A Family of Magnetic Coupling DC–DC Converters
With Zero-Voltage-Switching Over Wide Input Voltage Range and Load Variation

Guipeng Chen*, Jie Dong*, Yan Deng†, Yong Tao*, Xiangning He*, and Yousheng Wang*

*,†College of Electrical Engineering, Zhejiang University, Hangzhou, China

Abstract

This paper presents a family of soft-switching DC–DC converters with a simple auxiliary circuit consisting of a coupled winding and a pair of auxiliary switch and diode. The auxiliary circuit is activated in a short interval and thus the circulating conduction losses are small. With the auxiliary circuit, zero-voltage-switching (ZVS) and zero-current-switching are achieved for the main and auxiliary switches respectively, over wide input voltage range and load variation. In addition, the reverse-recovery problem of diodes is significantly alleviated because of the leakage inductor. Furthermore, the coupled inductor simultaneously serves as the main and auxiliary inductors, contributing to reduced magnetic component in comparison with the conventional zero-voltage-transition (ZVT) converters. Experimental results based on a 500 W prototype buck circuit validate the advantages and effectiveness of the proposed magnetic coupling ZVS converter.

Key words: DC–DC converters, Magnetic coupling, Zero-current-switching (ZCS), Zero-voltage-switching (ZVS)

I. INTRODUCTION

DC–DC converters have been widely applied in the industrial applications, such as in hybrid electric vehicles and renewable energy systems [1], [2]. To attain high density and efficiency, soft-switching techniques are utilized in high-frequency DC–DC converters. Among these techniques, zero-voltage-switching (ZVS) [3]–[26] including quasi-resonant ZVS, active-clamping ZVS, zero-voltage-transition (ZVT), and magnetic coupling ZVS converters have been intensively studied to eliminate turn-on losses, which is the dominant switching losses in the majority carrier devices, such as MOSFET.

Quasi-resonant converters (QRCs) achieve ZVS operation by employing LC resonance to create a zero-voltage turn-on condition [3]–[5]. Switching losses are effectively reduced, but the main switch suffers from excessively high voltage stress because of resonance. Moreover, the switching frequency varies widely under large load variations, thus making the design of passive components difficult. Pulse-width modulation (PWM) can be achieved in active-clamping ZVS converters with the assistance of an auxiliary switch [6]–[10]. However, the circulating current in the auxiliary circuit is large, thus increasing conduction losses. In addition, the voltage stress of the main switch remains high.

Conduction losses are reduced in ZVT converters since the auxiliary circuit is removed from the main circuit and only activated in a short interval around the turn-on of the main switch [11]–[21]. Furthermore, the voltage stress of the main switch is lower than that of QRCs and active-clamping converters. The work in [11] classified ZVT converters into three classes according to the implementation of the auxiliary voltage source (AVS): Class-A with switched AVS [12]–[15], Class-B with DC AVS [16]–[18], and Class-C with resonant AVS [19]–[21]. The characteristics of the main circuit in all these three converters are similar, whereas the performances of their auxiliary circuits are quite different [22]. The auxiliary circuit of Class-A ZVT converters is simple, but their auxiliary switch is zero-current-switching (ZCS) turned on and ZVS turned off, which causes additional switching losses. Improved ZCS turn-on and turn-off is achieved for the auxiliary switch in the Class-B and Class-C ZVT converters.
But in Class-C ZVT converters with resonant AVS, the design of the resonant tank is complex, and additional reactive energy is introduced. In Class-B ZVT converters, the configuration of the auxiliary circuit remains simple because the DC AVS can be directly obtained from the voltage source of the main circuit. However, the available input voltage range and load variation are limited to ensure that ZVS operation is achieved for the main switch [16], [17].

In references [23]-[26], coupled inductors were employed to achieve ZVS for synchronous DC–DC converters over a wide range of input voltage and load variation. The auxiliary circuit is relatively simple because it is composed of an auxiliary diode and a coupled winding. The magnetic component is reduced since the leakage inductor of the coupled inductor is utilized to function as the auxiliary inductor, which is indispensable in non-isolated QRCs, active-clamping ZVS converters, and ZVT converters. However, the circulating conduction loss in the auxiliary circuit is large because the auxiliary circuit is uncontrolled and activated in a long interval. In [27], circulating conduction losses were reduced with the addition of a switch in series with the auxiliary diode, and the auxiliary circuit was only activated in a short interval. Meanwhile, the controlled auxiliary circuit was also utilized in [28] to achieve ZVS operation for bidirectional DC–DC converters.

The present study proposes a family of magnetic coupling ZVS DC–DC converters. A pair of auxiliary switch and diode is utilized to control the auxiliary circuit, thereby contributing to small circulating conduction losses. Besides, with the assistance of the coupled winding, soft-switching is achieved for both main and auxiliary switches over wide input voltage range and load variation, along with a reduced magnetic component. The paper is organized as follows. Class-B ZVT DC–DC converters are briefly introduced, and a family of magnetic coupling ZVS buck, boost as well as buck-boost converters is derived in Section II. As an example, the operation principle of the proposed ZVS buck converter is shown in Section III. The analysis and comparison results with other ZVS buck converters are provided in Section IV. The experimental results validating the effectiveness of the proposed converter are presented in Section V. Finally, conclusions are drawn in Section VI.

II. TOPOLOGY DERIVATION

In order to gain a clear understanding of the soft-switching principle, Class-B ZVT DC–DC converters which implement DC AVS without additional components, are briefly introduced. ZVS and ZCS can be respectively achieved for main and auxiliary switches with a simple auxiliary circuit. However, ZVS realization is restricted by the input voltage and load condition. To improve the soft-switching characteristic, a family of magnetic coupling DC–DC converters is proposed, which realizes ZVS for the main switch over wide input voltage range as well as load variation. Also, the benefit of ZCS for the auxiliary switch is maintained. Furthermore, no extra auxiliary inductor is needed, contributing to reduced magnetic component.

A. Class-B ZVT DC–DC Converters

The general Class-B ZVT DC–DC converter composed of a basic PWM block and a simple auxiliary circuit which implements DC AVS without additional components, is illustrated in Fig. 1(a) [16]-[18]. The basic PWM block consists of the inductor $L_1$, main diode $D_1$, and main switch $S_1$ with parasitic capacitor $C_{ds}$. To achieve ZVS for the main switch $S_1$, the auxiliary circuit makes use of the inductor $L_a$, switch $S_a$, and diode $D_a$ to discharge and charge $C_{ds}$ and $C_a$, respectively. The operation of the converter is the same as that of the conventional DC–DC converter, except when the auxiliary circuit is active. In the active interval, key waveforms including the drive signals of the main and auxiliary switches, drain-to-source voltage $v_{ds}$, as well as auxiliary inductor current $i_{la}$, are shown in Fig. 1(b) [11], and the equivalent circuit is illustrated in Fig. 1(c). To simplify the operation principle, the current of inductor $L_1$ is assumed to be constant and denoted as $I_L$.

Before $t_0$, the auxiliary switch $S_a$ is in the off state, and the auxiliary inductor current $i_{la}$ is zero. The switch $S_a$ is in the off state, and the inductor current $I_L$ flows through the diode $D_1$.

Stage 1 ($t_0$–$t_1$): At $t_0$, the auxiliary switch $S_a$ is turned on. Then the auxiliary inductor $L_a$ is charged, and $i_{la}$ is linearly increased to $I_L$ at $t_1$. The interval $t_1$–$t_0$ is obtained in (2).

$$i_{la}(t) = \frac{V_{sd} - V_{cd}}{L_a}(t - t_0) = \frac{V_{sd} - V_{cd}}{L_a}(t - t_0)$$

$$t_1 - t_0 = \frac{L_a I_L}{V_{sd} - V_{cd}}$$

Fig. 1. Class-B ZVT DC-DC converter: (a) general topology, (b) key waveforms, and (c) equivalent circuit in the interval $t_0$–$t_1$. 
Stage 2 (t1–t2): In this stage, the equivalent capacitance \( C_{eq} \), which is the sum of parasitic capacitances \( C_{ds} \) and \( C_{dh} \), resonates with the auxiliary inductor \( L_a \), as illustrated in (3). Then the voltage \( V_{da} \) and current \( i_{da} \) can be derived in (4). At \( t_2 \), the voltage \( V_{da} \) decreases to zero, and the resonant process is finished. The interval \( t_2–t_1 \) and \( i_{da}(t) \) are given in (5) and (6), respectively.

\[
\begin{align*}
\frac{dV_{da}}{dt} &= I_L - i_{da}(t), \quad V_{da}(t_1) = V_{ad} \\
L_a \frac{di_{da}}{dt} &= V_{ds}(t) - V_{cd}, \quad i_{da}(t_1) = I_L \\
v_{da}(t) &= V_{cd} + (V_{ad} - V_{cd}) \cos \omega_1(t - t_1) \\
i_{da}(t) &= I_L + \frac{1}{\sqrt{L_a C_{eq}}} (V_{ad}^2 - 2V_{cd} V_{ad}) \\
\end{align*}
\]

(3)

(4)

(5)

(6)

where \( \omega_1 = \frac{1}{\sqrt{L_a C_{eq}}} \).

Stage 3 (t2–t3): After the complement of the resonant process, the current \( i_{da} \) flows through the anti-parallel diode of switch \( S_1 \). Therefore, \( i_{da} \) is linearly decreased, and the switch \( S_1 \) should be turned on to achieve ZVS operation before \( i_{da} \) decays to \( I_L \) at \( t_3 \). Thus, the phase angle \( \varphi \) between the two drive signals in Fig. 1(b) should satisfy (9).

\[
i_{da}(t) = \frac{V_{cd}}{L_a} (t - t_2) \\
t_3 - t_2 = \sqrt{\frac{L_a C_{eq} (V_{cd}^2 - 2V_{cd} V_{ad})}{V_{cd}}} \\
(t_2 - t_0) < \varphi / 2\pi < (t_3 - t_2)
\]

(7)

(8)

(9)

where \( T_s \) is the switching period.

Stage 4 (t3–t4): The operation in this stage is similar to that of stage 3, except that the current \( i_{da} \) flows from the anti-parallel diode to the MOSFET channel of switch \( S_1 \). \( i_{da} \) continues to decrease and drops to zero at \( t_4 \).

In order to achieve ZVS for the main switch \( S_1 \), the parasitic capacitance \( C_{ds} \) should be completely discharges in stage 2; thus, the voltage limitation is derived in (10) from (4) [16]. Meanwhile, the limit of maximum inductor current \( I_{L,max} \) is obtained in (11) from (2), (5), (8), and (9) to realize ZVS with a fixed \( \varphi \) over the whole load range. Therefore, the practical applications of Class-B ZVT converters in Fig. 1(a) are restricted.

\[
V_{cd} < 0.5V_{ad} \\
I_{L,max} < \frac{V_{ad} - V_{cd}}{V_{cd}} \sqrt{\frac{C_{eq}(V_{cd}^2 - 2V_{cd} V_{ad})}{L_a}}
\]

(10)

(11)

B. Proposed Magnetic Coupling ZVS DC–DC Converters

In order to achieve ZVS for main switches over wide input voltage range and load variation, a family of magnetic coupling ZVS DC-DC converters is proposed. The general magnetic coupling ZVS DC–DC converter is illustrated in Fig. 2(a). Compared with the Class-B ZVT converter in Fig. 1(a), a coupled inductor \( T_1 \) with turns ratio \( N_p:N_s=1:n \) is employed to implement the main inductor \( L_1 \) with the magnetizing inductor \( L_a \) and the auxiliary inductor \( L_a \), with the leakage inductor \( L_r \). Thus the magnetic component number is reduced. The secondary winding of the coupled inductor \( T_1 \), the auxiliary switch \( S_m \), and the diode \( D_{b} \) are connected in series between ports \( c \) and \( d \). The equivalent circuit in the interval when \( S_m \) and \( D_{d} \) conduct is shown in Fig. 2(b). Compared with Fig. 1(c), an additional voltage source \(-V_{cd}/n\) is added since the secondary voltage of the coupled inductor \( T_1 \) is clamped to \(-V_{cd} \). Hence, the limitation in (10) and (11) is modified in (12) and (13), respectively. According to (10), for the Class-B ZVT converter in Fig. 1(a), the ZVS operation of the main switch \( S_1 \) can be achieved only when \( V_{cd} \) is smaller than 0.5 \( V_{ad} \), which is greatly alleviated in the proposed converter as illustrated in (12). Moreover, the available ZVS load range is also enlarged with the comparison of (13) and (11).

\[
(1 - \frac{1}{n})V_{cd} < 0.5V_{ad}
\]

(12)

\[
I_{L,max} < \frac{V_{ad} - (1 - \frac{1}{n})V_{cd}}{(1 - \frac{1}{n})V_{cd}} \sqrt{\frac{C_{eq}(V_{cd}^2 - 2(1 - \frac{1}{n})V_{cd} V_{ad})}{L_r}}
\]

(13)
On the basis of the general magnetic coupling DC–DC converter in Fig. 2(a), the corresponding ZVS buck, boost, and buck-boost converters with different values for \( V_{ad} \) and \( V_{cd} \) in the Table I, are derived in Fig. 3. Besides, the magnetic coupling concept can be employed to achieve ZVS operation for various converters which consists of the basic PWM block.

### III. OPERATION PRINCIPLE

As the operation principles of the proposed magnetic coupling ZVS buck, boost, and buck-boost converters are similar, the ZVS buck converter is used as an example in this section, which is re-shown in Fig. 4(a). Key operating waveforms of the converter are shown in Fig. 4(b). The auxiliary switch \( S_a \) is turned on in advance to decrease the leakage inductor current \( i_{Lr} \) and create ZVS condition for the main switch \( S_1 \). In a switching period, the operation comprises seven stages, and the corresponding equivalent circuits of which are illustrated in Fig. 5.

Several assumptions are made to simplify the operation principle.

—\( L_m \) is large; hence, the magnetizing current \( I_L \) is assumed to be constant.

—The leakage inductance \( L_r \) is much lower than the magnetizing inductance \( L_m \).

—The parasitic capacitances \( C_d \) and \( C_d' \) are constant during the switching process, and their sum is denoted as \( C_{eq} \).

—All components are ideal, except for the parasitic capacitances of the main switch and diode.

From Fig. 4(a), the relationship among the leakage inductor current \( i_{Lr} \), auxiliary current \( i_n \), magnetizing current \( I_L \), and output current \( i_o \) is derived in (14) and (15).

Prior to \( t_0 \), \( S_1 \) and \( S_a \) are in the on state. The leakage inductor current \( i_{Lr} \) is increased, whereas the auxiliary current \( i_n \) is decreased.

Stage 1 \((t_0-t_1)\): At \( t_0 \), \( i_{Lr} \) rises to the magnetizing current \( I_L \), and \( i_n \) decays to zero; thus, \( D_a \) is reverse biased. In this stage, the energy is transferred from the input to the leakage inductor \( L_r \), magnetizing inductor \( L_m \), and output. Since the auxiliary current \( i_n \) is zero, the auxiliary switch \( S_a \) is ZCS turned off in this stage.

Stage 2 \((t_1-t_2)\): \( S_1 \) is turned off at \( t_1 \). The parasitic capacitors \( C_d \) and \( C_d' \) are respectively charged and discharged by \( I_L \). Therefore, the drain-to-source voltage \( v_{ds} \) increases linearly, as shown in (19).
\[
v_a(t) = \frac{I_L}{C_{eq}}(t - t_d) \quad (19)
\]
\[
v_{in}(t) = \frac{L_m}{L_r + L_m} (V_i - V_o - v_a(t)) \quad (20)
\]
\[
v_{L}(t) = -\frac{L_r}{L_r + L_m} (V_i - V_o - v_a(t)) \quad (21)
\]

Stage 3 \((t_3 - t_4)\): At \(t_3\), \(v_{di}\) increases to \(V_i\); hence, the main switch \(D_1\) is forward biased. During this stage, the energy is transferred from the leakage inductor \(L_r\) and magnetizing inductor \(L_m\) to the output.

\[
i_o(t) = i_{L}(t) = I_L \quad (22)
\]
\[
v_{in}(t) = -\frac{L_m}{L_r + L_m} V_o \quad (23)
\]
\[
v_{L}(t) = -\frac{L_r}{L_r + L_m} V_o \quad (24)
\]

Stage 4 \((t_4 - t_5)\): At \(t_4\), the auxiliary switch \(S_a\) is turned on, and the magnetizing voltage \(v_{in}\) is clamped to \((V_i - V_o)/n\). In this stage, the leakage inductor current \(i_{Lr}\) and the diode current \(i_o\) decrease linearly, as shown in (27). Based on (14), the auxiliary current \(i_a\) increases.

\[
v_{in}(t) = \frac{V_i - V_o}{n} \quad (25)
\]
\[
v_{L}(t) = -\frac{V_i + (n-1)V_o}{n} \quad (26)
\]
\[
i_o(t) = i_{L}(t) = I_L - \frac{V_i + (n-1)V_o}{nL_r}(t - t_5) \quad (27)
\]

Stage 5 \((t_5 - t_6)\): The leakage inductor current \(i_{Lr}\) and diode current \(i_o\) decay to zero at \(t_5\). Owing to the leakage inductor \(L_r\), the reverse-recovery problem of the main diode is greatly alleviated. Then, \(L_r\) resonates with \(C_{eq}\) as shown in (28).

\[
\begin{aligned}
&L_r \frac{di_{Lr}(t)}{dt} = \frac{(n-1)(V_i - V_o)}{n} - v_{in}(t) \\
&C_{eq} \frac{dv_{in}(t)}{dt} = i_o(t)
\end{aligned} \quad (28)
\]

Stage 6 \((t_6 - t_7)\): The voltage \(v_{di}\) decreases to zero at \(t_6\), and the current \(i_{Lr}\) flows through the body diode of \(S_1\); hence, \(S_1\) is ZVS turned on. The leakage inductor voltage \(v_{Li}\), shown in (29), must be larger than zero to increase the leakage inductor current \(i_{Lr}\) to the magnetizing current \(I_r\). Therefore, the turns ratio \(n\) should be larger than 1. And from (14), the auxiliary current \(i_a\) decreases.

\[
v_{Lr} = \frac{(n-1)(V_i - V_o)}{n} \quad (29)
\]
\[
n > 1 \quad (30)
\]
\[
i_a(t) = i_{Lr}(t) = i_a(t_5) + \frac{(n-1)(V_i - V_o)}{nL_r}(t - t_5) \quad (31)
\]

Stage 7 \((t_7 - t_8)\): At \(t_7\), \(i_{Lr}\) rises to zero. This stage is similar with the stage 6 except that the leakage inductor current \(i_{Lr}\) is positive. The switching period ends at \(t_7\) when the auxiliary current \(i_a\) drops to zero.

IV. ANALYSIS AND COMPARISON

A. Voltage Gain

Owing to the flux balance, the average voltage across the leakage inductor \(L_r\) and magnetizing inductor \(L_m\) in a switching period is zero. Therefore, the output voltage \(V_o\) is equal to the average diode voltage \(V_{di}\). From Fig. 4(b), the voltage gain of the proposed converter is derived in (32), which is similar with that of the conventional buck converter.

\[
M = \frac{V_o}{V_i} = \frac{V_{di}}{V_i} \approx D \quad (32)
\]

B. Average Magnetizing Current \(I_L\)

From the operation principle in section III, the leakage inductor current \(i_{Lr}\) in different stage is summarized in (33). Then the relationship between the average leakage inductor current \(I_{Lr}\) and magnetizing current \(I_L\) can be derived. Combing with (34) derived from (14) and (15), \(I_L\) can be calculated in terms of average output current \(I_o\). The relationship between \(I_L\) and \(I_o\) with \(L_r=6 \mu H\) and \(n=1.5\) as a function of voltage gain \(M\) is illustrated in Fig. 6. The average magnetizing current \(I_L\) is a little larger than the average output current \(I_o\), but the difference is small, particularly at a low voltage gain \(M\).
For the Class-B ZVT buck converter with the neglect of the leakage inductance, the RMS current of the auxiliary switch is derived in (37). In the interval $[t_s, t_1]$, the auxiliary diode current $i_{a,RMS}$ is reverse biased and the turn-off voltage is given in (39). In the interval $[t_1, t_2]$, the turn-off voltage of the main switch is derived in (42). In the interval $[t_2, t_3]$, the available ZVS load range for the ZVT buck converter in Fig. 1(a) is narrow and deteriorated with the increase of auxiliary inductance $L_m$. Likewise, the auxiliary switch $S_a$ is turned off in the stage 3, and the voltage stress is derived in (37).

$$V_{s1} = V_{dl} = V_i$$
$$V_{d2} = nV_{in} - (V_i - V_o) = (-\frac{nL}{L_m + L_a} - 1)(V_i - V_o)$$
$$\approx (n-1)(V_o - V_i)$$
$$V_{s3} = V_o - nV_{in} = V_o - n\frac{L}{L_m + L_a}(-V_o)$$
$$\approx V_o - (n-1)V_o$$

**D. Current Stress**

According to Fig. 4(b), in the intervals $[t_s, t_1]$ and $[t_0, t_1]$, the main switch $S_1$ conducts and the current $i_b$ is equal to the leakage inductor current $i_{Lr}$. Therefore, the RMS current of the main switch is derived in (38). In the interval $[t_1, t_2]$, the leakage inductor current $i_{Lr}$ flows through the main diode $D_1$ and hence the average current $I_{Lr}$ is given in (39). Similarly, the RMS current of the auxiliary switch $I_{a,RMS}$ and the average current of the auxiliary diode $I_{a,ave}$ can be obtained, as illustrated in (40) and (41).

$$I_{a,RMS} = \frac{1}{T_s} \int_{t_0}^{t_1} |i_{a}(t)| dt$$
$$I_{a,ave} = \frac{1}{T_s} \int_{t_0}^{t_1} |i_{a}(t)| dt$$

**E. Soft-switching Analysis**

For the proposed buck converter, in order to achieve ZVS for the main switch $S_1$, (42) must be satisfied from (12) and Table I. Combining with $n=1$ in (30), the available value of turns ratio $n$ at different voltage gain $M$ is depicted in Fig. 7. With $1 < n \leq 2$, the main switch can theoretically realize ZVS over the whole input voltage range.

$$n(V_i - 2V_o) < 2V_i - V_o$$

In practical, the maximum current $I_{L,\text{max}}$ is restricted in (11) and (13) to realize ZVS operation over the whole load range with a fixed phase angle $\phi$ for the Class-B ZVT buck converter in Fig. 1(a) and the proposed converter, respectively. Fig. 8 depicts the relationship between $I_{L,\text{max}}$ and the voltage gain $M$ with different turns ratios $n$. The available ZVS load range for the ZVT buck converter in Fig. 1(a) is narrow and deteriorated with the increase of auxiliary inductance $L_m$, which is significantly improved in the proposed ZVS buck converter. Moreover, with the decrease of turns ratio $n$, the available load range is enlarged. Therefore, the proposed buck converter can achieve ZVS operation for the main switch over both wide input voltage range and large load variation.

The decreasing ratio of the diode current $i_D$ and auxiliary current $i_a$ is limited by the leakage inductor $L_r$, as illustrated in Fig. 4(b). Therefore, the reverse-recovery problem of diodes $D_1$ and $D_a$ is greatly alleviated. Moreover, the auxiliary switch $S_a$ is ZCS turned off in the stage 1 since the auxiliary current $i_a$ already decays to zero at $t_1$. 

![Fig. 7. Available turns ratio $n$ to achieve ZVS as a function of $M$.](image)
A Family of Magnetic …

Fig. 8. Maximum magnetizing current $I_{L,max}$ to achieve ZVS over whole load range with a fixed $\varphi$ for different converters.

F. Comparison

The comparison results between the proposed and other ZVS buck converters are shown in Table III. The QRC buck converter in [4] is considerably simple and does not require an auxiliary switch, but the voltage stress of the main switch is excessively high. In [6], the voltage stress was reduced in the active-clamping buck converter, but it remained higher than that of the input voltage $V_i$. Moreover, the circulating current is large; thus, conduction losses are deteriorated. The ZVT buck converter in [17] can effectively reduce conduction losses because the auxiliary circuit is removed from the main circuit and only activated in a short interval. However, ZVS operation is lost in the application with $V_o<0.5V_i$. Nevertheless, all aforementioned converters need an auxiliary inductor. In the proposed buck converter, the ZVS operation can be achieved over wide input voltage range and load variation with the benefits of low voltage stress on the main switch, small circulating current and reduced magnetic component.

V. EXPERIMENTAL VALIDATION

In order to verify the effectiveness of the proposed topology, a 500 W prototype circuit of the magnetic coupling ZVS buck converter with topology parameters in Table II is built. The soft-switching characteristics of the converter are assessed at different input voltages 200 and 300 V under 10% and 100% load variations with open-loop experimental results, respectively.

The steady-state experiment waveforms at 100% load with input voltages $V_i=200$ V and $V_i=300$ V are shown in Fig. 9, Fig. 10 and Fig. 11, which are in well coincidence with the theoretical analysis. In Fig. 9, the leakage inductor current $i_{Lr}$ decreases after the turn-on of auxiliary switch $S_a$ to achieve ZVS turn-on for the main switch $S_1$, and then rapidly reset to the magnetizing current $I_L$. The auxiliary circuit is only activated in a short interval, and thus the additional conduction loss is reduced. It is noteworthy that the auxiliary current $i_a$ increases slightly and that the leakage inductor current $i_{Lr}$ decreases after the turn-off of the main switch $S_1$ due to the influence of the RCD, which is in parallel with the auxiliary switch $S_a$. From Fig. 10, ZVS is achieved for the main switch $S_1$ at both input voltages. Moreover, the reverse-recovery problem of the main diode $D_1$ is greatly alleviated. In addition, as shown in Fig. 11, the ZCS of the auxiliary switch $S_a$ is obtained, which is also independent of the input voltage. Therefore, the proposed ZVS buck converter achieves desirable soft-switching characteristics over a wide input voltage range.

<table>
<thead>
<tr>
<th>TABLE II</th>
</tr>
</thead>
<tbody>
<tr>
<td>PARAMETER SPECIFICATION</td>
</tr>
<tr>
<td>Parameter</td>
</tr>
<tr>
<td>Input voltage</td>
</tr>
<tr>
<td>Output voltage</td>
</tr>
<tr>
<td>Rated output power</td>
</tr>
<tr>
<td>Switching frequency</td>
</tr>
<tr>
<td>Turns ratio</td>
</tr>
<tr>
<td>Magnetizing inductance</td>
</tr>
<tr>
<td>Leakage inductance</td>
</tr>
<tr>
<td>Main switch</td>
</tr>
<tr>
<td>Main diode</td>
</tr>
<tr>
<td>Auxiliary switch</td>
</tr>
<tr>
<td>Auxiliary diode</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>TABLE III</th>
</tr>
</thead>
<tbody>
<tr>
<td>COMPARISON OF THE PROPOSED ZVS BUCK CONVERTER AND OTHER ZVS BUCK CONVERTERS</td>
</tr>
<tr>
<td>Voltage Stress</td>
</tr>
<tr>
<td>---</td>
</tr>
<tr>
<td>QRC [4]</td>
</tr>
<tr>
<td>Active-clamping [6]</td>
</tr>
<tr>
<td>ZVT [17]</td>
</tr>
<tr>
<td>Proposed</td>
</tr>
</tbody>
</table>
The experiment waveforms of the main switch $S_1$ and auxiliary switch $S_a$ at 10% load condition with input voltages $V_i=200$ V and $V_i=300$ V are illustrated in Fig. 12 and Fig. 13, respectively. ZVS and ZCS operation are also achieved. Note that the resonance caused by the non-ideality of semiconductor devices and parasitic inductances during the turn-on process of $S_1$ has slight influence on the ZVS operation, e.g. the drain-to-source voltage $v_{ds}$ increases slightly as shown in Fig. 12(b). Nevertheless, $v_{ds}$ finally decays to zero before the turn-on of $S_1$. Therefore, improved soft-switching characteristics of the proposed converter remain regardless of the load variation.

The measured efficiency of the proposed magnetic coupling buck converter at $V_i=200$ V and $V_i=300$ V is shown in Fig. 14(a), which achieves a maximum value of 97.2%. Thanks to the desirable soft-switching characteristic and the small auxiliary circuit conduction loss, the efficiency of the proposed converter is much improved over whole load range at both input voltages, in comparison with the conventional buck converter, as illustrated in Fig. 14(b)-(c). Compared with the ZVT buck converter in Fig. 1(a), the proposed converter achieves a slightly lower efficiency under heavy load with input voltage $V_i=200$ V in Fig. 14(b). Nevertheless, the efficiency of the ZVT converter under light load is harshly decreased, resulting from the loss of ZVS with a fixed phase angle $\phi$. Moreover, at $V_i=300$ V in Fig. 14(c), ZVS operation is still obtained in the proposed ZVS buck converter but lost in the ZVT buck converter. Therefore, the
A family of magnetic coupling ZVS buck, boost and buck-boost converters is proposed. Thanks to the additional

VI. CONCLUSIONS

Fig. 12. Experiment waveforms of the drive signal \( v_{gs1} \), drain-to-source voltage \( v_{ds1} \), and current \( i_s \) at 10% load with different input voltages: (a) \( V_i = 200 \) V and (b) \( V_i = 300 \) V.

Fig. 13. Experiment waveforms of the drive signal \( v_{gs2} \), drain-to-source voltage \( v_{ds2} \), and current \( i_s \) at 10% load with different input voltages: (a) \( V_i = 200 \) V and (b) \( V_i = 300 \) V.

Fig. 14. Measured efficiency: (a) proposed converter at \( V_i = 200 \) V and \( V_i = 300 \) V, (b) comparison at \( V_i = 200 \) V, and (c) comparison at \( V_i = 300 \) V.

Fig. 15. Photograph of the prototype circuit.

efficiency of the proposed converter is higher than that of the ZVT converter at \( V_i = 300 \) V. A photograph of the prototype circuit is shown in Fig. 15.
auxiliary voltage source provided by the coupled inductor. ZVS operation is achieved for the main switch over wide input voltage range and load variation with a reduced magnetic component. Besides, the auxiliary circuit obtains the advantages of ZCS operation and small circulating current. The proposed ZVS buck converter is taken as an example to clearly introduce the operation principle and analysis. Experiment waveforms and measured efficiency based on a 500 W prototype circuit at input voltages of 200 V and 300 V under 10% and 100% load conditions are given to validate converter performance.

ACKNOWLEDGMENT

This work was sponsored by the National Natural Science Foundation of China (51377144).

REFERENCES


Guipeng Chen received his B.E.E degree from the Department of Electrical Engineering, Zhejiang University, Hangzhou, China, in 2011. He is currently working toward his Ph. D degree at the College of Electrical Engineering, Zhejiang University, Hangzhou, China. His current research interests include soft-switching DC–DC converters, bidirectional DC–DC converters, and multi-port DC–DC converters.

Jie Dong received her B.E.E degree from the School of Electronic Information and Electrical Engineering, Shanghai Jiaotong University, Shanghai, China, in 2014. She is currently working toward her MA. Sc degree at the College of Electrical Engineering, Zhejiang University, Hangzhou, China.

Yan Deng received his B.E.E degree from the Department of Electrical Engineering, Zhejiang University, Hangzhou, China, in 1994, and his Ph.D. degree in Power Electronics and Electric Drives from the College of Electrical Engineering, Zhejiang University, in 2000. Since 2000, he has been a faculty member at Zhejiang University, where he teaches and conducts research on power electronics. He is currently an Associate Professor. His research interests are topologies and control for switch-mode power conversion.

Yong Tao received his B.S. degree in Electrical Engineering from Nanjing University of Science and Technology, Nanjing, China, in 2011. He is currently working toward his Ph.D. degree at the College of Electrical Engineering, Zhejiang University, Hangzhou, China. His research interests include microgrid control, distributed generation, and energy management.

Xiangning He received his B.Sc. and M.Sc. degrees from Nanjing University of Aeronautical and Astronautical, Nanjing, China, in 1982 and 1985, respectively, and his Ph.D. degree from Zhejiang University, Hangzhou, China, in 1989. From 1989 to 1991, he was a Lecturer at Zhejiang University. In 1991, he obtained a fellowship from the Royal Society of the U.K. and conducted research at Heriot-Watt University, Edinburgh, U.K., as a Post-Doctoral Research Fellow for two years. In 1994, he joined Zhejiang University as an Associate Professor. Since 1996, he has been a Full Professor at Zhejiang University. He is currently the Vice Dean of the College of Electrical Engineering, Zhejiang University. His research interests are power electronics and their industrial applications. He has published more than 280 papers and holds 22 patents.

Yousheng Wang was born in 1928. He received his B.S. degree from the Department of Electrical Engineering, Zhejiang University, Hangzhou, China. In 1950, he joined the Department of Electrical Engineering of Zhejiang University as a faculty member. He became a Professor in 1978. His research interests include power electronics and high-frequency induction heating. Mr. Wang received several national awards for his outstanding work in large electric generators and static induction heating equipment. He is a member of the China Academy of Engineering.