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Original scientific paper

Soft switched voltage multiplier cell (VMC) based DC-DC converter is proposed in this paper. The proposed converter provides the required high voltage gain without using a transformer. Thus, a compact converter with high power density is obtained. The proposed topology has higher efficiency due to soft switching of active device. Experimental results prove that the converter operates with a maximum efficiency of 95% and provides good load regulation.

Key words: Automotive applications, DC-DC converters, Power electronics, Soft switching

DC-DC pretvarač zasnovan na tehnologiji laganog prebacivanja za primjene u auto industriji. DC-DC pretvarači zasnovani na tehnologiji ćelija naponskog množila predloženi su u ovome članku. Predloženi pretvarači pružaju zahtjevanu visoku voltažu be uporabe transformatora. Slijedom toga, dobivamo kompaktni pretvarač s visokom gustoćom energije. Predložena topologija ima veću efikasnost zahvaljujući laganon predacivanju aktivnih komponenti. Eksperimentalni rezultati pokazuju da pretvarač radi sa maksimalno 95% korisnosti ujedno pružajući i dobru regulaciju tereta.

Ključne riječi: primjenene u automobilizmu, DC-DC konverteri, učinska elektronika, lagano prebacivanje

1 INTRODUCTION

With the advancement in automotive application, there is an incremental demand of efficient DC-DC converters. Automotive applications require high step up gain, high efficiency and reduced weight, volume and cost. Classical converter employs transformer which are magnetically coupled to achieve high step up voltage gain. However, the size of the transformer increases proportional to gain. Some of the drawbacks of using the transformer are leakage inductance and electromagnetic interference (EMI). These problems reduce the overall converter efficiency. The efficiency of classical converter is also affected when the switching frequency is increased with an idea of reducing the size. However by employing soft switching techniques the switching losses and electromagnetic interference generation can be minimized. In [1], a novel switching DC-DC converter is proposed in which large voltage step-down ratios can be achieved without a very small duty ratio and without a transformer. The absence of a transformer and the larger duty ratio permit operation at a high switching frequency and make the circuit amenable to partial integration and hybrid construction techniques. However, the converter is limited to three stages, low voltage and low power.

A switch capacitor based step up DC-DC converter is proposed in [2] and [3]. The operation of the power switches are determined by the PWM circuit. The converters are operated with high efficiency with low duty cycle. These converters are suited only for small power since they use more numbers of capacitors. In [4] and [5], a DC-DC converter using coupled inductors and diodes is proposed to provide high efficiency. High efficiency is achieved because the leakage energy is recycled and the output rectifier reverse-recovery problem is reduced. These converters are applicable for fuel cell based converters.

The introduction of voltage multiplier cell (VMC) based DC-DC converter is proposed in [6], [7], [8]. This provides a high step up gain. This converter also uses non-isolated topology which provides reduction in size and volume. However, the diode recovery losses are high in the existing VMC. Modified VMC based converters were proposed for renewable energy sources like photovoltaic and fuel cell based applications in [9]-[15]. Though many of these topologies use only single switch, all the topologies are hard switched converters. In addition, use of many passive components introduces complexities in creating a

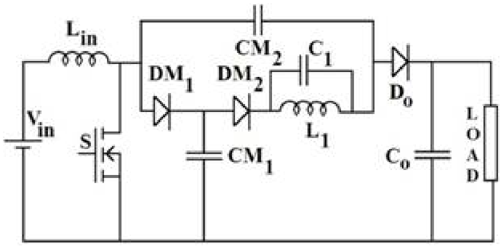

Fig. 1: Proposed converter circuit

modular structure.

Interleaved boost converter in conjunction with a VMC is proposed in [16]. Though the power handling capability and efficiency is improved, presence of two switches and coupled inductors make the circuit slightly bulky and complex. A three state switching cell was used along with VMC in [17]. This resulted in higher power handling capacity. However, due to the presence of two switches, eight diodes, inductor and an autotransformer, power density of the circuit will be lesser and hence may not suit the chosen wiper motor application. A single stage VMC with a built-in transformer was added to the conventional boost converter in [18] and [19]. The circuit provided good efficiency over a relatively higher power output. However, the voltage stress across the diodes was equal to the output voltage.

In this paper, the losses in the main switch and diodes are minimized by incorporating soft switching technique which improves the efficiency by 2% compared to the converter in [8]. Further, the diodes stresses are half of the output voltage.

2 BOOST CONVERTER WITH SINGLE MULTIPLIER CELL

Figure. 1 shows the circuit diagram of the proposed converter. The proposed converter is a combination of the classical boost converter and one VMC. This structure provides output voltage higher than the input voltage without use of magnetic elements. Additional voltage boost-up can be achieved by increasing the number of voltage multiplier cells as per requirements.

2.1 Circuit Description of the Existing Converter

The basic structure of the single phase VMC comprises of diodes $DM_1 - DM_2$ and capacitors $CM_1 - CM_2$. This cell can be used with classical converters such as buck, boost, and buck-boost configuration. The proposed circuit is formed by the switch S, boost inductor L_{in}, output capacitor C_o and output diode D_o, as presented in Fig 1. The conventional VMC operates without the resonant elements L_1 and C_1. The resonant tank is used to achieve soft switching of the power switch and suppress the negative effects of the reverse recovery current of all diodes.

2.2 Circuit Operation

The operation characteristics can be obtained when the converter operates in continuous conduction mode CCM. The operation can be classified into five modes considering the use of only one multiplier stage. ($M = 1$).

2.2.1 Mode 1

The switch is off at the start of mode 1. At this instant t_0, the energy stored in the input inductor L_{in} is transferred to the output capacitor C_o through the diode D_o. This energy is also transferred to the multiplier capacitor CM_1 through the diode DM_1. The resonant inductor current increases linearly with the input inductor current causing reduction in diode DM_1 current. The voltage across the capacitor CM_1 increases while the voltage across the capacitor CM_2 decreases in the same ratio. Hence, the constant voltage is maintained across the resonant tank. The resonant inductor current and capacitor voltage is computed from the previous states and are given by (1) and (2).

$$i_{L_1}(t) = \frac{V_{CM1}(t_5) + V_{CM2}(t_5) - V_o}{L_1} \times t \quad (1)$$

$$V_{CM1}(t) = V_{CM1}(t_5) + \left(\frac{i_{L_{in}}(t_o)t - \frac{V_{CM1}(t_5)+V_{CM2}(t_5)-V_o)t2}{L_r 2}}{C_{M1}} \right) \quad (2)$$

$$V_{CM2}(t) = V_{CM2}(t_5) - \left(\frac{\frac{V_{CM1}(t_5)+V_{CM2}(t_5)-V_o)t^2}{L_r 2}}{C_{M2}} \right) \quad (3)$$

2.2.2 Mode 2

This mode begins when the current in the diode DM_1 is zero at t_1. The resonant inductor current is equal to the input inductor current. The energy in the input inductor is transferred to the load through the diode D_o. The governing equations are given below.

$$i_{L_r}(t) = i_{L_{in}}(t) \quad (4)$$

$$V_{CM1}(t) = V_{CM1}(t_1) \quad (5)$$

$$V_{CM2}(t) = V_{CM2}(t_1) - \left(\frac{i_{L_{in}}(t_1)t}{C_{M2}} \right) \quad (6)$$

2.2.3 Mode 3

In this mode, the switch S is turned on using zero current switching commutation causing the reduction in resonant inductor L_r current, and the output diode Do current to zero at t_3. This blocks the output diode with low reverse recovery current. Due to the short duration of this stage, the capacitor CM_2 voltage can be considered constant.

$$i_{L_r}(t) = i_{L_{in}} - \frac{V_o}{2L_r}t \tag{7}$$

$$V_{CM_1}(t) = V_{CM_1}(t_1) \tag{8}$$

$$V_{CM_2}(t) = V_{CM_2}(t_2) \tag{9}$$

2.2.4 Mode 4

This mode commences when the diode DM_2 conducts and transfers the energy stored in the capacitor CM_1 to the capacitor CM_2 at time t_4. The voltage across the output capacitor is equal to the sum of output voltage of the classical boost converter and voltage across CM_2. The maximum voltage stress applied in all diodes and the power switch will be equal to the output voltage of the converter which is the combined voltages of the CM_1 and CM_2. The governing equations are given by

$$i_{Lr}(t) = \frac{(V_{CM1}(t_3) - V_{CM2}(t_3)) \, . \sin(\omega_0 t)}{\sqrt{L_r/Ceq}} \tag{10}$$

$$C_{eq} = \frac{C_{M1}C_{M2}}{C_{M1} + C_{M2}} \tag{11}$$

$$\omega_0 = \frac{1}{\sqrt{L_r C_{eq}}} \tag{12}$$

$$V_{CM1}(t) = V_{CM1}(t_3) - (V_{CM1}(t_3) - V_{CM2}(t)\cos(\omega_0 t) \tag{13}$$

$$V_{CM2}(t) = V_{CM2}(t_3) + (V_{CM1}(t_3) - V_{CM2}(t)\cos(\omega_0 t) \tag{14}$$

2.2.5 Mode 5

This mode starts when the current in the resonant inductor L_r becomes zero and the diode DM_2 is blocked. This causes the storage of energy in the input inductor until the switch is turned off and the cycle repeats to the first stage. The governing equations are given below.

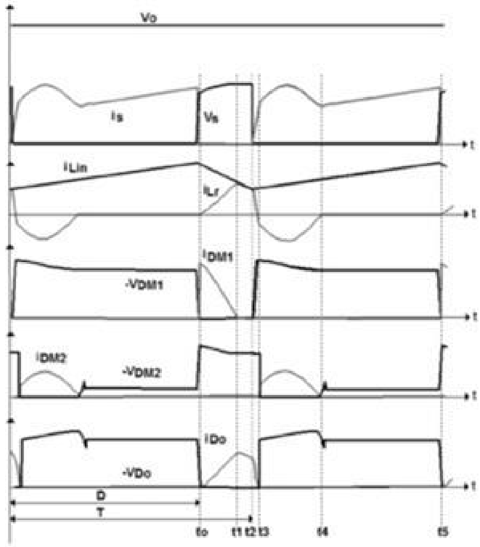

Fig. 2: Characteristic waveforms

$$i_{Lr}(t) = 0 \tag{15}$$

$$V_{CM1}(t) = V_{CM1}(t_4) \tag{16}$$

$$V_{CM2}(t) = V_{CM2}(t_4) \tag{17}$$

Fig 2 shows the characteristic waveforms of the proposed converter.

2.3 Design considerations

2.3.1 Static gain

The voltage across the capacitor CM_2 will be equal to the output voltage at the fourth stage. The capacitor is charged to twice the output voltage of the boost stage due to the stored energy in the input inductance. Thus, for single stage, the output voltage is given by

$$V_{CM2} = V_{CM1} = V_{in}\frac{1}{(1-D)} \tag{18}$$

$$V_0 = V_{CM2} + V_{in}\frac{1}{(1-D)} \tag{19}$$

Therefore for M stages, the output voltage will be multiplied by $M+1$. Thus the static gain is given by

$$q = \frac{V_0}{V_{in}} = \frac{(M+1)}{(1-D)} \tag{20}$$

2.3.2 Switch voltage

The maximum voltage in all diodes and power switch will be equal to the voltage multiplier capacitor CM_1 which is same as the output voltage of the converter.

$$V_{CM1} = Vs = Vd = V_{in} \cdot \frac{1}{(1-D)} \quad (21)$$

2.3.3 Input Inductance The design of the input inductance depends upon the current ripple of the nominal output current. The input inductance is given by

$$L_{in} = \frac{V_{in} D}{\Delta I_L f} \quad (22)$$

2.3.3 Voltage Multiplier Capacitor

The value of multiplier capacitor depends upon the maximum output power, the multiplier capacitor voltage and the switching frequency. The maximum output power is limited by the energy stored in the multiplier capacitor. When the load power increases above maximum value, the output voltage reduces, limiting the output power to maximum value. Thus the converter operates with constant output power even in overload condition until the output voltage reaches the output voltage of the boost converter. The power limitation of the circuit can increase the converter reliability even in overload operation, but a current protection circuit is needed in order to avoid short-circuit problem. Based on this condition, the value of the multiplier capacitor is given by

$$CM_1 \geq \frac{P_{o\max}}{V_{CM_1}^2 f} \quad (23)$$

2.3.4 Resonant Inductor

The resonant inductor is used to minimize the commutation losses. The variation in current is limited by the presence of resonance inductor, defined by

$$\frac{di}{dt} = \frac{Vo - V_{CM_2}}{L_r} \quad (24)$$

2.3.5 Switch Conduction Loss

The energy transfer from capacitor CM_1 to CM_2 does not change significantly. The switch current and RMS current can be determined approximately neglecting the effect of current ripple in the input inductance. The switch conduction loss is calculated as

$$IS_{rms} = \frac{P_{in}}{V_{in}} \sqrt{D} \quad (25)$$

$$P_{SW(cond)} = IS_{rms}^2 R_{DS_{on}} \quad (26)$$

2.3.6 Switch Commutation Losses

Switch commutation loss is the most important loss in the converters. Switch commutation loss depends upon the commutation current which is sum of the switch current and diode reverse recovery current. Since zero current switching is used in turn-on commutation, the commutation loss is greatly reduced. The power loss of the turn-off commutation is given by the area of the switch voltage Vs and switch current Is multiplied by the switch frequency. Thus commutation loss is given by

$$P_{SW(off)} = \left(\frac{1}{2} V_s I_s t_{off}\right) f \quad (27)$$

2.3.7 Diodes Conduction Loss

The average current in all diodes is equal to the output current. Therefore the conduction loss increases with low output voltage and high output power. Thus the calculation of conduction loss is necessary in order to determine the maximize efficiency.

The conduction loss of multiplier diodes is given by

$$I_{DM_1} = I_{DM_M} = I_{D_O} \quad (28)$$

$$P_D = 3 \frac{P_O}{V_O} V_f \quad (29)$$

2.3.8 Theoretical Efficiency

The expected theoretical efficiency of the converter can be determined depending upon the losses calculated.

$$\eta = \frac{P_o}{P_o + P_{SW(Cond)} + P_{SW(off)} + P_D + P_{Lin}} \times 100\% \quad (30)$$

3 SIMULATION RESULTS

The chosen converter with one multiplier stage was simulated with the following specifications; input voltage=12 V, output voltage=100 V, output power=100 W, switching frequency=50 kHz, duty cycle = 0.75. Fig. 3 shows the simulated output voltage waveform along with the gate pulse and inductor current waveforms. It is observed that the output voltage meets the desired specifications. Further, the input inductor current behavior can also be clearly seen. The input inductor current increases linearly during application of a gate pulse and decreases when the gate pulse is removed. This enables charging and discharging of the inductor.

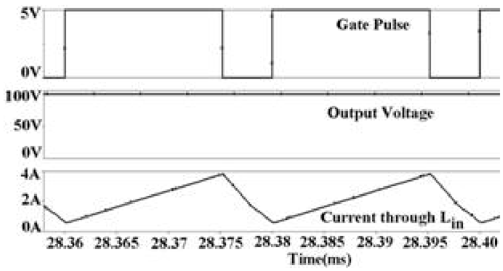

Fig. 3: Simulated output voltage, inductor current and gate pulse waveform of the proposed converter

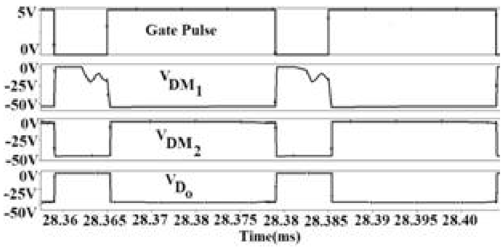

Fig. 4: Simulated voltage waveform across the multiplier diodes

Figure 4. shows the multiplier diode waveforms. It is observed that the voltage stress across all the multiplier diodes is around 50 V, which is within the stipulated limits. Further, the voltage rating of the diodes can be made to be similar due to similar voltage stress.

4 EXPERIMENTAL RESULTS

The experimental prototype of the proposed multiplier cell based converter was constructed and tested for its performance. A wind shield wiper motor was used as load. The gate pulse was generated by suitably programming the AT89C2051 microcontroller. After carefully isolating and buffering the microcontroller output, a low side driver TPS2812 was used to amplify the gate pulse. Later, it was applied to the gate of the power device IRF540N.

Figure 5 shows the gate pulse, inductor current and the output voltage. It can be clearly seen that the duty cycle is maintained at 0.75 at 50 kHz switching frequency. The output voltage is 106 V at 100 W output power. This is close to the designed output values. Further, to test the converter under slightly overload conditions, the load power was increased to 120 W. Figure 6 shows the output voltage and inductor current waveforms at this slight overload condition. Increment in input inductor current waveform can also be clearly seen.

Figure 7 shows the switching loss waveform that was observed across the power device. It is observed that the turn-on loss is much reduced compared to the turn-off loss.

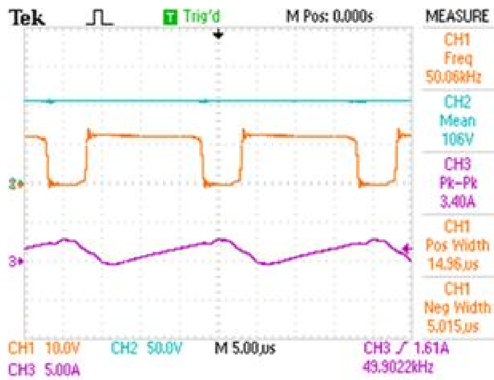

Fig. 5: Experimental output at full load condition

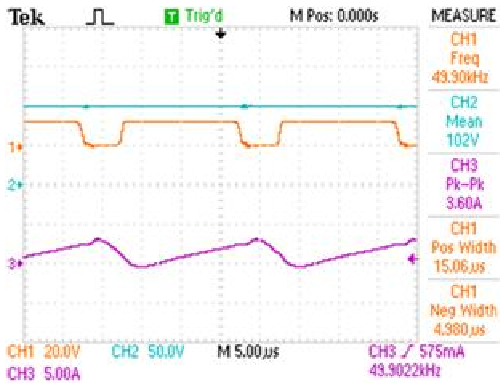

Fig. 6: Experimental output at slightly overload condition

This is due to the introduction of resonant tank elements. This has caused a significant improvement in efficiency when compared to the previous power circuit. Further, by using soft recovery diodes, the switching loss across the multiplier diodes was also reduced.

Figure 8 shows the voltage across the multiplier diodes. It is observed that similar to the simulation results, the diodes DM_1 and D_o operate complementary to the diode DM2. Further, the voltage stress across the diodes is below 50V. This is extremely advantageous since the diodes need to be rated for just half of the output voltage.

Figure 9 shows the efficiency curve of the proposed converter and the existing converter at various output power levels. Table 1 shows various loss components and its comparison with the existing converter. It is found that the reduction in power loss across the main switch and the multiplier diodes has contributed to improvement in efficiency. Figure 10 shows the photograph of the experimental setup.

5 CONCLUSION

In this paper, soft switched voltage multiplier cell based DC-DC converter was proposed. The proposed converter

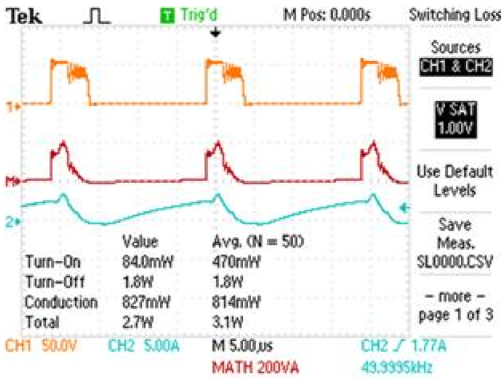

Fig. 7: Switching loss waveform

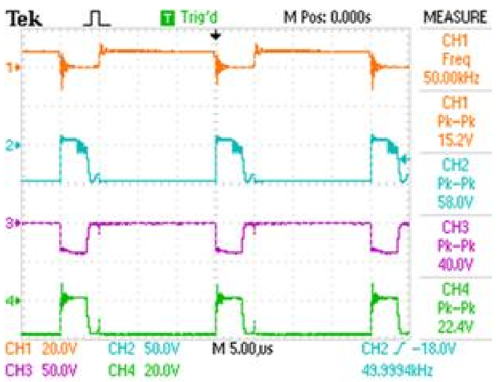

Fig. 8: Voltage across diodes

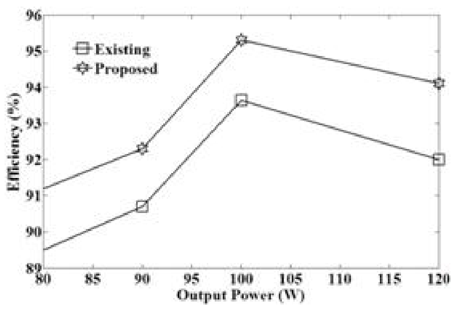

Fig. 9: Efficiency curve of the existing and proposed converter

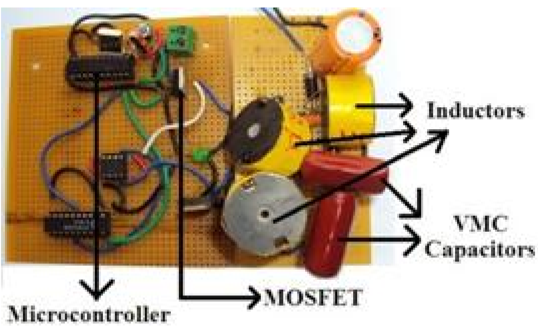

Fig. 10: Photograph of the experimental setup

Table 1: Various loss components for 100 W power

Element	Loss in watts	
	Existing	***Proposed***
Switch conduction	1.58	0.814
Turn-on	0.435	0.470
Turn-off	40.625	1.8
Diodes	3.6	1.016
Inductor	1	0.6
Total	7.24	4.7

used one multiplier stage to obtain the required voltage gain. The existing converter used hard switching as a result of which the efficiency was reduced. Experimental results obtained from the proposed soft switched converter prove that the proposed solution is more efficient and provides good load regulation. As the size of magnetic component is less, high power density can be achieved. Further, higher voltage gain can be easily obtained by adding required number of voltage multiplier cells. Thus, the proposed converter proves to be a good candidate topology for obtaining high gain compact converters with modular structure.

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