SIMULATION OF HIGH STEP UP DC-DC CONVERTER FOR GRID TIE THREE PHASE INVERTER USING RENEWABLE ENERGY

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ABSTRACT: This paper presents a high efficiency step-up DC-DC converter for low-DC renewable energy sources. The proposed DC-DC converter is controlled by asymmetrical pulse width modulation (APWM) technique and achieves Zero-current switching (ZCS) of all output diodes. Compared to the previous DC-DC converters, the voltage stresses of the semiconductor devices can be reduced in the proposed converter. Simulation results are obtained from a 200 W output power. The proposed DC-DC converter achieves a high efficiency of 97.5% at the rated load. Further project has been extended to three phase grid tie inverter using sinusoidal pulse width modulation technique.

Keywords: DC-DC converter, soft-switching, high voltage gain, zero-current switching (ZCS), Three Phase Inverter.

I. INTRODUCTION

The power generation systems using low-DC renewable energy sources such as photovoltaic module and fuel cell need a high step-up DC-DC converter to interface the low-DC voltage to the high DC voltage distribution network [1]. Lots of efforts have been made to develop high step-up DC-DC converters with a high efficiency [2]-[6]. Among the investigated topologies, the active-clamped step-up DC-DC converters in [4]-[6] are gaining its popularity, thanks to its high step-up ratio and soft-switching operation. The active-clamped step-up DC-DC converters in [4]-[6] has a high step-up gain by using the active-clamp circuit at the primary side and the voltage doubler rectifier at the secondary side. Moreover, the series-resonance between the transformer leakage inductor and the capacitor in the voltage doubler rectifier makes the output diodes to be turned off at zero current condition [5]. However, the power switches at the primary side operate under hard-switching condition which still causes high switching power losses and high heat dissipation problems. The half-bridge dc-deconverter has been presented to reduce switching power losses at high-voltage side [1]. The output diodes are turned off at zero current by using the voltage doubler rectifier. However, an additional half-wave rectifier is needed, which increases switching power losses. In order to overcome these problems, this paper proposes a high-simulation efficiency dc-dc converter for low-DC renewable sources. An improved active-clamped dc-deconverter is presented by using a dual active-clamping circuit. The voltage stress of power switches can be reduced at primary side. The performance of the proposed converter is verified using a 200 W simulation prototype. The simulation results confirm that a high efficiency of 97.5% is achieved at 50 V input voltage for 200 W output power with an improved dynamic performance.

In this paper a new DC-DC converter topology is proposed to step up the fuel cell voltage and provide a stable dc-link for the DC-AC inverter. The proposed DC-DC power conversion unit consists of a two parallel connected buck-boost converters. This produces an independently controllable dual voltage output. Block diagram of proposed converter is shown in Fig. 1. As it will be shown the use of the proposed topology along with a DC-AC inverter eliminates the need for a transformer to provide the required voltage gain. As a result, proposed topology has the following advantages:

- Operates from a single input voltage
- No transformer is required to achieve a voltage gain of 5 in per unit
- If a single phase is connected at its output the system can generate 230VAC output from a 50VDC input source without the use of a transformer.
II. INVERTER CONTROL SCHEME

The Fig. 1 shows the circuit diagram of the proposed stepup DC-DC converter. This converter combines an input capacitor $C_i$, a clamp capacitor $C_c$, main switches $S_1$ and $S_4$, an auxiliary switches $S_2$ and $S_3$, the resonant capacitor $C_r$, two output diode $D_{o1}$ and $D_{o2}$, an output capacitor $C_o$, and the secondary side leakage inductor $L_{lk}$. Each switch has its own parasitic capacitor $C_{S1} \sim C_{S4}$ and body diode $D_{S1} \sim D_{S4}$. The switches operate at a constant switching period $T_s(=1/f_s)$ with the asymmetrical pulse-width modulation (APWM). The transformer $T$ has the magnetizing inductor $L_m$ and leakage inductor $L_{lk}$ with the turns ratio of $1:N$ where $N = N_s/N_p$. The proposed converter operates in a continuous conduction mode so that the magnetizing inductor current $i_{Lm}$ flows continuously. The capacitors $C_c, C_o$ are large enough so that their voltages are considered constant as $V_c$ and $V_o$. $D_i$ is the duty ratio based on $S_1$ ($S_4$) turn-on time. Fig. 2(a) shows the operation waveforms of the primary side components. Fig. 2(b) shows the operation waveforms of the secondary side components. Fig. 3 shows the operation modes of the proposed converter for $T_s$. The proposed converter has six distinct operating modes for $T_s$ as follows:

Mode 1 [$t_0, t_1$]: At $t = t_0$, $S_1$ and $S_4$ are turned on. Since $V_{Lm} = V_i$, the magnetizing inductor current $i_{Lm}$ increases linearly as

![Fig.1: Proposed system configuration](image-url)
\[ lM(t) = lM(t_0) + \frac{V_i}{Lm} \cdot (t - t_0) \] (1)

From \( V = L \frac{di}{dt} \), \( di = \frac{V}{L} \cdot dt \) (2)

When \( nV_i \) is applied to the secondary winding of \( N_s \), the diodes \( Do1 \) is turned on. The series-resonant circuit consisting of \( Lk \) and \( Cr \) is formed, respectively. By the series resonance between \( Lk \) and \( Cr \), the energy stored in the capacitor \( Cr \) is transferred to the output capacitors \( Co \). The angular resonant frequency \( \omega_r \) of this series-resonant circuit is

\[ \omega_r = 2\pi f_r = \frac{1}{\sqrt{LkCr}} \]

where \( f_r \) is the resonant frequency. By referring the output diode current \( iDo1 \) to the primary side, the primary current \( ip \) is expressed as

\[ ip(t) = lM(t) \]

\[ ip(t) = lM(t) + niDo1(t) \]

\[ ip(t) = ip(t_0) + \frac{V_i}{Lm} \cdot (t - t_0) + niDo1(t) \] (3)

\[ iDo1(t) = \frac{V_0 - nV_i - Vr}{Zr} \cdot \sin \omega_r(t - t_0) \] (4)

Where \( Zr = \sqrt{Lk/\omega_r^2} \) (5)
Mode 2 \([t_1, t_2]\): At \(t = t_1\), the half-resonant period for the output diode currents \(i_{Do1}\) is finished. The output diode currents \(i_{Do1}\) is zero before \(Do1\) is turned off. Zero current switching of \(Do1\) is achieved without any diode reverse recovery problem at the end of Mode 2.

Mode 3 \([t_2, t_3]\): At \(t = t_2\), \(S1\) and \(S4\) are turned off. The primary current \(i_{p}\) charges \(CS1\) and \(CS4\), discharges \(CS2\) and \(CS3\). The voltage \(VS1\) across \(S1\) increases from zero to the voltage \(V_i\) and The voltage \(VS4\) across \(S4\) increases from zero to \(V_c\). Since the capacitor \(CS\) (= \(CS1 = CS4\)) is very small, the time interval during this mode is considered negligible compared to \(T_s\).

Mode 4 \([t_3, t_4]\): At \(t = t_3\), the auxiliary switches \(S2\) and \(S3\) are turned on with zero voltage. Thus, switching loss is reduced. Since \(V_{Lm} = -V_c\), the magnetizing inductor current \(i_{Lm}\) decreases linearly as

\[
i_{Lm}(t) = i_{Lm}(t_3) - \frac{V_c}{L_m} (t - t_3) \quad (6)
\]
When \(-nVc\) is applied to the secondary winding of \(N_s\), the output diode \(D_{o2}\) is turned on. The series-resonant circuit consisting of \(L_{lk}\) and \(C_r\) is formed, respectively. Through the output diode \(D_{o2}\), the energy is transferred to the resonant capacitor \(C_r\). By referring the output diode current \(i_{D_{o2}}\) to the primary side, the primary current \(i_{p}\) is expressed as

\[
i(t) = ip(t_3) - \frac{Vc}{L_m} \ast (t - t3) - niD_{o2}(t)\]  

(7)

where the output diode current \(i_{D_{o2}}\) is given by

\[
i_{D_{o2}}(t) = \frac{nV_{c}+V_{r}}{2\pi \omega_{r}} \sin \omega_{r}(t - t_{0})\]  

(8)

**Mode 5 \([t_4, t_5]\):** At \(t = t_4\), the half-resonant period of the output diode currents \(i_{D_{o2}}\) is finished. The output diode current \(i_{D_{o2}}\) is zero before \(D_{o2}\) is turned off. Zero-current switching of \(D_{o2}\) is achieved without any diode reverse recovery problem at the end of **Mode 5**.

**Mode 6 \([t_5, t_6]\):** At \(t = t_5\), \(S_2\) and \(S_3\) are turned off. The primary current \(i_{p}\) charges \(C_{s2}\) and \(C_{s3}\), discharges \(C_{s1}\) and \(C_{s4}\). The voltage \(V_{S2}\) across \(S_1\) increases from zero to the voltage \(V_{c}\) and The voltage \(V_{S3}\) across \(S_3\) increases from zero to \(V_{i}\). Since the capacitor \(C_s (= C_{s2} = C_{s3})\) is very small, the time interval during this mode is considered negligible compared to \(T_s\). The next switching cycle begins when \(S_1\) is turned on again. By imposing the voltage-second valence role on the magnetizing inductor \(L_m\), the capacitor voltages \(V_c\) and \(V_{r}\) can be expressed as

\[
V_c = \frac{D}{1-D} \frac{V_i}{n^{2}L_m - L_m} \frac{D}{1-D} \frac{V_i}{n^{2}L_m - L_m}\]  

(9)

\[
V_{r} = \frac{D}{1-D} nV_i\]  

(10)

where \(D\) is the duty cycle of the main switches \(S_1\) and \(S_4\). For \(L_{lk} << L_m\), \(9\) can be rewritten as for voltage-second rule on the secondary winding, the following equations can be obtained.

\[
V_{r} = \frac{D}{1-D} nV_i\]  

(11)

\[
(V_o - V_{r})DT_s = nV_c - (1D)T_s\]  

(12)

From \(9, 10, 11, 12\) we have

\[
V_o = \frac{n}{1-D} V_i\]  

(13)
The maximum voltage stress of $S_1$ and $S_3$ is confined to the input voltage $V_i$. The voltage stress of $S_2$ and $S_4$ is confined to the clamping capacitor voltage $V_c$. Fig. 4 shows the relation between the clamping capacitor voltage $V_c$ and the duty ratio $D$. The dual active-clamping circuit is used in the proposed converter. The clamping capacitor voltage in case of the dual active-clamping circuit is always lower than the clamping capacitor voltage in case of the conventional active-clamping circuit. It means that the switch voltage stress of the proposed converter is always lower than the switch voltage stress of the previous converter [4] using the conventional active-clamping circuit. Especially, when the duty ratio is below 0.5, the clamping capacitor voltage can be lower than the input voltage $V_i$. It is critically beneficial in low-voltage PV applications where more than 50% of the power losses are lost as switching power losses. The output diode currents $i_{Do1}$ and $i_{Do2}$ should be zero.

### III. SIMULATION RESULTS

Fig. 4 shows the Simulink model of the proposed converter and Fig. 5, 6 shows the primary current $i_p$ and output voltage of DC-DC converter, Fig. switch voltages $V_{S1}$ and $V_{S3}$ for 200 W output power. Fig. 5 shows the simulation waveforms at 50 V input voltage. As shown in Figs. 5, $V_{S1}$ and $V_{S3}$ are clamped at the input voltage. Fig. 6 shows the primary current $i_p$, clamping capacitor voltage $V_c$, and switch voltages $V_{S2}$ and $V_{S4}$ for 200 W output power. Fig. 6 shows the simulation waveforms at 50 V input voltage. The clamping capacitor voltage $V_c$ is 48 V. $V_{S2}$ and $V_{S4}$ are clamped at 48 V. Fig. 8 shows the measured power efficiencies of the converters at 50 V input voltage for different output load conditions. The proposed converter achieves the efficiency of 97.5% for 200 W output power. The previous active-clamped converter achieves the efficiency of 97.2% for 200 W output power. The efficiency of 0.3% is improved by the proposed converter at 50 V input voltage for 200 W output power. The previous half-bridge converter achieves the efficiency of 97.0% for 200 W output power. The proposed converter achieves the highest efficiency for the rated output power. Switching power losses are reduced by decreasing the voltage stress of power switches in the proposed converter. The power efficiency is increased by reducing switching power losses.
Fig. 7 shows the simulation waveforms for 200 W output power. The diode current flows in a resonant manner by the series resonance between the leakage inductor and the resonant capacitor. The diode current is zero before the output diode is turned off. Zero-current switching of each output diode is achieved at its turn-off instance.
Fig. 8 shows the output diode voltages $V_{Do1}$ and $V_{Do2}$ and output diode currents $i_{Do1}$ and $i_{Do2}$ for different output load conditions. The switching power loss caused by the diode reverse recovery current can be removed by zero-current turn-off of the output diode.

Fig. 9: Voltage Across Switch VS1

Fig. 10: Wave form of voltage across diode
Fig. 11 shows the output voltage of inverter and output voltage without filter for different output load conditions. The switching power loss caused by the switching of three phase inverter causes the output voltage to be stepped waveform.

Fig. 12 shows the output voltage of inverter and output voltage with filter for different output load conditions. The switching power loss caused by the switching of three phase inverter causes the output voltage to be stepped waveform without filter and filtering causes the output voltage to be pure sinusoidal which is directly fed to the grid.

IV. CONCLUSION

This paper presents a high efficiency step-up DC-DC converter for low-DC renewable energy sources. The operation of the proposed converter has been described. The simulation results have been presented at input voltage $V_i = 50$ V. The proposed converter reduces the switching power losses, increasing power efficiency. By using series resonant circuit, the proposed converter achieves ZCS turn-off to remove the reverse-recovery problem on output diodes. Simulation results have shown that the proposed converter achieves a high efficiency of 97.5%. By feeding DC output voltage of converter to three phase grid tie inverter gives pure sinusoidal voltage with 50 Hz frequency and suitable output AC voltage.

REFERENCES


BIOGRAPHY

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