High Efficiency modified Dual-Active Bridge DC-DC Converter for Photovoltaic Integration

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Abstract—The penetration of distributed renewable energy resources especially photovoltaic (PV) panels is expected to increase over the next few decades. We propose a novel switching scheme for a modified dual-active bridge dc-dc converter for integration of PV panels. The converter uses the leakage inductance of a high frequency transformer for its boost operation and enables power transfer from a low voltage port to a high voltage port. The converter operates in discontinuous conduction mode (DCM), simplifying control and achieving high efficiency results through soft switching. Analysis of converter along with simulation and experimental results are provided to validate the working of proposed converter design.

Index Terms—dc-dc converter, dual-active bridge, full-bridge, photovoltaic panels

I. INTRODUCTION

Growing concern for energy security and climate change has boosted interest in distributed renewable energy generation. As a result energy policies by governments are promoting energy resources like photovoltaic (PV) and wind energy, in an effort to cope with demand challenges and to move a step closer towards clean energy sources [1].

The integration of PV panels presents a power electronics challenge to develop a grid-parallel inverter. Multiple inverter system architectures exist, of which dc-dc-ac topology is one of the widely considered [2]. The dc-dc-ac topology, shown in Figure 1, offers high efficiency, good boost capability, is lightweight and offers scalability for multiple source integration. In this topology, the dc voltage from the PV panel is first boosted up by a dc-dc converter then a dc-ac inverter is cascaded to produce ac voltage. This paper focuses on the task of dc-dc converter for such applications.

The dc-dc converter, for such applications, needs to satisfy the following requirements: high step-up conversion ratio, stable output voltage during load and input voltage variation, and high efficiency. The first requirement excludes most transformerless converter topologies. The required gain in these topologies is achieved at high duty cycle which is not acceptable [3]. The second requirement can be achieved using a good feedback control. High efficiency can be achieved by optimizing the following factors [4]:

1. Zero voltage switching (ZVS) or Zero current switching (ZCS)
2. Low input current ripple
3. Proper utilization of the B-H curve and leakage inductance of transformer

A review of previously proposed converter topologies can be found in [2]. Dual-active bridge (DAB) dc-dc converter, originally proposed by De Doncker and Divan [5], is known for its high energy density, bidirectional power flow, galvanic isolation and high efficiency. Typical switching schemes include phase shift modulation, pulse-width modulation or a combination of these two. Due to the presence of transformer leakage inductance, these converters generally use zero voltage switching to minimize switching losses. An analysis of DAB converter operation using different switching schemes is covered in [6]. The bidirectional power flow capability of DAB is not required in PV panel integration so some controllable active switches can be reduced for such applications.

Full-bridge converter, originally proposed by Hitchcock [7], has been studied comprehensively in literature. Full-bridge converter with rectifier on the secondary side of transformer is also used for dc-dc conversion. The converter operation and analysis in continuous conduction mode (CCM), boundary conduction mode (BCM) and discontinuous conduction mode (DCM) can be found in [8]. The cost of this converter is low due to fewer controllable active elements but it does not provide bidirectional power flow and diverse switching schemes. Additionally load variation significantly affects the gain capability of full-bridge converter in DCM mode.

We propose a modification of dual-active and full bridge dc-dc converter topologies for PV panel integration. The converter design is shown in figure 2. The design uses a semi-active bridge on secondary side for rectification and fast current commutation. It utilizes the leakage inductance of a high frequency transformer for its boost operation. This allows a simple control strategy and soft switching of gates. Simulation and experimental results are presented to validate the claims and demonstrate working of the proposed design.
The paper is divided as follows: Section II provides operation and analysis of the proposed dc-dc converter and presents the derivation the transfer function of converter. Simulation and experimental results are provided in Section III and IV respectively, for a 700 W converter prototype.

II. PROPOSED DC-DC CONVERTER

A. Design and Operation

The proposed converter, figure 2, has an active h-bridge, S1-S4, on the low voltage side of a transformer, T, and a semi-active bridge, S5-S6 and D7-D8, on the high voltage side. The body diodes D1-D6 of active switches are denoted explicitly. The input voltage source is marked as $V_{in}$, output voltage as $V_o$, input capacitor filter as $C_{in}$, output capacitor filter as $C_{out}$ and the load as $R_L$.

Figure 3 shows equivalent circuit of the converter with transformer replaced by its model. The converter is operated in discontinuous conduction mode and four distinct modes in the inductor current waveform are observed during one half cycle of the base switching frequency. The circuit operation is shown in figures 4, 5, 6 and 7, for one half cycle of the base switching frequency. These distinct waveforms depend on the switching signals which are shown for each gate in figure 8. G1-G6 correspond to switching signal for S1-S6 respectively. The transformer primary voltage, secondary voltage, leakage inductance current and voltage waveforms for one half cycle are shown in figure 9. The circuit operation is explained below.

[t0–t1] – Fast Energy Storing Mode: At the start of this interval switch S2, S3 and S5 are closed at zero current condition. Closing S2 and S3 makes the primary voltage of transformer, $V_p$, equal to $V_{in}$ and S5 shorts the transformer secondary side through body diode D6, making $V_s = 0$. This means the inductor voltage is

$$V_L = V_{in} - 0 = V_{in}$$

The current in the leakage inductance of the transformer, $I_L$, will be

$$I_L = \frac{V_{in}}{L} \cdot t$$  \hspace{1cm} (1)

where $t$ varies within [t0-t1]. The leakage inductor is charged in this period at the slope of $V_{in}/L$. The current path in this period is shown in figure 4.

[t1–t2] – Slow Energy Storing & Delivery Mode: At the start of this period switch S5 is turned off while S2 and S3 are kept on. The secondary side current now completes its path through D7, load and body diode D6. This makes the voltage across the leakage inductor equal to

$$V_L = V_{in} - V_{op}$$

The current in the leakage inductance of the transformer, $I_L$, will be

$$I_L = \frac{V_{in} - V_{op}}{L} \cdot t + I_{t1}$$  \hspace{1cm} (2)

where $t$ varies within [t1-t2] and $I_{t1}$ refers to the inductor current at time $t_1$. The charging rate of inductor in this period decreases with the slope given by $(V_{in} - V_{op})/L$. However energy is also being transferred to the load in this period. The path of leakage inductor current is shown in figure 5 during this interval.

[t2–t3] – Energy Transferring Mode: At time t2, switch S2 is turned off while S3 is kept on. Inductor current can
not go instantly to zero, so it discharges the drain to source capacitance of S4, turns body diode D4 on and completes its path. The voltage across inductor is equal to

\[ V_L = -V_{op} \]

Therefore the inductor current will be

\[ I_L = -\frac{V_{op}}{L} \times t + I_{t2} \]  

where \( t \) varies within \([t2-t3]\) and \( I_{t2} \) is equal to inductor current at instant \( t2 \). The inductor is discharged during this period at the rate of \(-V_{op}/L\) and the current goes to zero at the end of this period. The current flow in this interval is shown in figure 6.

**[t3-t4] – Dead Band Mode:** This interval occurs because of the discontinuous conduction mode operation of the converter. No energy or current flow occurs through the switching elements in this interval. The length of this mode varies with change in load. The converter design should be such that the length of this interval should be greater than or equal to zero for rated load. Switch S3 is tuned off during this interval at zero current condition. This interval is shown in figure 7.

The switching scheme shows a fixed base switching frequency operation. Peak current through transformer depends on the time intervals \( t0-t1 \) and \( t1-t2 \). The switching frequency should be chosen such that the transformer core does not saturate during these intervals because of large flow of current. The switching signals for S1, S4 and S6 are also shown in figure 8. It can be seen that their operation is similar to S2, S3 and S5 during the negative cycle. This means the controller only needs to implement three independent signals, the rest can be obtained by complementing or phase shifting operation. For example G6 can be obtained by phase shifting G5 with 180° with their time period equal to \( T1 \). Similarly G1 and G2 are turned on for a time interval of \( T1+T2 \) with 180° phase difference. G3 and G4 are complementary of each other with time period equal to half of the base switching frequency.

**B. Discontinuous Conduction Mode Analysis**

As described earlier the converter operates in discontinuous conduction mode (DCM) to achieve high efficiency and easy controller implementation. It also ensures transformer core does not saturate. The DCM can be identified from the fact that current through the switching elements goes to zero during t3-t4. The mode boundary occurs when this time interval goes to zero. The converter can be analysed using DCM techniques similar to those discussed in [3] for buck and boost converters. The analysis of the proposed dc-dc converter in DCM operation is given below.

Let’s denote the duty cycle of switches S5 as \( D_1 \) which is equal to the time interval \([t0-t1]\). Let the time interval \( t1-t2 \) be marked as \( D_2 \). For controller simplification \( D_2 \) can be chosen as a constant times \( D_1 \); so \( D_2 = c \times D_1 \). Note that the
duty cycle of switches S1 and S2 is equal to \( D_1 + D_2 \). Let \( D_3 \) represent the time period \( t_2-t_3 \) during which inductor transfers its energy. The leakage inductor voltage waveform is shown in figure 9. By inductor volt-second balance equation the average voltage across inductor should be equal to zero. This can be written as

\[
\langle V_L \rangle = D_1 V_{in} + D_2 (V_{in} - \frac{V_o}{n}) - D_3 \frac{V_o}{n} = 0 \quad (4)
\]

Since \( D_2 = c*D_1 \), equation 4 can be simplified to give

\[
D_1 V_{in} + c D_1 (V_{in} - \frac{V_o}{n}) - D_3 \frac{V_o}{n} = 0 \quad (5)
\]

The charge balance equation dictates that average current through the capacitor must be zero, so the average current through the leakage inductance should be equal to the load current. Therefore using inductor current waveform from figure 9

\[
\langle I_L \rangle = \frac{1}{T_s} \left( \frac{2}{L} D_1 \left( \frac{V_{in}}{L} D_1 + \frac{V_{in}}{L} c D_1 + \frac{V_{in}}{L} D_1 - \frac{V_o}{nL} c D_1 \right) + \frac{1}{2} D_3 \left( \frac{V_{in}}{L} c D_1 + \frac{V_{in}}{L} D_1 - \frac{V_o}{nL} c D_1 \right) \right) = V_o n^2 \frac{2}{nR} \quad (6)
\]

Simplifying equation 6 gives,

\[
D_3 = \frac{-c V_{in} (DT_s)^2}{I_{pk} L} + \frac{V_o n T_s}{I_{pk} R} - c D T_s \quad (7)
\]

where \( T_s \) is the corresponding time period of base switching frequency and \( I_{pk} \) is,

\[
I_{pk} = \frac{V_{in}}{L} c D_1 T_s + \frac{V_{in}}{L} D_1 T_s - \frac{V_o}{nL} D_1 T_s
\]

Equation 5 and 7 can be simultaneously solved using a good mathematical tool to obtain the transfer function, \( M(D_1) \), for the proposed dc-dc converter.

C. Feedback Control

Feedback control was used to generate the switching signals automatically. A block diagram of the main blocks in the feedback loop are shown in figure 10. The output is sampled and scaled to calculate the error from a given reference. A simple PI controller is used for reference tracking and duty cycle generation.

D. Transformer Design

Transformer selection and design is an important factor in maximizing the efficiency of overall converter design. The following factors need to be considered in transformer design.

Maximum flux and switching frequency: The core size must be large enough to allow a large saturation flux. The switching frequency should be high enough to avoid core saturation in one half cycle.

Skin depth: The switching frequency is high in the proposed converter so the skin depth will be very small and the AC resistance of wire used in winding will be much larger than the dc resistance. So, multiple strands of enameled wire, with the diameter of each strand approximately equal to the skin depth should be used. This will minimize copper losses and allow greater power rating.

III. Simulation Experiments

Simulation of the proposed dc-dc converter was done in Matlab using SimPowerSys library tools. Input voltage was set at 24 V and output reference was set at 360 V. Power MOSFETs, with realistic on resistance and other parameters, were used in the simulation. Figure 11 shows the leakage inductance current waveform and figure 12 shows the output voltage and switching of multiple loads. The inductor current waveform demonstrates the four distinct modes which were explained earlier: fast energy storing, slow energy storing & delivery, energy transfer and dead band mode. Good load regulation was observed with the output voltage ripple measured to be in tens of millivolts.

The converter was tested in closed loop with a PI controller. Figure 12 shows that load variation brings a very small ripple in the output waveform, verifying the good load regulation of
IV. IMPLEMENTATION RESULTS

A 700 W prototype of the proposed converter was implemented in hardware. The converter was tested with an input voltage of 24 V and output reference of 220 V over a range of loads. The controller was implemented on an FPGA board which had an interfaced analog to digital converter for output sampling. Efficiency results of approximately 88% were achieved at rated load and we expect the efficiency to increase significantly with new low on resistance and high frequency switching devices. Figure 13 shows the efficiency of converter plotted over a range of loads. Output voltage variation was minimum with varying input and load. Hardware components used in converter are given in Table I along with their details.

V. CONCLUSION

We have presented a novel modified Dual-active bridge DC-DC converter for PV panel integration. The converter uses leakage inductance of a high frequency transformer for its boost operation and employs soft switching control scheme to achieve high efficiency results. Detailed working along with an analysis of DCM operation is also presented. The converter design was verified and implemented in hardware for validation. Efficiency can be significantly improved by developing a high frequency design and using new low ‘on’ resistance switching devices. The converter is suitable for integration of solar panels to a DC rail in solar farms.

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REFERENCES


