전력전자학술대회 논문집 2004. 7.12 ~ 7.15

# 새로운 무 손실 다이오드 클램프 회로를 채택한 두 개의 트랜스포머를 갖는 영 전압 스위칭 풀 브릿지 컨버터

윤현기, 한상규, 박진식, 문건우, 윤명중

한국과학기술원

# Zero-Voltage Switching Two-Transformer Full-Bridge PWM Converter With Lossless Diode-Clamp Rectifier

H.K. Yoon, S.K. Han, J.S. Park, G.W. Moon, and M.J. Youn

KAIST

***Abstract* — The two-transformer full bridge (TTFB) PWM converter has two transformers which act as the output inductor as well as the main transformer, *i.e.* as the forward and the flyback transformer. Although the doubled leakage inductor of the TTFB makes it easier to achieve the zero-voltage switching (ZVS) of the lagging leg switch along the wide load range, it instigates a serious voltage ringing in the secondary rectifier diodes, which would require the dissipative snubber circuit, cause the serious power dissipation, and increase the voltage stress across those diodes. To overcome these problems, a new lossless diode-clamp rectifier (LDCR) is employed as the output rectifier, which helps the voltage across rectifier diodes to be clamped on a half the output voltage ($V_o/2$) or the output voltage ($V_o$). Therefore, no dissipative snubber for rectifier diodes is needed and a high efficiency as well as low noise output voltage can be realized. The operations, analysis and design consideration of proposed converter are presented in this paper. To verify the validity of the proposed converter, experimental results from a 425W, 385-170Vdc prototype for the plasma display panel (PDP) sustaining power module (PSPM) are presented.**

## I. INTRODUCTION

Among various PWM DC/DC converters operating at the high frequency, the conventional Phase Shift Full Bridge (PSFB) converter is a preferred topology for middle/high power applications (over 400W), because this topology permits all switching devices to operate under the ZVS in the high frequency with the use of parasitic components [1].

However, the PSFB has a serious disadvantage such as the narrow ZVS range of the lagging leg. During the transition of the leading leg switch, the reflected large output inductor current to the transformer primary side can achieve the ZVS by discharging that switch output capacitor. However, during the transition of the lagging leg switch, the only energy stored in the small leakage inductor discharges that switch output capacitor. Therefore, since the energy available for charging and discharging the lagging leg switch output capacitor is insufficient, the hard switching operation of that switch is inevitable at the light load.

To achieve the ZVS of the lagging leg along the wide load range, PSFB is required to have the large leakage inductor. However, one accompanied problem of increasing the leakage inductance is that it can reduce the effective duty cycle, which causes the large circulating energy in the converter and degrades the overall system efficiency. The other problem caused by the large leakage inductance is the excessive voltage overshoot and ringing across the output rectifier diodes, due to the interaction between the primary leakage inductance of the transformer and the parasitic junction capacitance of the secondary rectifier diodes. To absorb the serious voltage ringing across the secondary rectifier diode, the dissipative resistor-capacitor (RC) snubber must be added to that rectifier diode. Unfortunately, since the energy stored in the snubber capacitor is not only very large but also dissipated through the snubber resistor, this RC snubber seriously degrades the overall system efficiency. Especially, in case of the high output voltage applications like the PSPM, it would be more serious.

To relieve the narrow ZVS range of the lagging leg switch in the conventional PSFB, the TTFB has been proposed in Fig.1 [2]. It has two transformers acting as not only a main transformer but also an output inductor, performing alternately as a forward transformer and a flyback transformer. Namely, while one transformer transfers the input power to the secondary side of the converter, the other stores the power as a flux form. The other pair of diagonally opposite switches is turned on thereafter, and continues this pattern of action. Therefore, since the series-connected two transformers can replace both a main transformer and output inductor, the inductor is not needed in the secondary side. In addition, since it has series connected two transformers, it has the somewhat large doubled leakage inductor, which can expand the ZVS range of the lagging leg switch. At a lighter load, since the magnetizing current of each transformer contributes to the ZVS operation, the ZVS of the lagging leg switch can be successfully achieved along the wide load range.

Nevertheless, the serious voltage ringing in the secondary rectifier diodes also exists in the TTFB. Since the total leakage inductance is naturally doubled due to the series-connected two transformers, it also instigates a serious voltage ringing in the secondary rectifier diodes like the conventional PSFB. Therefore, considering the abovementioned voltage ringing across output rectifier diodes, very high-voltage rated, low performance, and high-cost diode must be used for the secondary rectifier with the dissipative RC snubbers.

To overcome this problem, a ZVS TTFB PWM converter equipped with a new LDCR for the PSPM is proposed as shown in Fig.2. The primary side of the proposed converter is the same as TTFB and the secondary side consists of 5 diodes and 2 clamp capacitors, which is operated as a voltage-doubler. Especially, output rectifier diodes Do1b, Do2b can be clamped on a half the output voltage (Vo/2) and Dr, Do1a, Do2a on the output voltage (Vo), where Do1a, Do2a , Do1b, Do2b are high current fast recovery diodes for the powering purpose and Dr low current (below 1A) fast recovery diodes for the clamping purpose. Therefore, the proposed converter does not need any RC snubber and the voltage stress across rectifier diodes can be considerably reduced without voltage ringing, which favorably provides the high efficiency, low component temperature, high reliability, and high performance.

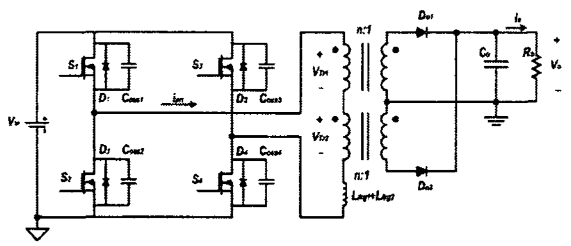


Fig. 1 Conventional two transformer full bridge converter

## II. OPERATION OF THE PROPOSED CONVERTER

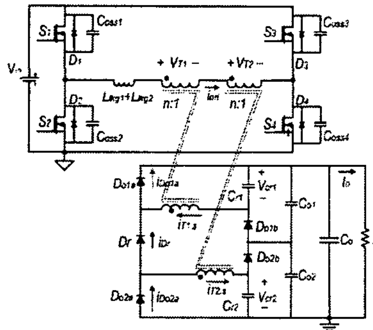


Fig. 2 Schematic diagram of the proposed circuit

Fig. 3 shows the operational key waveforms of the proposed circuit. The basic operation of the proposed converter is the same as that of the conventional TTFB. One half periods can be divided by 5 modes and its equivalent circuits are shown in Fig.4.

The operations at modes 1~5 are the same as those at modes 6~10 except for the current direction through switches S2 and S3. The switches of each leg (i. e. S1/S2 are leading leg switches and S3/S4 are lagging leg switches) turn on and off alternately with constant duty ratio and the phase difference between two legs determines the operational duty cycle of the converter where $D_{eff}T_s$ is the conduction time of each switch, $D_{free}T_s$ the phase shifted time, and $D_{lkg}T_s$ the loss time in a powering mode due to the leakage inductor.

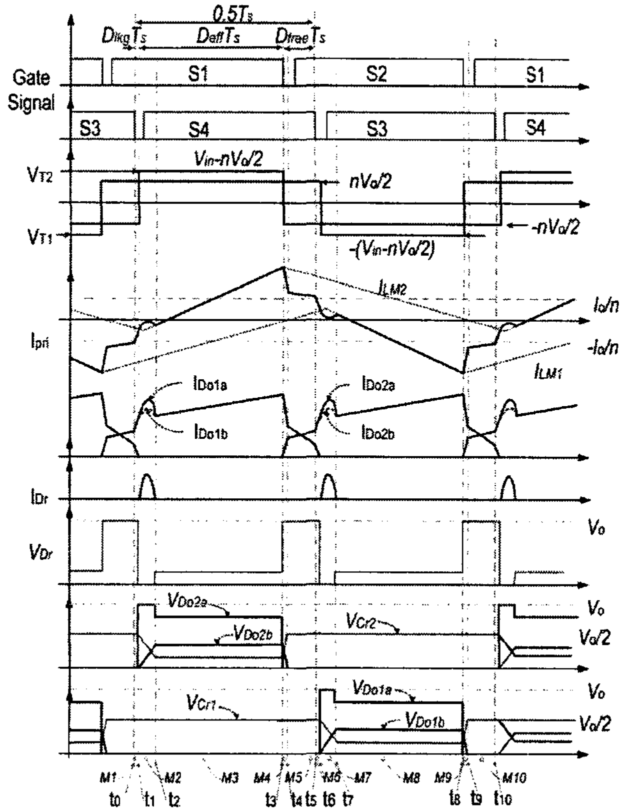
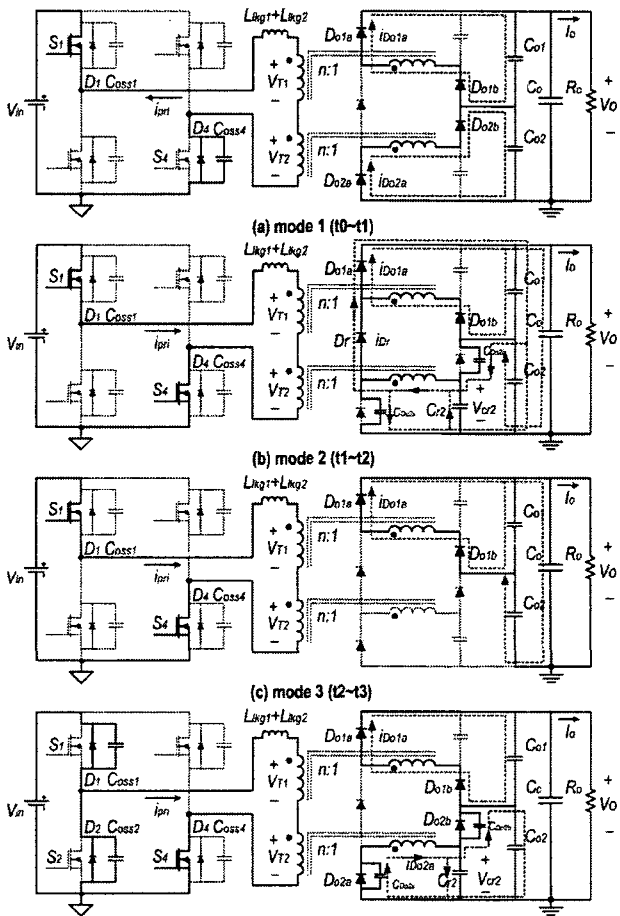


Fig. 3 Operational key waveforms of the proposed circuit

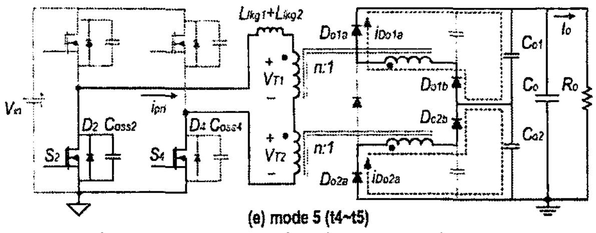


(a) mode 1 (t0~t1)

(b) mode 2 (t1~t2)

(c) mode 3 (t2~t3)

(d) mode 4 (t3~t4)



(e) mode 5 (t4~t5)

Fig. 4 Equivalent circuits of the proposed converter

For the convenience of the mode analysis in the steady state, several assumptions are made as follows:

- Power switches (S1~S4) are ideal except for their internal diodes (D1~D4) and output capacitors ($C_1$~$C_4$= $C_{oss}$).
- Rectifier diodes are ideal except for their internal junction capacitances.
- $C_{o1}$, $C_{o2}$, and $C_o$ are large enough to be constant voltage sources $V_o/2$, $V_o/2$ and $V_o$, respectively.
- Two transformers are identical. ($L_{m1}=L_{m2}=L_m$, $L_{lkg1}=L_{lkg2}=L_{lkg}$)

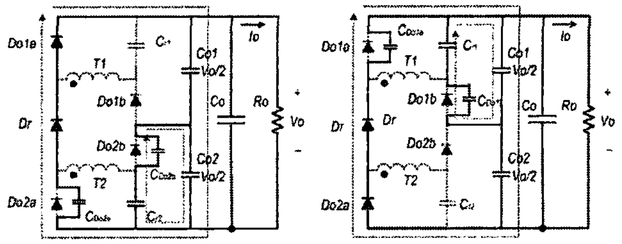
**Mode 1 ($t_0$~$t_1$):** After S3 is turned off at time $t_0$, the primary current flows back to the source through the output capacitor ($C_{oss4}$) of S4. The primary current charges the output capacitor ($C_{oss3}$) of S3 and discharges that of S4. However, the leakage inductance and primary current are low, ZVS condition of S4 is not enough. Thus, to achieve the ZVS condition, magnetizing inductance is used at next mode. The currents flows through secondary rectifier diodes, not commutated perfectly, the output voltage $V_o/2$ is reflected to the primary sides of both transformers in opposite polarity. Therefore, $V_{in}$ is applied to the total leakage inductor $L_{lkg1}+L_{lkg2}$ in the primary main power path. The primary current can be expressed as follows:

$$i_{pri}(t) = \frac{V_{in}}{2L_{lkg}}(t-t_0)+i_{pri}(t_0) \tag{1}$$

Since the freewheeling current is small, this mode ends fast. Mode 1 is ended at time $t_1$ when the commutation between the currents flowing through the rectifier diodes is completed.

**Mode 2 ($t_1$~$t_2$):** After the completion of commutation, the magnetizing inductance of transformer can help to achieve the ZVS of S4. Thus, it is easier to discharge the output capacitor ($C_{oss4}$) of S4 due to large magnetizing inductance. After the completion of the ZVS of S4, the switch S4 is conducting and is turned on.

The transformer T1 which acts as forward type transformer is transferring power to the secondary side, since rectifier diodes are commutated completely at mode1, the currents can flow through the rectifier diode $D_{o1a}$ and diode $D_{o1b}$. At the same time, the voltage of the diode $D_{o2a}$ is clamped on $V_o$ and that of the diode $D_r$ is discharging to 0 rapidly. Since the diode $D_{o1a}$ and diode $D_r$ are conducting at mode 2, the voltage of diode $D_{o2a}$ is clamped on $V_o$ by output voltage source shown in Fig.5 (a). As shown in Fig. 5(a), diodes $D_{o2b}$ and $C_r$ are clamped under $V_{Co2}$(=$V_o/2$) at mode 2.(The same operations in mode 7 as shown in Fig. 5 (b), diodes $D_{o1b}$ and $C_r$ are clamped under $V_{Co1}$(=$V_o/2$) and diodes $D_{o1a}$ and $D_r$ on $V_O$.) When the diode $D_r$ is conducting, the resonance between the leakage inductance ($L_{lkg1}+L_{lkg2}$) and equivalent secondary capacitance (which is output rectifier diodes junction capacitance ($C_{Do2b}$) and clamping capacitance ($C_{r2}$)) is happened. With the initial conditions of $i_{pri}(t_1)=I_{pri.t1}$, $V_{cr2}(t_1)=V_o/2$ and $V_{Do2b}(t_1)=0$, the current $i_{Dr}$ flowing through $D_r$ charges $C_{Do2b}$ and discharges $C_{r2}$ by the resonance operation.



(a) (b)

Fig. 5 Clamped paths of output rectifier diodes
(a) Clamped paths of $D_{o2a}$, $D_{o2b}$, $D_r$ in mode 2
(b) Clamped paths of $D_{o1a}$, $D_{o1b}$, $D_r$ in mode 7

The resonant current of the diode $D_r$ and the clamping voltage of the diode $D_{o2b}$ and $C_r$ can be expressed as follows:

$$i_{Dr}(t) = |nI_{pri.t1}|(1-\cos(w_O(t-t_1)))+\frac{V_{Llkg.t1}}{Z_O}\sin(w_O(t-t_1)) \tag{2}$$

$$v_{Do2b}(t) = (\frac{V_{in}}{n}-V_o)-|nI_{pri.t1}|Z_O\sin(w_O(t-t_1))-V_{Llkg.t1}\cos(w_O(t-t_1)) \tag{3}$$

$$v_{Cr2}(t) = (\frac{3}{2}V_o-\frac{V_{in}}{n})+|nI_{pri.t1}|Z_O\sin(w_O(t-t_1))+V_{Llkg.t1}\cos(w_O(t-t_1)) \tag{4}$$

*where*

$$w_O=\frac{n}{\sqrt{(2L_{lkg})(C_{r2}+C_{Do2b})}},\ Z_O=\frac{1}{n}\sqrt{\frac{(2L_{lkg})}{(C_{r2}+C_{Do2b})}}\ \text{and}\ V_{Llkg.t1}=\frac{V_{in}-nV_o}{n}$$

The current rating of clamping diode $D_r$ is estimated by eq.(2) and the voltage rating of output rectifier diodes $D_{o2b}$ and $D_{o1b}$ is estimated by eq.(3).

Mode 2 is ended at time $t_2$ when the conduction of the clamping diode $D_r$ is completed.

**Mode 3 ($t_2$~$t_3$):** After conduction of clamping diode $D_r$, since the secondary voltage $V_{T1.sec}$ (=$V_{in}/n$ - $V_O/2$) is smaller than the output voltage $V_o$, the rectifier diodes ($D_r$, $D_{o2a}$ and $Do_{2b}$) are reverse biased. The other rectifier diodes ($D_{o1a}$ and $D_{o1b}$) are conducting, the transformer T1 which acts as forward type transformer is transferring power to the secondary side. The transformer T2 acts as inductor and determines the slope of the primary current in this mode.

The energy stored in $L_{m.T2}+2L_{lkg}$ during this mode should be discharged through $D_{o2a}$ and $D_{o2b}$ in all modes except for mode 2~3.

Therefore, $i_{pri1}(t)$ can be expressed as follows:

$$i_{pri}(t) = \frac{V_{in}-nV_O/2}{L_m+2L_{lkg}}(t-t_2)+i_{pri}(t_2) \tag{5}$$

Mode 3 is ended at time t3 when S1 in turned off.

**Mode 4 ($t_3$~$t_4$):** When the switch S1 is turned off, primary current charges and discharges the output capacitor of S1 and S2, respectively and discharges secondary equivalent capacitance($C_{eq.s}$). It is leading leg transition to free-wheeling state. In this mode, primary current is dropped instantaneously by $C_{oss2}$ and $C_{eq.s}$. This mode has two resonant phases. First phase is that $V_{Coss2}$ is discharged from $V_{in}$ to zero which is ZVS operation of S2. It is easy to achieve the ZVS of S2 because the magnetizing inductor $L_{m1}$ still acts as an inductor and ipri($t_3$) $=I_{pri.t3}$ is large. Second phase is that $V_{Do2a}$ starts from $V_o$ to zero and $V_{Do2b}$ starts $V_o/2$ to zero. Considering the heavy load case, since the first phase is operated quickly, the second phase dominates in current drop. Assuming the first phase is completed, the primary current is can be expressed as follows:

$$i_{pri}(t) = I_{pri.t3} - (n\frac{V_o}{2})\sqrt{\frac{C_{eq.s}}{2L_{lkg}}}\sin(w_1(t-t_3)) \tag{6}$$

*where*

$$w_1 = \sqrt{\frac{1}{C_{eq.d}(2L_{lkg})}},\ C_{eq.d} = \frac{C_{eq.s}}{n^2}\ and\ C_{eq.s} = (C_{do2b} + C_{r2})$$

$C_{eq.d}$ is the sum of the equivalent capacitance of rectifier diodes and clamping capacitor of non-conduction rectifier at primary side. The clamping capacitor ($C_{r2}$) dominates to decrease primary current. It can reduce the circulating current in freewheeling state.

At the end of this mode, the $V_{T1}$+$V_{T2}$, which is two transformers voltage of primary side, is dropped to zero and secondary rectifier diode $D_{o2a}$, $D_{o2b}$ are conducting. Secondary rectifier diode $D_{o1a}$, $D_{o1b}$, $D_{o2a}$ and $D_{o2b}$ begins commutation.

**Mode 5 ($t_4$~$t_5$):** The primary current flows through S4 and S2 (and D2). Secondary rectifier diodes ($D_{o1a}$, $D_{o1b}$, $D_{o2a}$ and $D_{o2b}$) are conducting to provide the output current. If there is no voltage drop ideally in the primary side, the primary current will maintain its value at $i_{pri}(t_4)$ in this mode because the slope of the primary current equals to zero during mode 5. However, since there are several voltage drop, such as the turn-on resistance of switches, the winding resistance of transformers, body diode forward drop, unbalanced voltage of the $C_{o1}$ and $C_{o2}$, the primary current falls linearly to i($t_5$). The primary freewheeling current through S4 and D2 is can be expressed as follows:

$$i_{pri}(t) = -\frac{V_{drop}}{2L_{lkg}}(t-t_4) + i_{pri}(t_4) \tag{7}$$

Mode 5 is completed at time $t_5$ when S4 is turned off.

## III. ANALYSIS AND DESIGN CONSIDERATIONS

For the convenience of the calculation of voltage conversion ratio, several assumptions are made as follows:

- All components are ideal.
- The dead time between S1 and S2 is discarded.
- The effective period $D_{eff}T_s$ of S1 is less than 0.5 $T_s$
- The commutation time between two pairs of output diodes is discarded.

In steady state, the voltage conversion ratio can be obtained from the volt-second balance rule on the magnetizing inductance of the each transformer.

The voltage conversion ratio can be expressed as follows:

$$V_o \cong \frac{2}{N}D_{eff}V_{in} \tag{8}$$

### *A. Selection of Maximum Effective Duty $D_{eff}T_s$ and Turn Ratio of The Transformers, n*

The proposed converter for PSPM specifications has following specifications:

- Input voltage, $V_{in}$, 385 V
- Output Power, $V_o/I_o$, 170V/2.5 A
- Switching Frequency, $f_s$, 100 kHz

To guarantee the safe operation, the dead time between S1 and S2 is determined as $0.035T_S$, the $D_{free}T_S$ + $D_{lkg}T_S$ time as 0.045 $T_S$ in half- cycle duty, and maximum effective duty ($D_{eff}T_S$) as 0.42 $T_S$. The turn ratio of transformers can be calculated as $N = 1.88$ using the eq.(8).

### *B. Selection of Magnetizing Inductance, $L_m$ and Zero-Voltage Switching*

In conventional PSFB or TTFB converter for the PSPM, to achieve the ZVS of lagging leg switches by leakage inductance is difficult due to the low current in primary side of transformers.

The minimum leakage inductor energy required for the ZVS of lagging leg (using common MOSFET rated 500 V, 10 A has 250 pF output capacitor.) can be obtained as follows:

$$E_{ZVS.C} = 2\times\frac{2}{3}(C_{oss})V_{in}^2 = 2\times\frac{2}{3}(250p)\times 385^2 \cong 50\times10^{-6} \tag{9}$$

$$E_{ZVS.L} = \frac{1}{2}(L_{lkg})i_{pri}^2(t_5) \geq E_{ZVS.C} = 2\times\frac{2}{3}(C_{oss})V_{in}^2 = 50\times10^{-6} \tag{10}$$

For the given energy, to increase the ZVS range, the leakage inductance is increased too large or requires high initial current of leakage inductor at $t_5$. It is unreasonable condition.

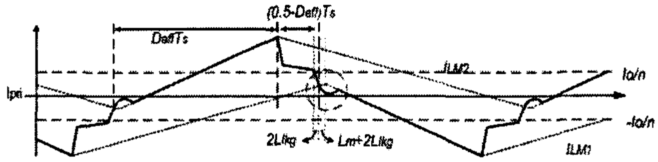



Fig. 6 The magnetizing currents of the two transformer for ZVS

The primary current $i_{pri}$, $i_{LM1}(t)$, and $i_{LM2}(t)$ are shown in Fig. 6. It is noted that since $i_{LM2}(t)$ swings with a positive offset and $i_{LM1}(t)$ swings with a negative offset, each magnetizing current can be considered as a current source, $I_o/n$ in this figure. Thus, the magnetizing inductance of the TTFB contributes to achieve the ZVS condition of the lagging leg switch.

Therefore, the ZVS condition of lagging leg switch is changed as equation (11):

$$\frac{1}{2}(L_{m1} + 2L_{lkg})\, i_{pri}^{\ 2}(t_5) \geq 2\times\frac{2}{3}C_{OSS}V_{in}^{\ 2} \tag{11}$$

To guarantee the safe ZVS operation,

$$E_{ZVS.L} \geq 100\times10^{-6}$$

$$L_m = 206uH,\ L_{lkg1} = 4uH\ and\ E_{ZVS.L} = 167\times10^{-6}$$

### C. Lossless Diode-Clamping Output Configuration

Output rectifier diodes $D_{o1b}$, $D_{o2b}$ can be clamped under a half the output voltage ($V_o/2$) which can be calculated by eq.(3). Therefore, the Schottky rectifiers having a low forward voltage drop can be used. The other output rectifier diodes (Dr, $D_{o1a}$ and $D_{o2a}$ ) are clamped on the output voltage ($V_o$). $D_{o1b}$ and $D_{o2b}$ are high current fast recovery diodes for the powering purpose and $D_r$ is low current (below 1A which current rating is calculated through eq.(2)) fast recovery diodes simply for the clamping purpose.

## IV. EXPERIMENTAL RESULTS

The 425W prototype of the proposed circuit has been constructed for PSPM. The parameters of this prototype circuit (shown in Fig.2) are selected as $n=1.88$, $L_{m1}=L_{m2}=210\mu H$, $L_{lkg1}=L_{lkg2}=4\mu H$, $C_{oss1\sim4}=250pF$, $C_{Do1a}=C_{Do2a}=55pF$, $C_{Do1b}=C_{Do2b}=500pF$, $C_{Dr}=55pF$, $C_{r1}=C_{r2}=6.8nF$, $C_{o1}=C_{o2}=2.2\mu F$, and $C_o=470\mu F/250V$.

The key experimental waveforms of the proposed converter at the full load are shown in Figs. 8, 9. Fig. 8 (a) ~ (d) shows the voltages and currents of lagging leg switches S2 and S4. It can be seen in this figure that the ZVS of both switches are achieved since the currents of the switches flow in the opposite direction before the switches are turned on. The primary and secondary current of each transformer are shown in Fig.8 (e)~(g). Fig. 9 shows the voltage waveforms of rectifier diodes at secondary side of T1, T2 transformers. There are no serious ringing voltages across diodes owing to a lossless diode clamp. In Fig. 10, the range of ZVS for the lagging leg switch S4 is down to about 10% of load current. This is due to the use of the magnetizing current as the ZVS operation.

In Fig.11, the measured efficiencies of the proposed converter, conventional TTFB without snubber, and conventional PSFB without snubber are shown according to the load current. As can be seen in this figure, the maximum efficiency is as high as 95.9 %.

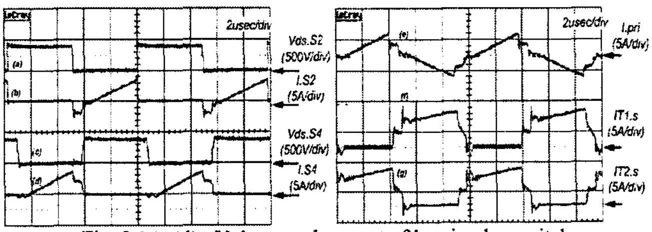



Fig. 8 (a)~(d) : Voltage and current of lagging leg switch
(e) : The primary current of transformer
(f)~(g) : The secondary current of transformer

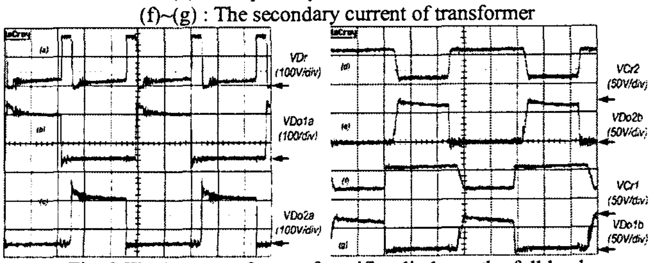



Fig. 9 The key waveforms of rectifier diodes at the full load

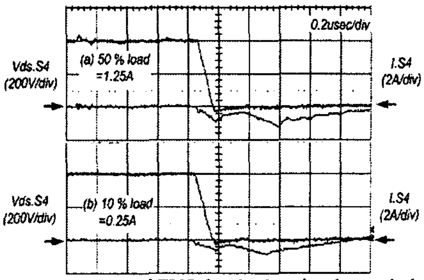



Fig. 10 The range of ZVS for the lagging leg switch S4

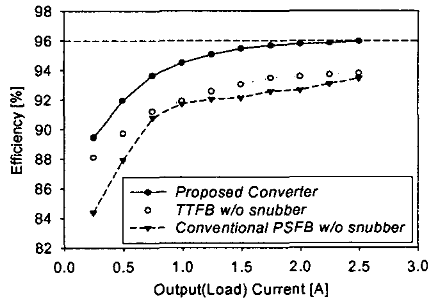



Fig. 11 The measured efficiency of proposed converter

## VII. CONCLUSION

In this paper, a zero-voltage switching two-transformer full bridge PWM converter with lossless diode-clamp rectifier for a plasma display panel sustaining power module has been proposed. The TTFB has two transformers which act as the output inductor as well as main transformer, i.e. as the forward and flyback transformer. Although the magnetizing inductor of the TTFB makes it easier to achieve the ZVS of the lagging leg switch along the wide load range and the new lossless diode-clamp rectifier is employed as the output rectifier, which helps the voltage across rectifier diodes to be clamped on a half the output voltage ($V_o/2$) or the output voltage ($V_o$). Therefore, no dissipated snubber for rectifier diodes is needed and a high efficiency as well as low noise can be realized.

A prototype has been designed to prove the validity of the proposed converter. The experimental results of the prototype converter have been presented for the specifications of 385V input and 170V output. The measured efficiency is as high as above 94% along a wide load range and maximum efficiency comes up to 95.9% at a rated load condition which has higher efficiency than a conventional PSFB and TTFB.

## REFERENCES

[1] J.A.Sabate, V.Vlatkovic, R.B. Ridley, F.C. Lee and B.H.Cho, "Design Considerations for High-Voltage High-Power Full-Bridge Zero-Voltage-Switched PWM Converter", *in Proc. IEEE APEC*, 1990, pp. 275-284

[2] Gwan-Bon Koo, Gun-Woo Moon and Myung-Joong Youn, "Analysis and Design of Phase Shift Full Bridge Converter With Series-Connected Two Transformers", *IEEE Trans. on Power Electronics*, Vol.19, No.2, 2004