A New Soft Switched Full Bridge Converter With Voltage-Doubler-Type Rectifier for High Voltage Applications

V. Delbin Jelaja and M. Rajaram

Abstract—A new soft switched full bridge converter with voltage doubler type rectifier is proposed to reduce the circulating loss in primary and the voltage stress in secondary. The conventional converter is having the drawbacks such as circulating loss in the primary, voltage spike across the rectifier diode. Also the conventional system is using large output inductor in the secondary side, because of this core loss occurs. In this paper we introduce a new technique to overcome the above said drawbacks. Without an auxiliary circuit, zero – voltage switching (for leading - leg switches) and zero – current switching (for lagging – leg switches) are achieved in the entire operating range, which reduces the circulating loss in the primary side. To implement the converter without an additional inductor, the leakage inductance of the transformer is utilized as the resonant inductor. The resonance between the leakage inductance of the transformer and the rectifier capacitor can reduce the current stresses of the rectifier diodes and the conduction losses. The clamp diode is used to clamp the voltage spike across the rectifier diode which releases the voltage stress of the rectifier diode. Due to its many advantages this converter is attractive for high voltage and high power applications. The analysis and design considerations of the proposed converter are presented. The simulation results confirm efficiency above 98% which is 3% to 5% more than the conventional converters.

Index Terms—Phase shift full bridge (PSFB) converter, zero – voltage and zero – current switching (ZVZCS), voltage - doubler – type rectifier

I. INTRODUCTION

The zero - voltage and zero – current switching converters are very attractive for high voltage and high power applications. The phase shifted ZVS PWM converter is often used in many applications because its topology permits all switching devices to operate under zero–voltage switching by using circuit parasitics such as power transformer leakage inductance and the rectifier capacitors. However, because of Phase Shifted PWM Control, the converter has a disadvantage that circulating current flows through the power transformer and switching devices during freewheeling intervals. The circulating current is the sum of reflected output current and transformer primary magnetizing current. Due to the circulating current, root mean square (RMS) current stresses of the transformer and switching devices are still high compared with those of the conventional hard- switching PWM FB converter. The zero-voltage and zero–current switching (ZVZCS) technique is proposed based on PSFB converter to reduce the circulating loss. The key point of this technique is to use the leakage inductance of the transformer as resonant inductor, with the help of that we can achieve Zero–voltage, zero–current switching in the entire operating range. The conduction loss is high [1] for the diode in series with primary switch due to the large primary current, though the circulating loss can be eliminated. Inserting an auxiliary inductor with three coupling windings in primary side of the transformer [2], primary current is reset by input voltage source through the auxiliary coupling windings in the transformer, but the conduction loss in auxiliary circuits is large because of high input current, and the voltage ringings across the rectifier diodes are still high. By an active switch in secondary side [3], [4], the primary current in transformer can be reduced. But the control is complex because the auxiliary switch should be driven corresponding to the driving logic of the primary switches. In [5] the primary current is reset by utilizing dc blocking capacitor and adding a saturable inductor in the primary. In spite of the simple additional circuit, this converter is not suitable for high power

Fig. 1. Previous Secondary sides current reset circuits.

Switching in the entire operating range. The conduction loss is high [1] for the diode in series with primary switch due to the large primary current, though the circulating loss can be eliminated. Inserting an auxiliary inductor with three coupling windings in primary side of the transformer [2], primary current is reset by input voltage source through the auxiliary coupling windings in the transformer, but the conduction loss in auxiliary circuits is large because of high input current, and the voltage ringings across the rectifier diodes are still high. By an active switch in secondary side [3], [4], the primary current in transformer can be reduced. But the control is complex because the auxiliary switch should be driven corresponding to the driving logic of the primary switches. In [5] the primary current is reset by utilizing dc blocking capacitor and adding a saturable inductor in the primary. In spite of the simple additional circuit, this converter is not suitable for high power
applications due to the core loss of the saturable inductor. Fig 1 represents the previous secondary side current reset circuits. Fig 1(a) shows the snubber circuit. Here the peak voltage across the rectifier can reach twice of the steady state value. Fig 1(b) shows the H-bridge snubber. Here the lossy components are more, voltage stress across the rectifier reaches twice of the steady state value and large circulating current flows through the primary which reduces the overall system efficiency. Fig 1(c) represents the transformer auxiliary winding, which increases the complexity of the transformer and it requires additional RC/RCD snubber. Fig 1(d) shows the coupled output inductor, here the voltage across the rectifier cannot be reduced properly and the switching noise can also be more. Fig 1(e) shows the PWM converter, it implies additional rectifier conduction losses and secondary parasitic ringing. Fig 1(f) shows the energy recovery snubber, the drawback of the circuit is the generation of additional losses and switching noise can be produced in the secondary side rectifiers.

This paper proposes a PSFB converter without circulating loss in primary side. Furthermore the voltage stress and current stress of the rectifier diode is reduced to a much low value compared with previous ZVZCS converter. The circuit diagram is shown in Fig 2. Here D1, D2 are main rectifier diodes. Dc is the clamping diode which is used to clamp the voltage spike across the rectifier diode. To implement the converter without an additional inductor, the leakage inductance of the transformer is utilized as the resonant inductor. In addition the proposed converter employs the voltage – doubler rectifier. In addition the resonance between the leakage inductor of the transformer and the rectifier capacitors can reduce the current stress of the rectifier diodes and conduction losses. Normally the voltage doubler is used in an ac–dc rectifier with the step up operation, and it is rarely adopted in dc–dc converters because of the voltage tolerance of the capacitor. However since the voltage doubler can behave as an inductance – capacitance filter, which eliminates the large output inductor. Due to the elimination of large output inductor, it features simple structure, lower cost, lighter weight and small mass. Further more, the proposed converter has wide zero-voltage switching ranges of lagging leg switches with low current stresses of the primary power switches by using the magnetizing current.

II. OPERATIONAL PRINCIPLE

The circuit diagram and key waveforms of the proposed converter are shown in figs 2 and 3 respectively. The operation of the proposed converter can be divided into twelve modes. One switching cycle of the proposed circuit is divided into two half cycles. 1. \( t_0 \sim t_2 \) and 2. \( t_2 \sim t_4 \). Since the operation principles of the two half cycles are symmetric, only the first half cycle is explained. One half cycle can be divided into six modes and its equivalent circuits are shown in fig.5.

A. Stage 1 \( t_0 \sim t_2 \)

At \( t_0 \), \( M_2 \) turns on, the input \( V_{in} \) is across the transformer primary winding. The primary current \( I_p \) increases linearly until it reaches the reflected output inductor current as follows.

\[
I_p(t) = \frac{1}{n} \left( V_m - V_{o1}(t_o) \right) \sin \omega_r(t - t_o) + I_m(t - t_o)
\]

where \( \omega_r = \frac{n}{\sqrt{L_m C_r}} \), \( z_0 = \frac{1}{n} \sqrt{\frac{L_m}{C_r}} \).

The magnetizing current, which also rises with the resonance between the magnetizing inductor and the rectifier capacitors, is obtained as follows.

\[
I_p(t) = I_m(t_0) \cos \omega_m(t - t_0) \frac{n V_{o1}(t_o)}{Z_m} \sin \omega_m(t - t_0)
\]

where \( \omega_m = \frac{n}{\sqrt{L_m C_r}} \), \( z_m = n \sqrt{\frac{L_m}{C_r}} \).

On the other hand, since the resonant frequency \( (f_m = \frac{\omega_m}{2\pi}) \) is much slower than the switching frequency, \( I_m \) can linearly be approximated as follows:

\[
I_m(t) = I_m(t_0) + \frac{n V_{o1}(t - t_o)}{L_m}(t - t_0)
\]
this interval is shown in Fig.5 (b). In order to calculate the magnitude of the resonant voltage and current, the equivalent circuit can be simplified into a resonant circuit as shown in Fig.5. Where $V_c (t)$ is the voltage across the rectifier capacitor. The voltage $V_{rec}(t) = V_c(t)$. This interval ends when the voltage increases to $(V_0 + V_{c1}) = \frac{3V_0}{2}$

$$I_{LK} = I_0 + \frac{V_{in}}{Z_r} - \frac{V_{c1}}{Z_r} \sin \omega_r (t - t_i)$$  \hspace{1cm} (4)

$V_c(t) = \frac{V_{in}}{n} (1 - \cos \omega_r (t - t_i)) + V_{c1} \cos \omega_r (t - t_i)$  \hspace{1cm} (5)

$$I_p(t) = \frac{I_{LK}(t)}{n}$$  \hspace{1cm} (6)

where $z_r = \sqrt{I_{LK}/C}$.

$$c = c_1 + c_2, \quad \omega_r = \frac{1}{\sqrt{L_{lk}C}}, \quad V_{c1} = \frac{V_0}{2}$$

**C. Stage 3 (t_2 ~ t_3)**

When the voltage $V_{rec}$ reaches to $3V_0/2$, the clamping diode $D_C$ conducts. Thus, the voltage $V_{rec}$ is clamped to $3V_0/2$. The equivalent circuit of this stage is shown in Fig.5(c). Because of reflected voltage $V_{in}/n$ from input is lower than $3V_0/2$ the current in $L_{lk}$ begins decreasing at $t_2$. The voltage across $L_{lk}$ is $[3V_0/2 - V_{in}/n]$. This stage ends when the current $I_p$ decreases to

$$\frac{di_p}{dt} = \left(\frac{3V_0}{2n} - \frac{V_{in}}{n^2}\right)/L_{LK}$$  \hspace{1cm} (7)

**E. Stage 5(t_4 ~ t_5)**

At $t_5$, $M_1$ turns off. The intrinsic capacitors of $M_1$ and $M_2$ are charged by reflected load current. When voltage across $M_1$ reaches to $V_{in}$ the body diode of $M_2$ conducts. Thus, $M_2$ can achieve ZVS turning on. The voltage transition mode of lagging leg switches is similar to conventional PSFB converter. Furthermore, the voltage increases very fast because of large primary current reflected from load current. When the body diode of $M_2$ conducts, the voltage across primary winding is clamped to zero. Thus, the current in primary is begins decreasing, and the decreasing slope is described as follows. This interval ends when current $i_p$ is reset to zero.

$$\frac{dV_p}{dt} = \frac{V_C}{n} \frac{1}{L_{lk}}$$  \hspace{1cm} (8)

**Stage 4(t_3 ~ t_4)**

When $D_C$ cut off $V_{rec}$ keeps at $V_{in}/n$. The voltage spike across the rectifier diode is reduced.
Stage 6 (t5 \rightarrow t6)

After t5 the converter is in freewheeling stage. Because the primary circulating current is reset to zero, the circulating loss in primary side is eliminated without considering the transformer magnetizing current. Due to the clamping diode the voltage spike across the rectifier diode is eliminated. After t5 the other half switching cycle begins. Due to the symmetry operation characteristics, the principles and the equivalent circuits of the other half switching cycle are same to that mentioned above.

III. ANALYSIS OF THE PROPOSED CONVERTER

A. ZVS and Circulating Energy

In the conventional PSFB converter the ZVS operation of the leading leg switches can easily be achieved due to the large output filter inductor. But due to large output inductance core loss occurs due to the saturation of core. This is eliminated with the help of the proposed converter: To implement the converter without additional inductance, the leakage inductance of the transformer is utilized as the resonant inductor. Due to its many advantages, including high efficiency, minimum number of devices, and low cost, this converter is attractive for high-voltage and high-power applications.

B. Reduction of Current stress and Conduction loss by using Resonance

A disadvantage of the proposed converter is that the current stresses of the rectifier diodes are rather large. This is because the proposed converter employs the voltage doubler, which causes the voltage - current doubling effects. However, the resonance between the leakage inductor and the rectifier capacitors reduces not only the current stresses of rectifier diodes but also conduction losses.

\[ V_c(t) = \frac{V_{in}}{n} \left[ 1 - \cos(\omega_r (t-t_1)) + V_{cl} \cos(\omega_r (t-t_1)) \right] \]  
\[ I_p(t) = \frac{I_{LK}(t)}{n} \]

where \( \omega_r = \frac{1}{\sqrt{L_{in} C_r}} \), \( z_r = \sqrt{L_{lk} / c_r} \), \( V_c = \frac{V_0}{2} \)

C. Voltage stress of the Rectifier Diode

The auxiliary clamp diode is used to clamp the voltage spike across the rectifier diode, which releases the voltage stress of the rectifier diode.

D. Filter Requirements

The filter requirements of the proposed converter are different from that of conventional PSFB Converter because there is an auxiliary voltage source at the input of the filter inductors during the freewheeling interval. The current ripple (peak top peak) is derived according to different duty cycles.

\[ I_{pp} = \frac{V_o}{4L_o} (1-D)T_s \]  

The current ripple in function of input \( V_{in} \) is derived as

\[ I_{pp} = \frac{V_o(V_{in} - nV_o)}{2L_o(2V_{in} - nV_o)} T_s \]  

The ripple reduction in the proposed converter is about 60 - 80 % during the input voltage range compared with the conventional PSFB.

IV. DESIGN CONSIDERATIONS

To validate the features of the proposed converter, a prototype converter has been designed using the following procedures, with the specifications that the input voltage \( V_{in} = 390-410 \) V; the output voltage \( V_o = 205 \) V and the switching frequency \( f_s = 60 \) KHz.

E. Selection of Resonant Frequency

To reduce the current stresses of rectifier diodes and conduction losses the resonant frequency, which is decided by the leakage inductance and the rectifier capacitance, should be selected using

\[ c_{o1} = c_{o2} = \frac{n^2}{8\pi^2 L_{lk} f_s^2} \left[ \frac{f_r}{f_s} \right]^2 \]

And Fig. 6 after the leakage inductance is determined. To analyze the converter, two quantities are defined as frequency ratio.

\[ F = \frac{f_r}{f_s} \]
\[ Q = \frac{4\omega L_k}{R_o} \]  

\[ \Delta t_{\text{dead}} = C_0 \frac{V_{\text{in}}}{I_m(t_1)} / 2 \]  

V. EXPERIMENTAL RESULTS

A 540-W prototype of the proposed converter was built. The parameters of this prototype circuit are listed in Table 1. Experimental waveforms are shown in Fig. 7. As mentioned above due to the lack of output inductor the voltage stress across the rectifier is highly reduced, we can obtain high voltage without ringing. In addition with the help of leakage inductance of the transformer the circulating loss is eliminated. So during the freewheeling period, only a small magnetizing current is flowing through the primary side of the transformer. With the help of clamping diode the voltage stress across the rectifier diode is considerably reduced. The resonance between leakage inductance of the transformer and rectifier capacitor reduce the current stress, voltage spike of the rectifier diodes and conduction losses. Using voltage doubler circuit we can obtain high voltage without any additional circuits. From the experimental results we conclude that, this circuit is applicable for high voltage and high power applications.

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<tr>
<th>Item</th>
<th>Symbol</th>
<th>Value/Part</th>
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<td>Input voltage</td>
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<tr>
<td>Output voltage</td>
<td>( V_o )</td>
<td>205V</td>
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<tr>
<td>Max. Power Rating</td>
<td>( P_{\text{max}} )</td>
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<tr>
<td>Switching frequency</td>
<td>( F_s )</td>
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<td>Turns Ratio</td>
<td>( n : l )</td>
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<tr>
<td>Leakage inductance</td>
<td>( L_{\text{in}} )</td>
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<tr>
<td>Magnetizing Inductance</td>
<td>( L_m )</td>
<td>1.4mH</td>
</tr>
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<td>Capacitance of double cell</td>
<td>( C_1, C_2 )</td>
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<td>Output capacitance</td>
<td>( C_o )</td>
<td>1000 ( \mu ) F</td>
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<tr>
<td>Rectifier Diodes</td>
<td>( D_{1-2} )</td>
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</tbody>
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Fig. 7 (a). Gate pulse of \( S_1 \)

Fig. 7 (b). Gate pulse of \( S_2 \)

Fig. 7 (c). Gate pulse of \( S_3 \)

Fig. 7 (d). Gate pulse of \( S_4 \)

Fig. 7 (e). Voltage across primary winding

Fig. 7 (f). Voltage across secondary winding

Fig. 7 (g). Output voltage

Fig. 7 (h). Output current
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[26] Measured Efficiency (%)

[35] 100

[42] 84

[42] 86

[42] 88

[42] 90

[42] 92

[42] 94

[42] 96

[42] 98

[47] [3]

[47] [4]

[47] [2]

[47] [1]

Fig. 8. Efficiency comparison under load variation

VI. CONCLUSION

A new soft switched full bridge converter with voltage doubler type rectifier is presented. By employing an auxiliary clamp diode the voltage stress across the rectifier diode is reduced, and low voltage rate diode can be utilized to reduce the conduction loss. Also the lead, lag switches achieves ZVS on. In addition since all energy that is stored in the leakage inductor of the transformer is transferred to the output side, the circulating energy is considerably reduced. A prototype has been used in the experiments to prove the validity of the proposed converter. The measured efficiency with wide load range, i.e., as high as around 98%, demonstrated higher efficiency than in conventional PSFB converters. Therefore, the improved efficiency of the proposed converter demonstrates it is suitable for high voltage and high power applications.

REFERENCES


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