A 5 kW Boost Converter with Various Passive/Active Snubbers for Reducing Component Stress and Achieving High Efficiency

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Abstract—This paper presents a 5 kW boost converter with various passive/active snubbers for reducing current and voltage stress, and achieving high efficiency. The proposed converter includes a clamp branch diode and capacitor, a coupling inductor, and a PWM converter to function as active clamp. As compared to a conventional boost converter with a passive inductor-capacitor-diode (LCD) snubber, the proposed has the merits of low current stress imposed on the main switch, clamped switch voltage and limited reverse recovery current. In this study, three types of active snubbers, buck, boost and flyback, are respectively implemented with a boost converter. Theoretical analysis and experimental results have verified that the proposed circuit configuration is attractive for high power applications.

Index Terms—boost converter, flyback active snubber, coupling inductor, inductor-capacitor-diode snubber, reverse recovery current.

I. INTRODUCTION

Renewable energy resources have drawn a lot of attention. Photovoltaic (PV) energy is most popular as it is clean, maintenance-free, and abundant. In order to obtain maximum power from PV modules, tracking the maximum power point (MPP) of PV arrays is usually an essential part of a PV system, which is mostly realized with a boost converter [1]-[6].

For high power applications, component stress, switching loss and EMI noise are increased due to high \( \frac{di}{dt} \) of diode reverse-recovery current and high \( \frac{dv}{dt} \) of MOSFET drain-source voltage, resulting in low reliability and even violating regulation. Hence, passive and active clamp snubbers were applied to the boost converter [7]-[10]. In [11], four soft-switching boost converter topologies with passive snubbers have been built for high power applications. However, when the circuit operates under heavy load condition, the overall conversion efficiency is still suffered. To overcome the problem, a coupled winding from the boost inductor is inserted in series with the auxiliary diode \( D_s \), as shown in Fig. 1. Although the inductor-capacitor-diode snubber can limit reverse recovery current and clamp switch voltage, it still suffers from high switch current stress, and might also result in high switching loss and high conduction loss.

To release the above mentioned drawbacks, the passive snubber can be replaced by an active clamp one [12]-[14]. Among the proposed active clamp snubbers, the buck active clamp is first adopted with the boost converter, as shown in Fig. 2(a). It can clamp the switch voltage while inductor current \( i_{LS} \) tracking inductor current \( i_{Lm} \) at the main switch turn-off transition. However, clamping capacitor voltage \( v_{CS} \) cannot be drained out during the main switch turn-on state, resulting in significant turn-off loss. Thus, a boost converter with a boost active clamp snubber, as shown in Fig. 2(b), is introduced to relieve the above drawback. With a boost active clamp snubber, the energy stored in \( C_s \) can be effectively transferred to the output in every switching cycle and a ZVT feature can be attained at the main switch turn-off transition. Although the circuit has the merit of a ZVT feature, it still suffers from high voltage spike imposed on the main switch \( S_m \) at the switch turn-off transition.
To solve the above mentioned problem, a boost converter with a flyback active clamp snubber, as shown in Fig. 2(c), is therefore proposed. It can relieve the drawbacks of high current and high voltage stresses imposed on the main switch at turn-on and off transitions. The active snubber only deals with around 1% of the full load power, and it is operated in DCM to avoid high voltage or current spike.

Mode 2 [Fig. 4(b), \(t_1 < t_2\)]: When equation (1) is satisfied, current \(i_{LS}\) will start to track current \(i_{Lm}\) with a resonant manner, and capacitor \(C_S\) will start to release energy. At time \(t_2\), current \(i_{LS}\) is equal to current \(i_{Lm}\). Meanwhile, the voltage of main switch \(S_m\) and capacitor \(C_S\) will reach the maximum value simultaneously, and an equivalent circuit is shown in Fig. 4(b). Soft switching zero-voltage-transition (ZVT) feature is therefore attained during \(t_0\) to \(t_2\). In the mode, snubber capacitor \(C_S\), equivalent inductance \(L_X (= L_{K2} + L_S)\) and buffer capacitor \(C_b\) are in resonance. Currents \(i_{LS}(t)\) and \(i_{CS}(t)\), and voltages \(v_{LS}(t)\), \(v_{Cb}(t)\) and \(v_{CS}(t)\) can be derived as

\[
\begin{align*}
i_{LS}(t) &= \frac{C_y}{C_S} I_{Lm} \left[ 1 - \cos \omega_{oi}(t-t_1) \right], \\
i_{CS}(t) &= I_{Lm} \left( \frac{C_x}{C_S} \right) I_{Lm} \left[ 1 - \cos \omega_{oi}(t-t_1) \right], \\
v_{LS}(t) &= (Z_{oi} I_{Lm} \left( \frac{C_x}{C_S} \right) \sin \omega_{oi}(t-t_1)), \\
v_{CS}(t) &= \frac{I_{Lm}}{C_s + C_b} \frac{1}{\omega_{oi}} \sin \omega_{oi}(t-t_1) - (t-t_1) + V_{C3(t1)} .
\end{align*}
\]

and

\[
v_{CS}(t) = \frac{1}{C_x} [ I_{Lm}(t-t_1) \times (1 - \frac{C_x}{C_S} \frac{I_{Lm}}{C_s} \sin \omega_{oi}(t-t_1)] + V_{CS(t1)} ,
\]

where \(V_{C3(t1)}\) and \(V_{CS(t1)}\) are the initial value of capacitors \(C_b\) and \(C_S\) at \(t_1\), \(I_{Lm}\) is a constant value, and capacitor \(C_S\), resonant frequency \(\omega_{oi}\) and characteristic impedance \(Z_{oi}\) are respectively expressed as follows:

\[
C_x = \frac{C_b C_S}{C_s + C_b} ,
\]

\[
\omega_{oi} = \sqrt{\frac{L_x C_x}{}},
\]

\[
Z_{oi} = \frac{L_x}{\sqrt{C_x} } ,
\]

and

\[
L_x = L_S + L_{K2} .
\]

Mode 3 [Fig. 4(c), \(t_2 < t_3\)]: Before \(t_3\), the energy stored in buffer capacitor \(C_b\) was not completely drained out; thus, the capacitor will not stop discharging until its voltage drops to zero. The equivalent circuit is shown in Fig. 4(c). The energy stored in capacitor \(C_S\) is

\[
W_{CS} = \frac{1}{2} C_S \cdot V_{CS}^2(t_2) .
\]

Mode 4 [Fig. 4(d), \(t_3 < t_4\)]: When the energy stored in \(C_b\) has been completely released to the output at \(t_3\), diode \(D_m\) will conduct. In this interval, the voltage across the main switch will drop back to output voltage \(V_o\), and moreover, the circuit operation in this mode is identical to a regular turn-off...
state of a conventional boost converter. The equivalent circuit is shown in Fig. 4(d). It should be noted that voltage $V_{di}$ might not be higher than $V_o$ under light load condition and with a small $L_{K2}$.

**Mode 5** [Fig. 4(e), $t_4 < t_d$]: The driving signals of both boost converter and flyback snubber are synchronously started at $t_d$. In this mode, the boost converter achieves a zero-current-transition (ZCT) soft switching feature, and current $i_{CS}$ drops to zero gradually.

In the flyback snubber, the energy stored in capacitor $C_S$ will be delivered to magnetizing inductance $L_{mf}$, current $i_{mf}$ is therefore built up, and the equivalent circuit is shown in Fig. 4(e). During the energy-transfer process, both components $C_S$ and $L_{mf}$ are in resonance. Currents $i_{CS}(t)$, $i_{mf}(t)$ and $i_{df}(t)$ are identical; thus, current $i_{CS}(t)$ and voltage $v_{CS}(t)$ can be derived as

$$i_{CS}(t) = \frac{V_{mf}}{Z_{O4}} \sin \omega_{O4}(t-t_4),$$

and

$$v_{CS}(t) = -V_{mf} \cos \omega_{O4}(t-t_4),$$

where the resonant frequency $\omega_{O4}$ and the characteristic impedance $Z_{O4}$ are respectively expressed as follows:

$$\omega_{O4} = \frac{1}{\sqrt{L_{mf}C_S}},$$

and

$$Z_{O4} = \frac{L_{mf}}{\sqrt{C_S}}.$$

**Mode 6** [Fig. 4(f), $t_5 < t_d$]: Afterwards, boost diode $D_m$ is in reverse bias, and the equivalent circuit is shown in Fig. 4(f). The $di/dt$ of the boost diode reverse-recovery current is primarily limited by leakage inductance $L_{K2}$.

**Mode 7** [Fig. 4(g), $t_6 < t_d$]: In this mode, boost converter and flyback snubber are also maintained in the on state. The energy from capacitor $C_S$ is still delivered to magnetizing inductance $L_{mf}$. The equivalent circuit is shown in Fig. 4(g).

**Mode 8** [Fig. 4(h), $t_7 < t_d$]: At $t_7$, capacitor voltage $v_{CS}$ drops to zero. In this mode, the time of driving signal $V_{gdf}$ is slightly greater than the discharging time of capacitor $C_S$. The purpose is to ensure that the energy in capacitor $C_S$ can be completely released, creating a ZVT operational opportunity at next turn-off transition. The equivalent circuit is shown in Fig. 4(h).

**Mode 9** [Fig. 4(i), $t_8 < t_d$]: When switch $S_f$ is turned off at $t_8$, the energy stored in inductance $L_{mf}$ starts to transfer to buffer capacitor $C_b$ by way of $D_f$, and the equivalent circuit is shown in Fig. 4(i). During this interval, both magnetizing inductance $L_{mf}$ and buffer capacitor $C_b$ are in resonant manner; as a result, current $i_{CB}(t)$ and voltage $v_{CB}(t)$ can be derived as

$$i_{CB}(t) = I_{mf(\omega_8)} \cos \omega_{O8}(t-t_8) + \frac{V_{mf}}{Z_{O8}} \sin \omega_{O8}(t-t_8),$$

and

$$v_{CB}(t) = Z_{O8}I_{mf(\omega_8)} \sin \omega_{O8}(t-t_8) - V_{mf} \cos \omega_{O8}(t-t_8),$$

where $I_{mf(\omega_8)}$ is the initial value of magnetizing inductance at $t_8$, and resonant frequency $\omega_{O8}$ and characteristic impedance $Z_{O8}$ can be determined as

$$\omega_{O8} = \frac{1}{\sqrt{L_{mf}C_b}},$$

and

$$Z_{O8} = \frac{L_{mf}}{\sqrt{C_b}}.$$

**Mode 10** [Fig. 4(j), $t_9 < t_d$]: Because the energy stored in magnetizing inductance $L_{mf}$ was completely transferred to capacitor $C_b$ at $t_9$, currents $i_{df(t)}$, $i_{mf(t)}$ and $i_{df(t)}$ and voltage $v_{df(t)}$ are equal to zero in this interval. Voltage $v_{CB}$ is clamped till next switching cycle, time $t_9$. The equivalent circuit is shown in Fig. 4(j). A complete switching cycle ends at $t_9$.
III. DESIGN OF THE CONVERTER

This section will cover the design of the power converter and selection of the major components. The proposed converter can be divided into a boost converter power stage and a flyback active clamp snubber. A detailed design procedure is described as follows:

A. Design of the boost converter [15]:

1) Main switch (Sm)

To operate the converter at a 5 kW power rating, the main switch is realized with two MOSFET devices in parallel connection. In the experiment, two IXYS model IXFH36N50P with a low $R_{d(on)}$ value (0.17 $\Omega$) were adopted for the proposed boost converter.

2) Main inductor (Lm)

The inductor was designed based on (20), which can be operated at continuous-conduction mode (CCM):

$$L_m > L_g = \frac{V_o T_s}{2 I_{OB}} D(1-D)^2,$$

where $L_g$ is the boundary inductance, $T_s$ is the switching period, $I_{OB}$ is the boundary output current and $D$ is the duty cycle.

In addition, core loss, saturation flux density, and frequency response of the inductor are also needed to be considered. Hence, a toroidal core CM777125 is selected for the main inductor. The winding of 2 paralleled 13 AWG copper wires with 33 turns was designed.

B. Design of the flyback active clamp snubber:

The purpose of using a flyback active clamp snubber is to transfer energy from snubber capacitor $C_b$ to buffer capacitor $C_b$, which can attain a ZVT soft switching feature for main switch $S_m$. The key components of $L_S$, $C_S$ and $C_b$ are designed in the following.

Design of snubber inductor $L_S$ and capacitor set $C_S$ and $C_b$ can be achieved with Matlab software package. The simulation is conducted with the following conditions:

1) both ZCT and ZVT soft switching features are still sustained,
2) the processed power is less than 1% of the full load condition, and
3) the capacitor-set ratio($C_b/C_S$) is around 2.

In addition to the above conditions, two circuit modes operate alternately. In mode 1, current $i_{lm}$ flows through the low impedance-path capacitor $C_S$. Relationship among $v_{CS}$, $V_O$ and $V_{CB}$ can be expressed as follows:

$$v_{CS} < V_O - V_{CB}(t_1),$$

$$i_{LS} = 0, \quad i_{CS} = i_{lm}, \quad \text{and} \quad C_S \frac{dv_{CS}}{dt} = i_{CS}.$$

When capacitor $C_S$ is charged to be high enough, it means that equation (1) is satisfied, and the converter enters mode 2 operation. Current $i_{lm}$ will flow through the path of $L_S - L_{Si} - D_2 - C_O$ with a resonant manner, which creates a ZVT operational opportunity for main switch $S_m$. The following relationship can be obtained:

$$v_{CS} \geq V_O - V_{CB}(t_1),$$

where

$$C_S \frac{dv_{CS}}{dt} = i_{CS}, \quad L_S \frac{di_{LS}}{dt} = v_{CS} - (V_O - v_{CB}), \quad C_b \frac{dv_{CB}}{dt} = i_{CB},$$

and $i_{CS} = i_{lm} - i_{LS}$.

Based on the above conditions, snubber inductance $L_S$, processed power of the flyback snubber, capacitor set $C_S$ and $C_b$, and voltage $v_{CS}$ and $v_{CB}$ can be derived. Simulated results from the above expressions are shown in Figs. 5 to 7. According to these simulated results, higher snubber inductance $L_S$ can reduce diode reverse-recovery loss, while the flyback snubber needs to process higher power and higher voltage will cross the snubber capacitor, resulting in lower conversion efficiency. To choose proper component values, the major parameter consideration is snubber inductance $L_S$. This component value can not only alleviate diode reverse-recovery loss, but can determine the power level of the flyback snubber.

![Fig. 4. Various circuit modes illustrating the operation of the boost converter with a flyback active clamp snubber.](image)

![Fig. 5. Snubber inductance $L_S$ versus processed power of the flyback snubber under various snubber capacitor sets. (Output power = 5 kW)](image)

![Fig. 6. Snubber inductance $L_S$ versus voltage stress of $C_S$ under various snubber capacitor sets. (Output power = 5 kW)](image)
Given a processed power level of the flyback snubber, inductance $L_S$, capacitance $C_S$ and $C_b$ can be determined from the plots of the simulated results shown in Fig. 5. Coupled inductor $L_S$ and its leakage inductance are used to limit the reverse recovery current of diode $D_{on}$. It is first chosen as $L_S = 2.5 \mu H$ to limit the current effectively. Next, in considering the voltage stresses of the capacitor set, the middle curve in Fig. 5 is selected, and $C_S$ and $C_b$ will be 22 nF and 47 nF, respectively. According to the principle of energy conversation, the relationship between capacitors $V_{Cb}$ and $V_{CS}$ can be expressed as

$$V_{cb} = m \cdot V_{CS},$$

where

$$m = \frac{C_S}{\sqrt{C_b}}. \hspace{2cm} (22)$$

Voltage $V_{CS}$ is always higher than or equal to $V_{O}$; thus, capacitance $C_b$ is usually required to be greater than capacitance $C_S$ to buffer a voltage lower than or equal to $V_O$. Another consideration is the voltage stress imposed on switch $S_m$. Hence, capacitor $C_b$ should be somehow larger than $C_S$ to yield a voltage lower than $V_O$. In the design, capacitance $C_b$ is chosen as $47 \text{nF}/400V$.

IV. EXPERIMENTAL RESULTS

To verify the converter performance, five prototypes of 5 kW, the conventional boost converter, the boost converter with a passive snubber, the one with a buck active clamp snubber, the one with a boost active clamp snubber and the one with a flyback active clamp snubber, were designed and built. The specifications are listed as follows:

- input voltage $V_i$: 200 V$_{dc}$
- input current $I_i$: 25 A
- switching frequency $f_s$: 20 kHz
- output voltage $V_O$: 360 V$_{dc}$ and
- output power: $P_{O(max)} = 5$ kW

Fig. 8 shows voltage $v_{ib}$ and current $i_{ib}$ waveforms at turn-on and off transitions of main switch $S_m$ under a 3 kW power rating, illustrating that hard-switching manner occurs in the conventional boost converter. When the power rating goes higher than 3 kW, the conventional boost converter does not work properly. Fig. 9 shows those of $S_m$ at turn-on and off transitions of the boost converter with a passive snubber. It can be seen from Fig. 9 (a) that high resonant current circulates through the main switch. Fig. 9 (b) illustrates a ZVT feature at turn-off transition. Fig. 10 shows the measured main switch $S_m$ of the boost converter with a buck snubber at turn-on and off transitions. Fig. 10(a) illustrates that reverse-recovery current is alleviated by snubber inductor $L_S$ and $L_{k2}$, and meanwhile, the circulation current problem can be relieved. However, hard switching still occur at turn-off transition because the energy stored in $C_S$ cannot be completely released to the output, as shown in Fig. 10(b). Measured results from the boost converter with a boost snubber are shown in Fig. 11, in which Fig. 11(a) illustrates that the boost converter combined with a boost active clamp snubber can also avoid serious current spike at turn-on transition, and Fig. 11 (b) illustrates a ZVT feature at turn-off transition. Fig. 12 shows measured current and voltage waveforms of main switch $S_m$ of the boost converter with a flyback snubber at turn-on and off transitions. From Fig. 12 (a), it can be seen that reverse-recovery current can be well limited. In Fig. 12(b), it can be observed that high voltage stress can be further suppressed as compared to that of the converter with either a buck or a boost active clamp snubber.
The proposed converter can achieve higher efficiency compared to the other four types of converters, specifically the boost converter with a flyback snubber. As voltage stress and ZVT feature have been also achieved by the ZCT feature of the main switch have been attained, and low current stress and ZVT feature have been also achieved by the proposed boost converter with a flyback snubber. Experimental results have shown that the proposed converter with a flyback snubber and its counterparts under 5 kW have been implemented in the laboratory to verify its feasibility. Theoretical analysis has been also described in detail.

Fig. 11. Measured voltage $v_{ds}$ and current $i_{ds}$ waveforms of main switch $S_m$ at (a) turn-on and (b) turn-off transitions from the boost converter with a boost snubber and with a 5 kW power rating.

Fig. 12. Measured voltage $v_{ds}$ and current $i_{ds}$ waveforms of main switch $S_m$ at (a) turn-on and (b) turn-off transitions from the boost converter with a flyback snubber and with a 5 kW power rating.

Fig. 13 shows plots of the efficiency versus output power from 1 kW to 5 kW under various passive/active snubbers. From these plots, we can see that the efficiency of the converter with an active or a passive snubber comes out almost the same, while they have different current and voltage stresses imposed on the main switch. Even though the proposed converter with a flyback snubber requires a more sophisticated control than that with a passive snubber, the proposed can achieve a lower current stress at switch turn-on transition, reducing EMI level and circulation current significantly.

V. CONCLUSIONS

In this paper, the proposed boost converter with a flyback snubber and its counterparts under 5 kW have been implemented in the laboratory to verify its feasibility. Theoretical analysis has been also described in detail. Experimental results have shown that low current stress and ZCT feature of the main switch have been attained, and low voltage stress and ZVT feature have been also achieved by the proposed boost converter with a flyback snubber. As compared with the other four types of converters, the proposed converter can achieve the highest efficiency, while sustains the lowest current and voltage stresses. It is relatively suitable for high power applications.

VI. REFERENCES


