A novel phase-shift full-bridge converter with voltage-doubler and decoupling integrated magnetics in PV system

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Abstract. A novel phase shift full bridge (PSFB) converter with voltage-doubler and decoupling integrated magnetics in photovoltaic (PV) systems is proposed. Considering the demand that the output voltage is higher than the input voltage in PV systems, the voltage-doubler is added to achieve higher voltage gain compared with the traditional PSFB. In order to avoid current oscillation caused by the voltage-doubler and obtain the wide zero voltage switching (ZVS) ranges, an external inductor is imposed on the circuit. Especially, to obtain much higher power density, the external inductor and transformer are integrated into one magnetic core. The operation and voltage gain of proposed converter are analyzed. Also, in order to reveal the effects the integrated magnetics gives to the converter, the decoupling condition and the expression of leakage inductor of integrated magnetics are obtained in detail. Finally a 100 W prototype converter is made and the experimental results are given to verify the analysis.

Key words: integrated magnetics, voltage-doubler, PSFB converter, PV system.

1. Introduction

Using the photovoltaic (PV) power system as an alternative energy resource has been widely discussed due to the rapid growth of power electronics techniques [1–3]. There are three different systems widely used in PV applications—the centralized inverter system, the string inverter system, and the module integrated converter system [3, 4]. The multi-string inverter is the further development of the string inverter, where several strings are interfaced with their own dc-dc converter to a common dc-ac inverter [5–9]. The main challenge is to develop a dc-dc converter that can lift the low voltage.

A Phase-Shift Full-Bridge (PSFB) converter is widely used for high voltage high power applications due to its advantages such as simple structure and zero-voltage switching (ZVS) [10–15]. However, full ZVS operation can only be achieved in a limited load and input-voltage range, unless a relatively large inductance is provided in series with the primary winding of the transformer which can be implemented by adding an external inductor. Several converters achieve ZVS in all primary switches over an extended load and input voltage range by utilizing energy stored in the inductive components of an auxiliary circuit [16–20]. Unfortunately, these converters are not appropriate to achieve high power density, high reliability, and low cost because of extra devices and/or complicated control circuitry.

In the modern power electronics industry the demand for integrated magnetics has become much stronger because of the following reasons [21–28]:

1) The number of magnetic components can be reduced.
2) The size and cost of power electronic circuits can be reduced.
3) Sometimes a controlled coupling between magnetic components is required to achieve special functions.

In brief, it is an effective method to improve power density. When the PSFB converter is used in PV system as dc-dc converter, it has two main problems:

1) The voltage gain is higher, it means that the output voltage is much higher than the input voltage though the high voltage gain can be solved through the increasing of turns ratio of high-frequency transformer, it decreases the power density due to the increasing of the volume of transformer.
2) Due to the high voltage gain, the number of turns of secondary winding of transformer is more than that of primary winding of transformer, therefore, the leakage inductor of transformer used as resonant inductor to achieve wide ZVS range is not enough.

In this paper, in order to overcome above problems, a novel PSFB converter with voltage-doubler and decoupling integrated magnetics is proposed. The voltage-doubler is adopted to increase voltage gain. In order to avoid current oscillation that caused by the voltage source parallel with capacitors of voltage-doubler directly through transformer, an external inductor between voltage source and voltage-doubler is set. Moreover, the external inductor is also used as resonant inductor to achieve wide ZVS range, therefore, the problem that only leakage inductor of transformer is used to achieve wide ZVS range is not enough is solved. Especially, in order to improve power density and save volume of magnetics, the external inductor and transformer are integrated into one magnetic core. The integrated magnetics are decoupled in order not to
influence the operation of converter, and the leakage inductor of integrated magnetics is also used as resonant inductor to achieve ZVS besides the external inductor.

The paper studies the operation of the individual modes of the converter thoroughly. The key features in the individual modes are also discussed in detail. The theoretical analyses are done to explain the converter operations including the realization of ZVS and the calculation of voltage gain. Also, the decoupling condition of integrated magnetics is analyzed and the leakage inductor expression of integrated magnetics is obtained. Experimental results of a 100-W converter with 48-V input and 380-V output are provided at the end of the paper to confirm the theoretical analysis.

2. Operational principles

Figure 1 shows the circuit diagram of the conventional PSFB converter, in order to increase voltage gain and save volume of magnetics, conventional rectifier circuit is replaced by voltage-doubler and transformer T are integrated into one magnetic core, the proposed converter is shown in Fig. 2. In Fig. 2 the winding of inductor Lr and transformer T are in the left and right legs, and the winding of transformer T is in the middle leg. The operating waveforms of the proposed converter in the steady state are shown in Fig. 3.

In order to perform a mode analysis, several assumptions are made as follows:

1) Turns ratio of transformer T is n, and Lr includes leakage inductor.
2) Switches S1-4 are ideal except for parasitic capacitors (Coss1 = Coss2 = Coss3 = Coss4 = Coss) and internal diodes (D1 = D2 = D3 = D4).
3) The output voltage V0 is constant.
4) Voltage doubler is ideal, rectifier capacitors C1 and C2 are identical, and each voltage is V0/2, Dr1 and Dr2 are rectifier diodes.

![Fig. 1. Conventional PSFB converter](image1)

![Fig. 2. Proposed PSFB converter with voltage-doubler and decoupling integrated magnetics](image2)

As shown in Fig. 3, each switching period is subdivided into six modes and their topological states are shown in Fig. 4. From t0 to t1, since the primary current ipri(t) and idr1(t) are continuous, that is, the voltage V0 of secondary side of transformer T is the voltage of C1, which means V0 = V0/2.

1) Mode 1 (t0 - t1). The input power is transferred to the secondary side through S1 and S3. Dr1 is turned on and S1 is charged by idr1(t), therefore, the voltage V0/2n is reflected on the primary side of T, and V01 - V0/2n is put on Lr, that is, Lr determines the slope of the primary current ipri(t), and it can be described as follows:

\[
i_{pri}(t) = \frac{V_{in} - V_0/2n}{L_r}(t - t_0).
\]

2) Mode 2 (t1 - t2). When S1 is turned off, Coss1 and Coss3 are charged and discharged by resonance with Lr respectively, and then the primary current ipri(t) and the voltage of Coss1 and Coss3 can be described as follows:

\[
i_{pri}(t) = i_{pri}(t_1) \cos \omega(t - t_1)
\]

\[
v_{coss1}(t) = i_{pri}(t_1) \cdot Z \cdot \sin \omega(t - t_1)
\]

\[
v_{coss3}(t) = V_{in} - i_{pri}(t_1) \cdot Z \cdot \sin \omega(t - t_1)
\]

where \(\omega = \frac{1}{\sqrt{2}C_{oss} \cdot L_r}, \ Z = \sqrt{\frac{L_r}{2C_{oss}}} \).
3) Mode 3 ($t_2 - t_3$). The voltage of $C_{oss3}$ is discharged to zero, the anti-diode $D_1$ of switch $S_3$ is turned on naturally, and then the $S_3$ can be turned on to achieve ZVS. The primary current $i_{pri}(t)$ can be described as follows:

$$i_{pri}(t) = i_{pri}(t_2) - \frac{V_o/2n}{L_r}(t - t_2). \quad (5)$$

4) Mode 4 ($t_3 - t_4$). $S_1$ is turned off, $C_{oss2}$ and $C_{oss4}$ are discharged and charged by resonance with $L_r$ respectively. Then the primary current $i_{pri}(t)$, and the voltage of $C_{oss2}$ and $C_{oss4}$ can be described as follows:

$$i_{pri}(t) = i_{pri}(t_3) \cos \omega(t - t_3) \quad (6)$$

$$v_{oss2}(t) = V_{in} - i_{pri}(t_3) \cdot Z \cdot \sin \omega(t - t_3) \quad (7)$$

$$v_{oss4}(t) = i_{pri}(t_3) \cdot Z \cdot \sin \omega(t - t_3) \quad (8)$$

where $\omega = \frac{1}{\sqrt{2C_{oss}L_r}}$, $Z = \sqrt{\frac{L_r}{2C_{oss}}}$.

5) Mode 5 ($t_4 - t_5$). The voltage of $C_{oss2}$ is discharged to zero, the anti-diode $D_2$ of switch $S_2$ is turned on naturally, and then the $S_2$ can be turned on to achieve ZVS. The primary current $i_{pri}(t)$ can be described as follows:

$$i_{pri}(t) = i_{pri}(t_4) - \frac{V_{in} - V_o/2n}{L_r}(t - t_4). \quad (9)$$

6) Mode 6 ($t_5 - t_6$). The primary current $i_{pri}(t)$ goes though $S_2$ and $S_3$, $D_{r2}$ is turned on and $C_{r2}$ is charged by $i_{Dr2}(t)$, therefore, the voltage $-V_o/2n$ is reflected on the primary side of $T$, and the analysis is similar with mode 1.

From mode 6, another circle is begun which is similar with mode 1 to mode 5 analyzed above.

3. Analysis of voltage gain

Compared with conventional PSFB which is widely used to step down the input voltage, the proposed converter is used to lift input voltage. In order to analyze the voltage gain $M (V_o/V_{in})$, the waveform of primary current $i_{pri}$ is shown in Fig. 5a. In Fig. 5a when $t_0 \leq t \leq t_1$, $i_{pri}$ increases linearly as $i_{pri1}$, when $t_1 \leq t \leq T$, $i_{pri}$ decreases as $i_{pri2}$ which is composed of four parts which are described in formula (2), (5), (6), (9) respectively ($t_1 \leq t \leq t_2$, $t_2 \leq t \leq t_3$, $t_3 \leq t \leq t_4$, $t_4 \leq t \leq T$). Compared with mode 3 ($t_2 \leq t \leq t_3$) and mode 5 ($t_4 \leq t \leq T$), the charging and discharging time of mode 2 ($t_1 \leq t \leq t_2$) and mode 4 ($t_3 \leq t \leq t_4$) are instantaneous and then can be omitted, which are shown in Fig. 5b. In Fig. 5b $i_{pri1}$ is described as $i_{pri1} (t_{on} \leq t \leq t_{off})$ and $i_{pri2} (t_{off} \leq t \leq T)$. $I_{max}$ and $I_{av}$ are, respectively, the max value of $i_{pri}$ and the average value of $i_{pri}$. $I_{av}$ can be described as follows:

$$I_{av}T = \int_{t_0}^{t_{on}} i_{pri1}^{+}dt + \int_{t_0}^{t_{off}} i_{pri1}^{-}dt + \int_{t_{off}}^{T} i_{pri2}^{-}dt \quad (10)$$

$i_{pri1}^{+}$, $i_{pri1}^{-}$ and $i_{pri2}^{-}$ can be described as follows:

$$i_{pri1}^{+} = \frac{V_{in} - V_o/2n}{L_r}(t - t_0) \quad (11)$$

$$i_{pri1}^{-} = \frac{V_{in} - V_o/2n}{L_r}(t - t_{on})$$

$$i_{pri2}^{-} = \frac{V_{in} - V_o/2n}{L_r}(t - t_{off})$$
4. Design of integrated magnetics

4.1. Decoupling of integrated magnetics. The proposed PSFB converter is used to lift voltage, which means that the number of turns of transformer primary winding is less than that of transformer secondary winding. Furthermore, the leakage inductor in primary winding is much smaller, therefore, the ZVS range is narrow when only the leakage inductor is used for resonance. In order to solve this problem, an external inductor is added, but it increases volume which results in the decreasing of power density. In this paper, the integrated magnetics which are shown in Fig. 6 are used to solve this problem. In Fig. 6a transformer \( T \) and inductor \( L_r \) are integrated into one EE magnetic core, \( N_{L1} \) and \( N_{L2} \) are, respectively, the number of turns of primary winding and secondary winding of transformer \( T \), \( N_P \) and \( N_S \) are, respectively, the number of turns of primary winding and secondary winding of transformer \( T \), \( N_{L1} \) and \( N_{L2} \) are the number of turns of inductor \( L_r \), \( \Re_1 \) and \( \Re_2 \) are, respectively, the magnetic resistance in middle leg, left leg and right leg, \( i_P \), \( i_s \) and \( i_L \) are, respectively, the current of primary winding, secondary winding and inductor \( L_r \), the transformer winding is in the middle leg, and the inductor winding is divided into two parts which are in left and right legs respectively, \( \varphi_{LT1} \) and \( \varphi_{LT2} \) are, respectively, the fluxes in transformer windings (middle leg) which are imposed by inductor winding (left and right legs), \( \varphi_{LT1} \) and \( \varphi_{LT2} \) are, respectively, the fluxes in inductor winding which are imposed by transformer winding, and the magnetic circuit is shown in Fig. 6b. In order to prevent their fluxes from coupling with each other, which make it difficult to analyze the operation of converter, the integrated magnetics are decoupled.

![Image](https://example.com/image.png)

Fig. 5. Waveforms of primary current: (a) waveform of primary current, (b) simplified waveform of primary current

Based on power conservation, \( I_{av} \) also can be described as follows:

\[
I_{av} = \frac{V_o^2}{V_{in} R} \tag{12}
\]

Based on (10), (11), (12), the voltage gain \( M \) \( (V_o/V_{in}) \) can be got as follows:

\[
M = \frac{V_o}{V_{in}} = \frac{D_2}{\frac{1}{2} + \frac{1}{4n^2} + \frac{1}{2} \frac{L_{sf}}{R} D_1} \tag{13}
\]

where \( D_{on}T = t_{on} - t_0 \), \( D_{off1}T = t_{off} - t_{on} \), \( D_{off2}T = T - t_{off}, f = 1/T, R \) is load,

\[
\begin{align*}
D_1 &= \frac{1}{2}D_{on}^2 + \frac{1}{2}D_{off1}^2 + D_{on}D_{off1} - \frac{1}{2}D_{off2}^2 \times \left(\frac{1}{2}D_{on}^2 + \frac{1}{2}D_{off1}^2 + D_{on}D_{off1} - \frac{1}{2}D_{off2}^2\right) \\
D_2 &= \frac{1}{2}D_{on}^2 + \frac{1}{2}D_{off2}^2 + D_{on}D_{off1} - \frac{1}{2}D_{off2}^2
\end{align*} \tag{14}
\]

From (13) and (14), it can be got that voltage gain \( M \) is decided by \( D_1 \), \( D_2 \), turns ratio \( n \), resonant inductor \( L_r \), frequency \( f \) and load \( R \). It means that such parameters should be adjusted together to satisfy the voltage gain \( M \) demanded.

![Image](https://example.com/image.png)

Fig. 6. Integrated magnetics: (a) integrated transformer and inductor, (b) equivalent magnetic circuit
Based on Fig. 6b, the fluxes $\psi_{T1}$ and $\phi_{T2}$ can be described as follows:

$$\begin{align*}
\phi_{T1} & = \frac{N_L i_L R_{22}}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}} \\
\phi_{T2} & = \frac{N_L i_L R_{21}}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}}.
\end{align*}$$

(15)

The flux linkage $\psi_{TL}$ of transformer $T$ which is imposed by inductor $L_r$ is described as follows:

$$\psi_{TL} = \frac{(N_P - N_S)(\phi_{T1} - \phi_{T2})}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}} (N_L R_{22} - N_L R_{21}).$$

(16)

The fluxes $\phi_{LT1}$ and $\phi_{LT2}$ can be described as follows:

$$\begin{align*}
\phi_{LT1} & = \frac{(N_P i_P - N_S i_S) R_{22}}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}} \\
\phi_{LT2} & = \frac{(N_P i_P - N_S i_S) R_{21}}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}}.
\end{align*}$$

(17)

The flux linkage $\psi_{LT}$ of inductor $L_r$ which is imposed by transformer $T$ is described as follows:

$$\psi_{LT} = N_L \phi_{LT1} - N_L \phi_{LT2} = \frac{N_P i_P - N_S i_S}{R_{21} R_{22} + R_1 R_{22} + R_1 R_{21}} (N_L R_{22} - N_L R_{21}).$$

(18)

When $\psi_{LT} = \psi_{TL} = 0$, the inductor $L_r$ and transformer $T$ are decoupled, the decoupling condition can be got as follow from (16) and (18):

$$\frac{R_{21}}{R_{22}} = \frac{N_{L1}}{N_{L2}}.$$  

(19)

In order to design it conveniently, make air gaps of left leg and right leg be the same, which means that $R_{21} = R_{22}$, and then the condition of decoupling is got from (19):

$$N_{L1} = N_{L2}. $$

(20)

Assuming that $B_1$, $B_{21}$ and $B_{22}$ are, respectively, the flux density of middle leg, left leg and right leg, they can be described as follows:

$$
\begin{align*}
B_1 & = \frac{(N_P i_P - N_S i_S)(R_{21} + R_{22})}{(R_{21} + R_{22}) S_1} \\
B_{21} & = \frac{i_L (N_{L1} + N_{L2})}{(R_{21} + R_{22}) S_2} - \frac{(N_P i_P - N_S i_S) R_{22}}{(R_{21} + R_{22}) S_2} \\
B_{22} & = \frac{i_L (N_{L1} + N_{L2})}{(R_{21} + R_{22}) S_2} + \frac{(N_P i_P - N_S i_S) R_{21}}{(R_{21} + R_{22}) S_2}.
\end{align*}
$$

(21)

where $S_1$, $S_{21}$ and $S_{22}$ are, respectively, the cross section areas of middle leg, left leg and right leg.

The integrated magnetics are designed to make sure that each leg is not saturated according to (21).

### 4.2. Leakage inductor of integrated magnetics

Compared with conventional PSFB converter which adopts leakage inductor of transformer as resonant inductor, the proposed PSFB converter’s resonant inductor includes two parts: (1) Inductor $L_r$ integrated in magnetics. (2) Leakage inductor of integrated magnetics. Therefore, it is necessary to analyze the leakage inductor of integrated magnetics.

![Fig. 7. Leakage flux of integrated magnetics: (a) front view of integrated magnetics, (b) top view of integrated magnetics](image)

The leakage fluxes of integrated magnetics are shown in Fig. 7. $V_L$, $V_P$ and $V_S$ are, respectively, the voltage of inductor winding, primary winding of transformer $T$, and secondary winding of transformer $T$, they can be described as follows:

$$
\begin{align*}
V_L & = N_L \frac{d}{dt}(\phi_L + \phi_{IL} + \phi_{ISL1} + \phi_{ISL2}) \\
V_P & = N_P \frac{d}{dt}(\phi_T + \phi_{IP} + \phi_{IPS}) \\
V_S & = N_S \frac{d}{dt}(-\phi_T + \phi_{IS} - \phi_{IPS} + \phi_{ISL1} - \phi_{ISL2})
\end{align*}$$

(22)

where $N_L$ ($N_L = N_{L1} + N_{L2}$) is the number of turns of inductor $L_r$, $\phi_T$ and $\phi_L$ are, respectively, the fluxes of transformer $T$ and inductor $L_r$, $\phi_{IL}$, $\phi_{IP}$ and $\phi_{IS}$ are, respectively, the leakage fluxes of inductor winding, primary winding and secondary winding of transformer $T$, $\phi_{ISL1}$, $\phi_{ISL2}$ and $\phi_{IPS}$ are, respectively, the leakage fluxes which goes through inductor winding in left leg and secondary winding of transformer $T$, inductor winding in right leg and secondary winding of transformer $T$, and they can be described as follows:
where, \( R_{ljk} \) is leakage magnetic resistance according to its corresponding leakage flux \( \phi_{ljk} \), \( i = L, S, P \), \( j = L, S, P \), \( k = 1, 2 \).

Further more, in order to get the expression of leakage inductor, the formula (22) can be deduced as follows:

\[
\begin{pmatrix}
V_L \\
V_P \\
V_S
\end{pmatrix} = \begin{pmatrix}
L + L_{IL} + L_{IL2} & 0 & 0 \\
0 & L_P + L_{IP} + L_{IPS} & -M_{ISP} - M_{SP} \\
M_{IL1} - M_{IS2} & -M_{ISP} - M_{SP} & M_{IL1} - M_{IL2} \\
M_{IS1} - M_{IS2} & -M_{ISP} - M_{SP} & M_{IS1} - M_{IS2}
\end{pmatrix}
\begin{pmatrix}
\frac{dL}{dt} \\
\frac{dP}{dt} \\
\frac{dS}{dt}
\end{pmatrix}
\]

(24)

where

\[
\begin{align*}
L &= \frac{N_i^2}{R_{l1} + R_{l2}}, \quad L_{IL} = \frac{N_i^2}{R_{IL}}, \\
L_{IL1} &= \frac{N_{IP}N_S}{2R_{lSL1}}, \quad L_{IL2} = \frac{N_{IS}^2}{2R_{lSL2}}, \\
M_{IL1} &= \frac{N_{IP}N_S}{R_{l1} + R_{l2}}, \quad M_{IL2} = \frac{N_{IS}^2}{R_{l1} + R_{l2}}, \\
L_P &= \frac{N_i^2}{R_{l1} + R_{l2}}, \quad L_{IP} = \frac{N_{IP}N_S}{R_{lPS}}, \\
M_{IP} &= \frac{N_{IP}N_S}{R_{lPS}}, \\
M_{IPS} &= \frac{N_{IP}N_S}{R_{lIP}}, \\
M_{PS} &= \frac{N_{IP}N_S}{R_{lPS}}, \\
L_{IS} &= \frac{N_i^2}{R_{l1} + R_{l2}}, \quad L_{IS} = \frac{N_i^2}{R_{IS}}, \\
L_{ISP} &= \frac{N_i^2}{R_{lPS}}, \quad L_{ISP} = \frac{N_{IS}^2}{R_{lISP}}, \\
M_{IS1} &= \frac{N_{IP}N_S}{R_{lSL1}}, \quad M_{IS2} = \frac{N_{IP}N_S}{R_{lSL2}}, \\
M_{ISP} &= \frac{N_{IP}N_S}{R_{lIP}}, \\
M_{IS} &= \frac{N_{IS}^2}{R_{lIS1}}, \quad M_{IS} = \frac{N_{IS}^2}{R_{lIS2}}
\end{align*}
\]

(25)

The total leakage inductor \( L_i \) for resonance can be described as follows:

\[
L_i = L_{LI} + L_{TPi} + \frac{N_i^2}{N_S}L_{TSi}
\]

(26)

where, \( L_{LI} \) is total leakage inductor of the inductor \( L_r \), \( L_{TPi} \) and \( L_{TSi} \) are respectively, the total leakage inductor of primary and secondary winding of transformer \( T \), and they can be described as follows:

\[
\begin{align*}
L_{LI} &= L_{IL} + L_{IL1} + L_{IL2} + M_{IL1} - M_{IL2} \\
L_{TPi} &= L_{IP} + L_{IPS} - M_{IPS} \\
L_{TSi} &= M_{IS1} - M_{IS2} + L_{IS} + L_{ISP} + L_{IS1} \\
&\quad + L_{IS2} - M_{ISP}
\end{align*}
\]

(27)

Since leakage magnetic resistance \( R_{lSL1} \) equates to \( R_{lSL2} \) approximately, and then \( M_{IL1} = M_{IL2}, M_{IS1} = M_{IS2} \), the formula (27) can be simplified as follows:

\[
\begin{align*}
L_{LI} &= L_{IL} + L_{IL1} + L_{IL2} \\
L_{TPi} &= L_{IP} + L_{IPS} - M_{IPS} \\
L_{TSi} &= L_{IS} + L_{ISP} + L_{IS1} + L_{IS2} - M_{ISP}
\end{align*}
\]

(28)

According to (28), \( L_{LI} \) is only composed of its self leakage inductors, since the mutual leakage inductors \( M_{IL1} \) and \( M_{IL2} \) between inductor \( L_r \) and transformer \( T \) are counteracted by each other, it isn’t influenced by transformer. The leakage inductor of transformer \( T \) is composed of \( L_{TPi} \) and \( L_{TSi} \), since the mutual leakage inductors \( M_{IS1} \) and \( M_{IS2} \) between \( L_r \) and transformer \( T \) are counteracted by each other, it isn’t influenced by inductor \( L_r \). Therefore, besides the fluxes between inductor \( L_r \) and transformer \( T \), the leakage fluxes between them are also decoupled. The decoupling integrated magnetics can be seen as discrete inductor \( L_r \) and transformer \( T \), and then the leakage inductors can be measured by common short experiment.

5. Experimental results

5.1. Decoupling of integrated magnetics. The parameters of integrated magnetics are shown as follows: the resonant inductor is \( L_r = 12.5 \mu H \), the inductor of transformer primary winding is \( L_{TP} = 138 \mu H \), the inductor of transformer secondary winding is \( L_{TS} = 2.8 \mu H \), the leakage inductor of transformer primary winding is \( L_{TPi} = 2.1 \mu H \), the leakage inductors of transformer secondary winding is \( L_{TSi} = 24 \mu H \), the turns ratio is \( N_S : N_P = 4.5 \), the air gap of three legs are the same are 0.4mm. \( L_{TPi} \) and \( L_{TSi} \) are measured by short circuit experiment, since \( L_r \) less than 10% of \( L_{TPi} \), compared with \( L_{TPi} \), \( L_{LI} \) is so small that it can be omitted. Based on (26) it can be calculated that \( L_1 = 4.2 \mu H \). In order to test the decoupling of integrated magnetics, a 5 V and 100 KHz sinusoidal voltage source is imposed on integrated magnetics. When voltage source is imposed on the primary winding of transformer \( T \), the waveform of the voltage of transformer secondary side \( v_s \) and inductor \( v_l \) are shown.
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In Fig. 8a, \( v_s \) is 22.6 V according to turns ratio, and \( v_l \) is zero approximately, it means that the voltages of transformer and inductor don’t influence on each other due to decoupling. When voltage source is imposed on the inductor winding, the waveform of the voltage of transformer secondary side \( v_s \) and primary side \( v_p \) are shown in Fig. 8b. In Fig. 8b if inductor and transformer are coupled closely, according to turns ratio \((N_L : N_P = \sqrt{L_{TP}/L_r} = 3.3 \) and \(N_L : N_S = \sqrt{L_{TS}/L_r} = 14.8)\) \( v_p \) should be 16.5 V and \( v_s \) should be 74 V, from Fig. 8b it can be seen that \( v_p \) and \( v_s \) are less than 10% of 16.5 V and 74 V, therefore, inductor and transformer are decoupled with each other approximately.

![Fig. 8. Waveforms of decoupling integrated magnetics: (a) waveforms of \( v_s \) and \( v_l \), (b) waveforms of \( v_p \) and \( v_s \)](image)

5.2. PSFB converter. Based on the designed parameters, a 100 kHz, 100 W, 380 V prototype dc-dc power module with an input voltage range of 48 V has been constructed. The parameters are as follows: IRF3815s and BYM36Cs are used as switching devices and voltage-double diodes, voltage-double capacitors are selected as 0.1 \( \mu F \). The key experimental waveforms are shown in Figs. 9–13, it can be seen that all the waveforms agree well with the theoretical analysis. Figure 9 shows the primary current \( i_{pri} \) with different \( L_r \). In Fig. 9a when \( L_r = 7.3 \) uH the \( i_{pri} \) decreases to zero before next circle due to the small \( L_r \), therefore, the oscillation of \( i_{pri} \) is caused. In Fig. 9b and c, with the increasing of \( L_r \) the \( i_{pri} \) becomes continuous, the oscillation of \( i_{pri} \) is reduced, and the ZVS range is widen. Figure 10 shows the waveform of \( v_{AB} \), when \( v_{AB} \) becomes zero the leading switches can achieve ZVS. Figure 11 shows the waveforms of lagging switch, it can be seen that when gate signal \( v_{GS} \) is on, the voltage of switch \( v_{DS} \) has been decreased to zero, and then the lagging switch achieves ZVS. Figures 12–13 show the waveforms of \( v_{TS} \) and the voltage \( v_{Cr1} \) and \( v_{Cr2} \) of voltage-doubler, when \( v_{Cr1} \) is charged \( v_{TS} \) equates to \( v_{Cr1} \), when \( v_{Cr2} \) is charged \( v_{TS} \) equates to \( v_{Cr2} \) since \( V_o = v_{Cr1} + v_{Cr2} \), the output voltage \( V_o \) doubled the \( v_{TS} \). Figure 14 shows the converter efficiencies under some load conditions. Figure 15 shows the photograph of Prototype 100-W converter.
6. Conclusions

In this paper, a novel PSFB converter with voltage-doubler and integrated magnetics in PV system is proposed. The voltage-doubler is adopted to increase voltage gain and an external inductor which is used not only to avoid current oscillation but also as resonant inductor to achieve ZVS is added. Especially, the external inductor and transformer are integrated into one magnetic core to improve power density. The operational principles have been presented by the mode analysis, and the design equations for the achievement of ZVS and the voltage gain are derived. Moreover, the decoupling condition of integrated magnetics and the leakage inductor expression of integrated magnetics are obtained. Based on the design of ZVS, voltage gain and integrated magnetics, a prototype has been designed to prove the validity of the proposed converter. The experimental results of a 100 W prototype converter have been presented. The efficiency of the proposed converter is obtained about 93% at a rated condition.

The proposed converter is suitable for the multi-string inverter as dc-dc converter in PV system requiring the high efficiency and high power density.

REFERENCES

A novel phase-shift full-bridge converter with voltage-doubler and decoupling integrated magnetics in PV system


