of this state, inductor current iL3 reaches its maximum value at I_{max} and magnetic current i_{Lm} reaches its minimum value at I_{Lm2}.

Step4 [t_3,t4]: this state begins when the switch S1 turns off. At this moment, current iS1 is equal to:

$$i_{s1}(t_3) = (1-n)I_{max} - I_{Lm2} \qquad (8)$$

This current begins to charge C1 and discharge C2 and this state ends when capacitor C2 discharges fully. Therefore, the time of this situation can be considered as follows:

$$\Delta_{t4} = \frac{(C_1 + C_2)(V_1 + V_2)}{(1-n)I_{max} - I_{Lm2}} \qquad (9)$$

Step5 [t_4,t5]: this state begins when the voltage V_{S2} becomes zero and D2 body diode turns on. Then, gate pulse of S2 switch applies. In this mode, V_{L1} voltage is equal to V2. Therefore, magnetic inductor current increases linearly.

$$i_{Lm}(t) = I_{Lm2} + \frac{V_2}{L_m}t \qquad (10)$$

Since the voltage of across secondary winding coupled inductor L1 is nV2. Therefore, the across voltage of auxiliary inductor VL3 is equal to –(1-n)v2 . So, the current iL3 decreases linearly:

$$i_{L3}(t) = I_{max} - \frac{(1-n)V_2}{L_3}t \qquad (11)$$

From (10) and (11) , the current iL is calculated by according to (12):

$$i_L(t) = I_{Lm2} + \frac{V_2}{L_m}t + nI_{max} - \frac{n(1-n)V_2}{L_3}t \qquad (12)$$

Also, the current switch S2 is equal to:

$$i_{s2}(t) = I_{Lm2} - (1-n)I_{max} + (\frac{V_2}{L_m}t + \frac{(1-n)^2V_2}{L_3})t \qquad (13)$$

Step6 [t5,t6]: In this mode, the current of body diode D2 is transferred to switch S2 and the same slope of previous mode is continuing. Since, the voltage Vs2 remains zero at the moment of turning on S2, switch S2 turns on in terms of ZVS. At the end of this state, the current iL3 reaches its minimum value at –Imin and the magnetic current reaches its maximum value at iLm1. Therefore, the time of this state is equal to:

$$\Delta_{t6} = \frac{I_{Lm1} + (1-n)I_{min}}{\frac{(1-n)^2V_2}{L_3} + \frac{V_2}{L_m}} \qquad (14)$$

B. performance of the proposed converter in the Buck mode

The performance of proposed converter buck mode is similar to the performance of boost mode, except that the flow direction magnetic current iLm and the buck mode current iL is the opposite.

IV. DESIGN CONSIDERATIONS

A. inductor design L3

In the proposed converter, the main inductor current ripple iL decreases rather than conventional bidirectional buck-boost converter. The current characteristic of main coil without ripple is obtained by using simple auxiliary circuit.

From (5) and (12), flow current without can be calculated:

$$L_3 = n(1-n)L_m \qquad (15)$$

B. Imax and Imin values

As seen in figure3, the values of Imin and Imax are the same. Thus, the status of 2 and 5, Imax and Imin can be calculated.

$$I_{max} = I_{min} = \frac{(1-n)V_1}{2L_3}DT_s \qquad (16)$$

C. voltage gain of proposed converter

Voltage gain of proposed converter is equal to conventional bidirectional dc-dc converter. According to the balance voltage – second on the across voltage waveforms of the magnetic inductor Lm, voltage gain is calculated as follows:

$$\frac{V_2}{V_1} = -\frac{D - D_{loss}}{1 - D - D_{loss}'} \qquad (17)$$

Where $D=(t_3-t_1)/T_s$ and the values of D_{loss}' and D_{loss} are calculated in boost mode, it has the same result in buck mode. From figure 3 and equation (6) Dloss can be calculated as follows:

$$D_{loss} = \frac{\Delta_{t2}}{T_s} = \frac{I_{Lm1} + (1-n)I_{min}}{\frac{(1-n)^2V_1}{L_3} + \frac{V_1}{L_m}}f \qquad (18)$$

Where f is the switching frequency.

Also, from figure 3 and equation (13), $D_{loss'}$ can be calculated as follow:

$$D_{loss}' = \frac{\Delta_{t6}}{T_s} = \frac{I_{Lm1} + (1-n)I_{min}}{\frac{(1-n)^2V_1}{L_3} + \frac{V_2}{L_m}}f \qquad (19)$$

Also:

$$D_{loss}' = D_{loss}\frac{V_1}{V_2} \qquad (20)$$

Equation (20) is replaced in equation (17):

$$\frac{V_2}{V_1} = -\frac{D - 2D_{loss}}{1 - D} \qquad (21)$$

D. ZVS condition

To provide ZVS condition for auxiliary inductor switches L3 must be able to charge and discharge their parasitic capacitors.

Therefore:

$$\frac{1}{2}L_3I_{min}^{\ 2} > \frac{1}{2}(C_1 + C_2)(V_1 + V_2)^2 \qquad (22)$$

In addition to the performance ZVS of switches S1 and S2, their dead time must be considered. Because gate- source voltage switches must be applied after reaching their across voltage zero and before changing their body diode flow current. Therefore, to maintain the current direction of switches in the dead time, the auxiliary coil L3 should be enough small. The maximum values of this inductor can be determined.

On the other hand, for the performance of switch S1 in terms of ZVS condition in boost mode, the current value is1 must be negative at t1.

Therefore,

$$(1-n)I_{min} + I_{Lm1} > 0 \qquad (23)$$

Also, for the ZVS performance of switch S2 in boost mode, the value current is2 must be negative at the moment t4.

$$(1-n)I_{max} + I_{Lm2} > 0 \qquad (24)$$

As can be seen, ZVS switches are obtained by n less than unity. This result will be valid for the buck mode.

V. The simulation result of proposed converter

The proposed bidirectional, non isolated converter is simulated and implemented with the specifications:

V_2=80v ،V_1=48v and n=0.3، L_3= 32 uH ، L_m=155 uH and Po=70W ، Ts=10µs

Fig 4. until 7 show simulation waveforms switches for boost mode in direct direction and buck mode in reverse direction. As can be seen, these results match with theoretical analysis and circuit switches operate ZVS in bidirectional power flow. When the switches turn on, currents are negative. Therefore, switches cause soft switching. Because full load condition is the worse mode to provide ZVS condition, therefore, both switches always operate in terms of ZVS.

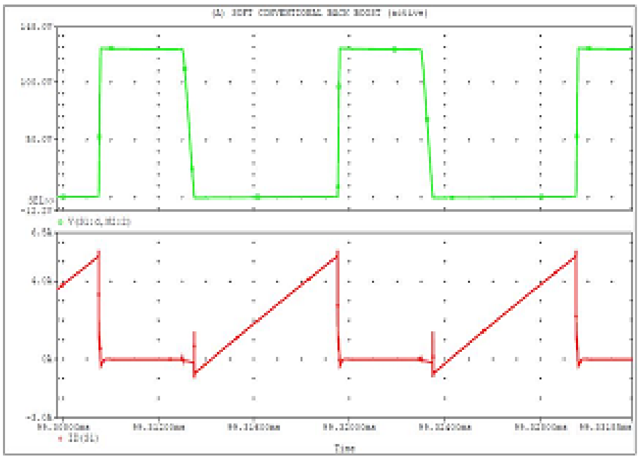

Fig 4: Simulation waveforms of drain voltage - source (upper waveform) and current (lower waveform) S1 switch to boost mode (voltage scale 100 V / div, scale current 5 A / div and the time scale 2.5µs/div)

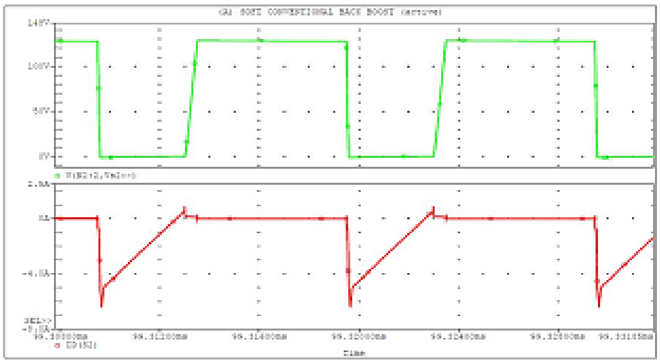

Fig 5: Simulation waveforms of drain voltage - source (upper waveform) and current (lower waveform) S2 switch to boost mode (voltage scale 40 V / div, scale current 2 A / div and the time scale 2.5µs/div)

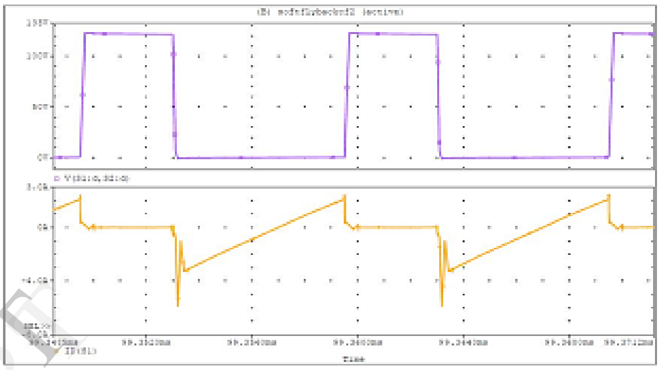

Fig 6: Simulation waveforms of drain voltage - source (upper waveform) and current (lower waveform) S1 switch to buck mode (voltage scale 50 V / div, scale current 5 A / div and the time scale 2.5µs/div)

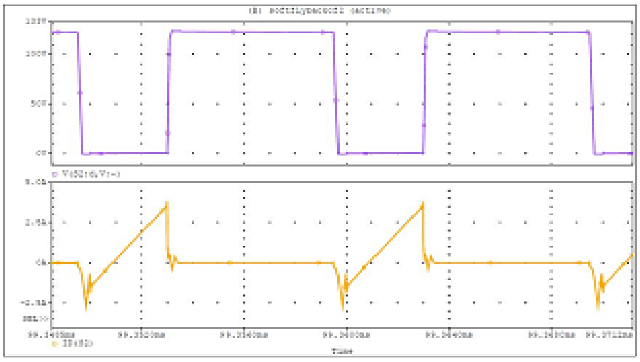

Fig 7: Simulation waveforms of drain voltage - source (upper waveform) and current (lower waveform) S2 switch to buck mode (voltage scale 60 V / div, scale current 5 A / div and the time scale 2.5µs/div)

VI. CONCLUSION

In this paper, a new bidirectional, non-isolated dc-dc converter is proposed. To provide the ZVS conditions, a simple auxiliary circuit is used without an additional switch. The soft switching condition of switches for buck and boost modes is set in both directions. Therefore, the switching losses efficiency of this converter at heavy loads decreases and the recovery reverse body diode switch problems are solved. In addition, converter switches are switching PWM and complement each other. So, it makes to simplify the control circuit. Simulation results and experimental results confirm the accuracy of the theoretical analysis proposed converter.

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