FPGA based Hybrid Resonant Switching DC/DC converter for Electric Vehicles

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Abstract*: A *hybrid switching DC-DC converter for Hybrid Electric Vehicle with reduced on state conduction losses, voltage stress and switching losses of the power semiconductor devices is presented in this paper. Zero current switching and zero voltage switching is achieved for the leading and lagging legs of the inverter by using bipolar switching method. This reduces the circulatory losses in the transformer primary. Additionally it aids in exploiting the leakage inductance of the transformer to resonate with the output capacitor. Consequently this reduces the components count and the converter size. This results in efficient energy transfer as the inductive energy and capacitive energy acquired by the output inductor and resonant capacitor are simultaneously transferred to the load during the freewheeling interval. This boosts the converter efficiency when compared to the conventional converter. During active intervals, inductor is charged and the increased capacitor voltage is offered to the load. The bipolar circuit voltage is clamped during the freewheeling interval and reduces the peak voltage overshoots that crop up during the diode reverse bias period. Simulation for the DC-DC converter is rendered with PSpice software and the experimental results shows the desired output is achieved with reduced losses under variable load conditions.

Index Terms*: Hybrid electric vehicles, DC/DC Resonant converters, battery chargers, Zero Voltage Switching, Zero Current Switching.

I. INTRODUCTION

Generally a high voltage battery for powering the Plug-in hybrid electric vehicle [1,2] (PHEV) possess higher density to store energy in a way to drive electric traction system. Power factor correction at the supply point is also important which has isolated dc-dc converter next to it. Power factor correction helps in improving the quality of the input current and the dc voltage regulation. DC-DC converter is helpful in charging the high voltage pack. It also isolates the service mains from the traction battery pack.

The conventional phase shifted full bridge dc/dc converter is the prominent converter used as the second Interval of PHEV charger system. Zero Voltage Switching (ZVS) can also be realized by exploiting the parasitic capacitance of the switches as well as the leakage inductance of the transformer. Conventional converter has many limitations in achieving ZVS for wide voltage range [3-7].The range of ZVS of lagging leg inverter switches is found to be limited. Resonance is achieved from half to full load condition. Voltage across leakage inductor (lk) is responsible for achieving ZVS in lagging – leg switches however this condition fails when load conditions are low.

To overcome the above problem, the values of the leakage inductance can be increased or a serial inductor can be added externally. Adding a higher value inductance may introduce a delay in the primary current. This delay may further lead to loss in the transformer secondary. To compensate these losses, a transformer with higher turn's ratio must be employed. But this results in high voltage stress for the switches in bridge rectifier. A number of methods are reported in the literature for exploiting the inductive energy accumulated in supplementary circuits, instead of adding external series inductance to increase the ZVS range [5,6]. Three major drawbacks of conventional methods are as follows: First, the batteries of electric vehicle applications may require wide range of load and duty cycle. ZVS cannot be attained at wider range. Secondly, the primary current on the primary part circulates during freewheeling period. This contributes to unnecessary conduction losses. The third major drawback is the voltage overshoot across the diodes in the output side. This is due to parasitic ringing occurs in the junction capacitors and also in transformer leakage inductance. RCD[7] snubber circuits are used to reduce voltage overshoots in the bridge rectifier. But this increases the losses and decreases the converter efficiency. However solving this problem intensifies the complexity and reduces the system reliability. Several clamp circuits with effective passive components are proposed [8,9]. Though these techniques are able to give a solution for reducing the voltage stress, the value of these voltage stress rely on the duty cycle in addition to the output voltage value [10-12]. Therefore, these techniques were commonly not considered for battery charger circuit's applications.

II. HYBRID SWITCHING RESONANT DC-DC CONVERTER

Fig.1 shows the proposed circuitry for hybrid switching resonant DC-DC converter. It consists of an inverter having a higher frequency with reduced voltage ripple at the load end and also includes four switches ($S_1 - S_4$) connected to the isolation transformer primary end and diode bridge rectifier on secondary end for synchronous rectification. This topology employs ZVS for the four switches which lead to a diminished switching losses and higher converter efficiency. Switches (S_1, S_2, S_3 and S_4) operate in pairs, hence the term bipolar. The switches S_1/S_2 are considered as leading leg pair and S_3/S_4 as switches at the lagging leg. ZVS of the leading leg can be attained using transformer leakage

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inductance. However, since the ZVS in lagging leg switches S_3 and S_4 is attained using energy accumulation in leakage inductance of the transformer, it fails to accomplish ZVS [13,14] under low load conditions.

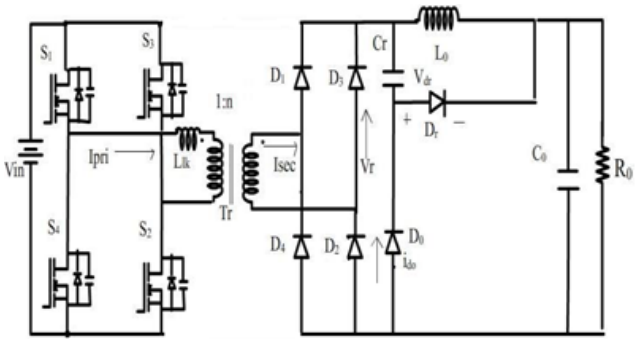


Fig.1 Proposed resonant dc/dc converter

Hybrid resonant waveforms and pulse width modulated current waveforms are used to examine the resonant converter. During active intervals, the energy gets discharged from the input supply to the load along a resonant path via transformer. This enhances the power transfer from input power supply to load. All along the freewheeling period, the accumulated energy in the resonant path will be sent to the load. Fig.1 depicts the proposed converter which provides an additional circulating current which in turn forces the primary current to zero at the beginning of freewheeling period. It should be noted that, during this interval the voltage across C_r is the same as the voltage across the uncontrolled converter. This decreases the voltage stress (dv/dt) across the rectifiers. The working principle of the classical DC-DC converter relies on the stored inductive energy [15] whereas the proposed converter has an added capacitive energy transmitted to the output by hybrid switching method. The circulating current is nonexistent for the period of the freewheeling interval [16-19].

III. DETAILED OPERATION OF HYBRID SWITCHING RESONANT CONVERTER

The converter function is reviewed with the aid of three operational modes:

a. Mode 1: The resonant current that passes through the diode D_r (I_{dr}) forms the basis for the classification of Mode 1 into six topological Intervals. Fig 2 depicts the waveforms for different Intervals of Mode 1. The voltages in the primary and secondary part of the transformer are V_{pri} and $V_{\sec}$; $ipri$ and $i\sec$ are the current flowing through transformer primary and secondary end. iLo represents output inductor current and it is higher than $i\sec$ as shown in the figure 2.

Interval 1: [Fig 3(a)]. Interval 1 starts before t_1. During this period, the currents flowing through both primary and secondary windings are zero. Resonant capacitor C_r turns off the bridge rectifier diodes by applying reverse voltage. The stored energy in the resonant capacitor is released through diode D_0 and inductor Lo. At time t_1, switch S_2 turns ON. As a result, the primary and secondary currents of the transformer increase. This mode changes once the secondary end current I_S is higher than the inductor current (iLo) at the output.

Interval 2: This interval starts when the secondary end current is greater than the output inductor current which turns ON diode D_r and brings D_0 under reverse bias condition. The Interval 2 operation of mode1 is shown in Fig.3(b).This result in an increase in the power delivered to output, because the energy transferred is the parallel combination of sinusoidal resonant mode and linear PWM modes.

$$Tr = n.(ilo + idr) \qquad (1)$$

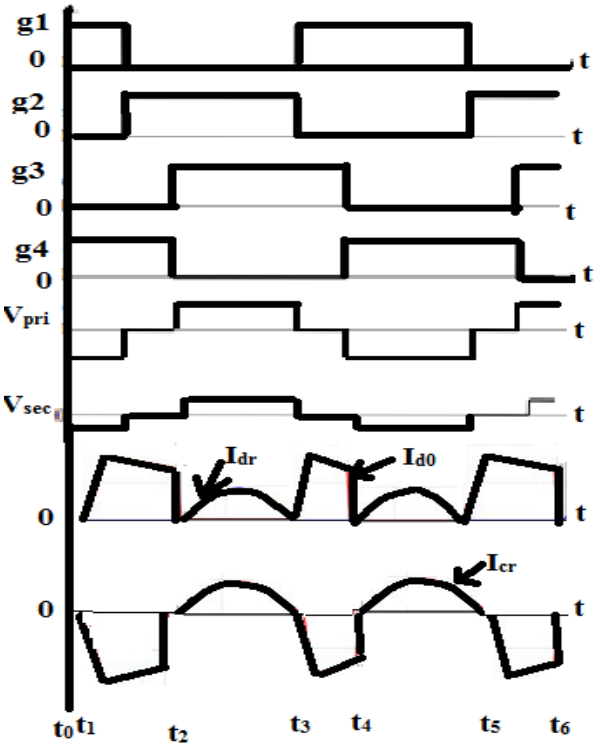


Fig.2 Converter Waveforms (Theoretical)

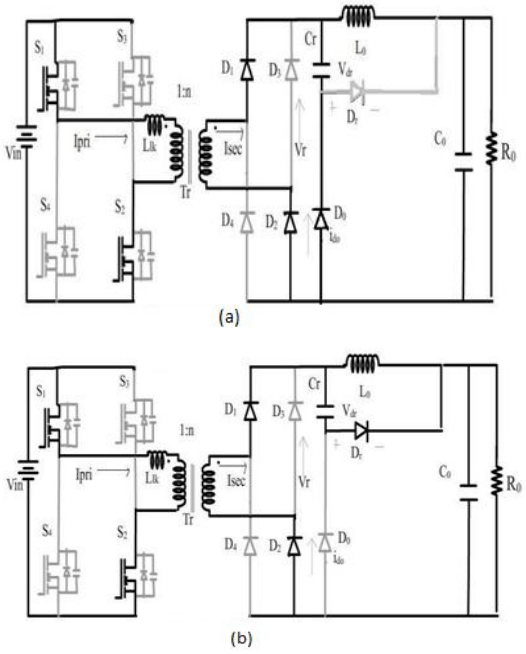


Fig.3 Topological intervals of Mode 1 (a) Interval 1 (b) Interval 2

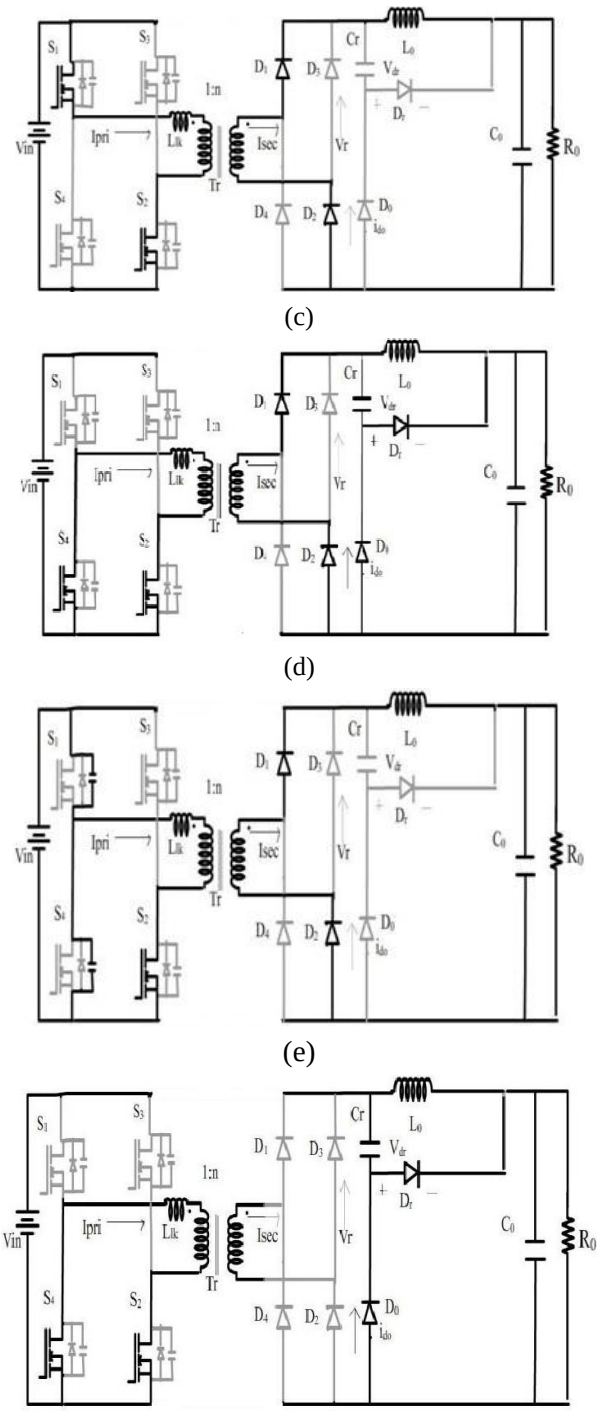


Fig 4.Mode 1 operation (c) Interval3 (d) Interval4 (e) Interval 5 (f) Interval 6

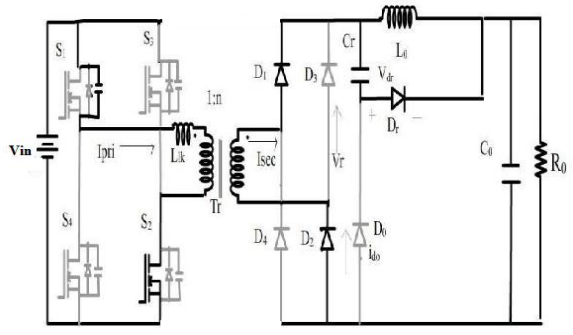


Fig.5 Mode 3 operation

Interval 3: Interval 3 operation of mode 1 starts at time instant t_2 as represented in Fig.4(c). The diode current (Idr) resonates to zero at time instant t_2. The stored energy in transformer windings are dissipated to load until the current through the secondary equals to Ilo.

Interval 4: [Fig 4(d)]. In this interval the switch S_1 remains off and the transformer primary current ($ipri$) charges and discharges the junction capacitance of switch S_3. As the drain– source voltage ($Vds3$) of switch S_3 attains zero, $ipri$ flows through the S_3 body diode and once the dead time passes, S4 is turned ON with ZVS.

Interval 5: [Fig 4(e)]. Secondary side current is lesser compared to Ilo, which in turn brings the diode D_0 under forward bias condition. The voltage across the bridge rectifier matches the capacitor voltage (V_{cr}). Thus the primary current becomes zero turning off S_1 and S_3.

Interval 6:[Fig 4(f)]. Currents flowing in through the transformer turns are zero. Stored energy in C_r and L_0 is dissipated to the load, until S_4 switch is turned ON. The bridge rectifier is under reverse bias condition as the resonant capacitor (C_r) voltage is usually more than zero.

b. Mode 2: The operation of Mode 2 is analogous to the previous mode. The variation in Mode 2 occurs when the resonant current Idr approaches zero and switch S_1 turns OFF.

c. Mode 3 [Fig 5]: In mode 3,resonant diode current is more than zero. When converter operates in freewheeling period, resonance halts and Switch S1 is turned OFF. Primary current comprises of the resonant diode current and the current at the output inductor. As the current flowing through S_3 approaches zero, charging and discharging of the junction capacitor occurs. Primary current flows through body diode once S_1 and S_3 are in OFF state. The following equation is obtained after making use of flux balance condition for inductor:

$$VCr*Dr+(nVin-Vo)*(D-Dr)+(1-D)*(V0-Vcr) \qquad (2)$$

IV. HYBRID SWITCHING RESONANT DC / DC CONVERTER CHARACTERISTICS:

In the conventional converter during active period, the voltage drop occurs across two diodes in the bridge rectifier whereas in the suggested topology the voltage drop is only across one diode. This is due to the voltage drop across C_r (resonant capacitor) is greater than zero. Fast recovery diode has an advantage of high rate of commutation and less reverse recovery current. As the transformer secondary voltage is high, a large voltage drop occurs across the diode bridge rectifier which might cause higher conduction losses. But this does not alleviate the diodes reverse recovery problem and can be resolved by considering the transformers leakage inductance. The energy is stored in C_r and L_o, when the energy transfer occurs between the supply and the load. During the freewheeling period this energy is transferred to the load. This provides additional current to ZVS and sets the primary current to zero during the

start of freewheeling period. Hence switches at the leading leg can realize ZVS for an extensive power range. The suggested converter is more efficient than ZVS converter [16]. By employing ZVS and PWM control switching losses are significantly reduced.

V. DESIGN CONSIDERATIONS

This converter has constant ON-time and matches half the resonant period. The OFF-time is controlled. The ON-time equation is obtained as

$$\text{Ton} = \text{D.Ts} = \frac{\text{Tr}}{2} = \pi \text{Cr}\sqrt{\text{Lr}} \tag{3}$$

Since the transformer leakage inductance is used to attain resonance, L_r is substituted with $L_{lk.}$ Resonant time period is

$$\text{Tr} = 2\pi \text{CrL}_{\text{lk}} \tag{4}$$

To attain resonance, the dead time required among the switches S_1 and S_4 can be obtained by the following expression:

$$Cs1 + Cs4.Vin + CTr \le Ipri\ .\ \delta.t_1 \tag{5}$$

where t_1 is S_1 and S_4dead time. The design equations of the transformer are given by

$$\frac{\text{N1}}{\text{N2}} = \frac{\text{V1}}{\text{V2}} \frac{\sqrt{\text{I}_1}}{\sqrt{I_2}} \tag{6}$$

$$\frac{I_1}{I_2} = \frac{\mu \text{N1} \pi \text{r} 2}{\mu \text{N} 2 \pi \text{r} 2} \tag{7}$$

The peak ripple current equation is given by

$$\Delta i2pk = \frac{\left(\frac{1}{nM} - 1\right).\left(2 - \frac{1}{nM}\right).Vo\frac{ts}{2}}{Lo} \tag{8}$$

Where $M = \frac{Vo}{Vin} = \frac{1}{2-D}$ and M represents conversion ratio and n represents turns ratio of the transformer. The inductor current comprises of the diode current Dr and capacitor resonant current through C_r . Hence resonant capacitor charge balance equation is given by the condition that the resonant capacitor

voltage must be greater than zero and is should satisfy the following equation:

$$vcr < VCr \tag{9}$$

The critical value of the capacitance is given by

$$Cr >> \frac{(2-D).Po}{8*n^2*Vin^2*f} \tag{10}$$

Where Po represents the load power and Vin is the input voltage of the converter. The resonant capacitor current icr is expressed as

$$icr = (1\text{-}d)\ ilo \tag{11}$$

where ilo is the current through the inductor.

VI. SIMULATION RESULTS

Converter design parameters are given in Table 1. The dead time is given for the triggering pulses of switches from S_1 to S_4 to accomplish ZVS as illustrated in Fig.6 (a). The drain source voltage (Vds1, Vds2, Vds3 and Vds4) of inverter switches are given in Fig.6 (b). The voltage across the transformer primary and secondary can be noticed in Fig 6(c). The bridge rectifier voltage stress is almost identical to the secondary voltage of the transformer and hence only one voltage drop during the freewheeling interval. The load current I_0 and inductor current (I_{Lo}) are depicted in Fig 6(d).

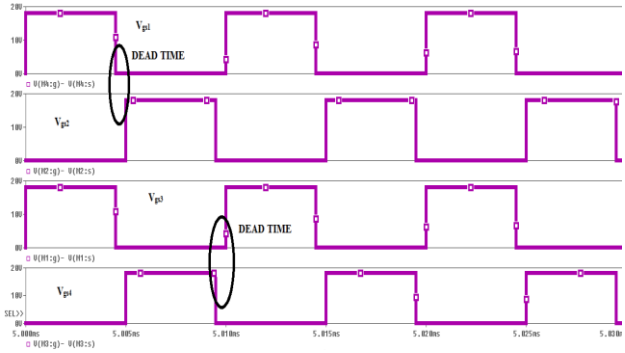


Fig.6 (a) Triggering pulses of four switches (S1- S4)

From the Fig 6(f) it can be observed that the inductor current at the load end is the addition of the current through the resonant capacitor C_r and the diode. Fig.7 reveals that the converter's operating range is obtained at a duty cycle of 0.45.

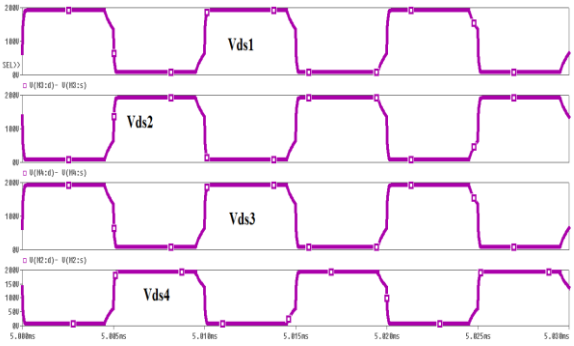


Fig.6(b) Drain source voltage (Vds) of the four switches

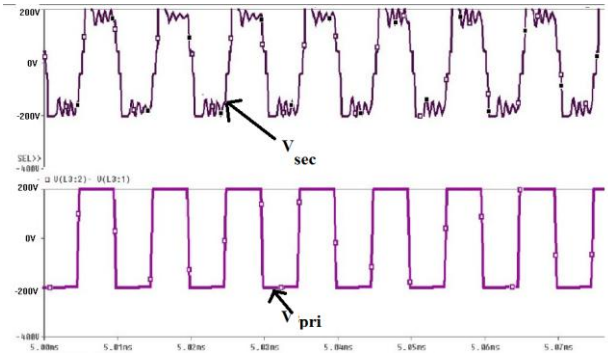


Fig.6(c) Transformer secondary and primary winding voltage

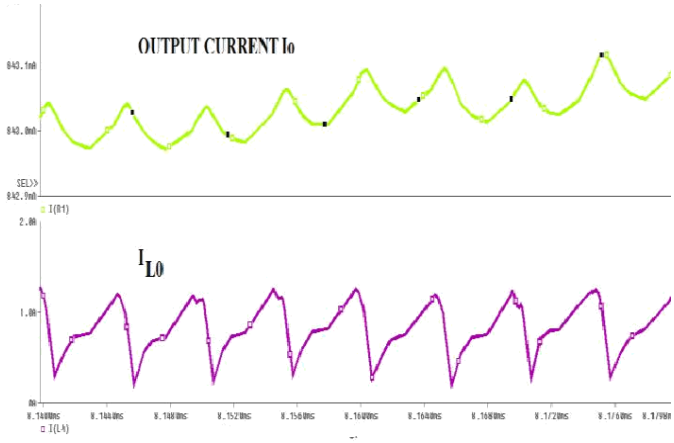


Fig 6(d) Load current and the current through the inductor L_o

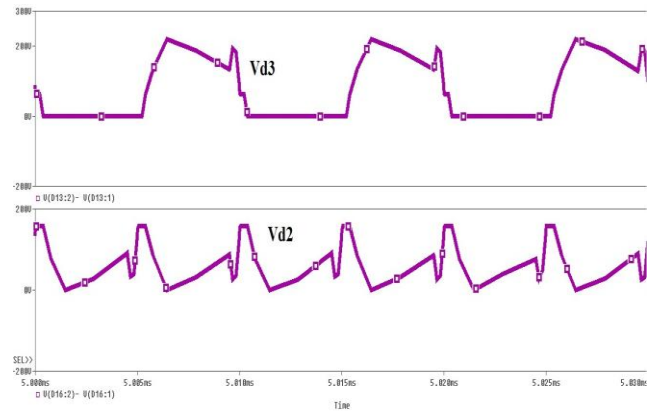


Fig.6 (e) Voltage across the uncontrolled converter

VII. HARDWARE RESULTS

A 50W experimental model is realized to validate the performance of hybrid resonant DC/DC converter. Hardware specifications are listed in Table.2 with an isolation transformer. Secondary of the transformer is coupled to the diode bridge rectifier (DBR) which consists of schottky diodes.

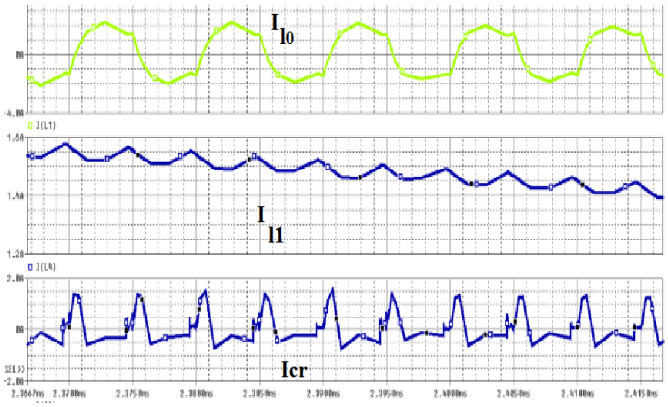


Fig. 6(f) Waveforms illustrate inductor current, primary current through the transformer and resonant capacitor current

Table 1: Converter Parameters

Parameter	Value
Input dc voltage	200V
Output load voltage	175V
Power rating	100W
Resonant capacitor (Cr)	470 nF
Output inductor (L_o)	270mH
Output capacitor(C_o)	47uF
Diodes ($D_1, D_2, D_3, D_4 D_r$)	MUR1560

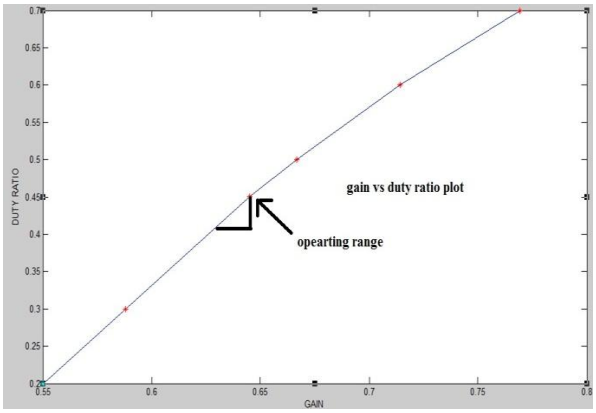
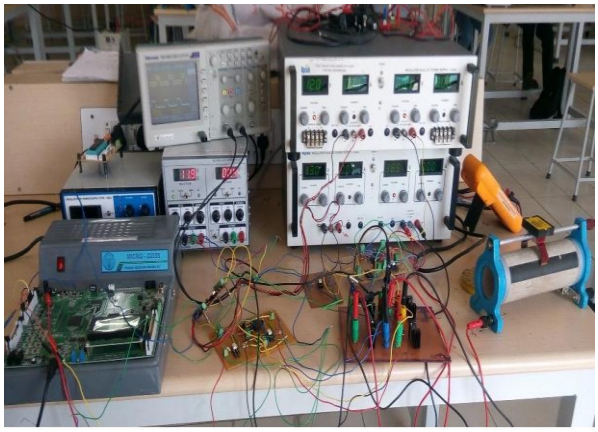


Fig.7 Gain Vs Duty ratio

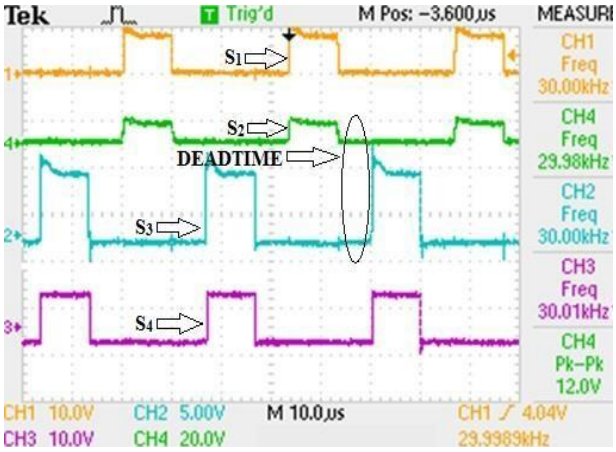
This rectifies the voltage at the input for the full bridge converter. The resonant part which contains a resonant diode, inductor and the capacitor of the topology is used to attain the zero voltage switching for the proposed converter. To generate the pulses for the proposed converter Spartan –6 XC6SLX25 FPGA board is employed. PWM output from FPGA is terminated in a 34 Pin connector through Level Translator for converting 3.3V to 5V. There are 8 pins to generate PWM pulses. FPGA board has a clock frequency of 100MHz. By dividing this internal clock frequency the desired switching frequency can be obtained. Figure 9(a) illustrates the experimental setup of the proposed converter.

Table 2 Converter Specifications

Symbol	Parameter	Value
P0	Power rating	50W
Vin	Input voltage	100VDC
VOUT	Output voltage	65VDC
fs	Switching frequency	30kHz
Imax	Max current	2A
Lr	Resonant inductor	2.7 mH
C_o	Output capacitor	47uF/150v
C_r	Resonant capacitor	0.47uF/150V (polypropylene)
Tr	Transformer turns ratio	(39:24) (V1/L1:150/4mH) (V2/L2:100/3.1mH)
S_1,S_2,S_3,S_4	Switches	IRF740
D_1,D_2,D_3,D_4,D_0,D_r	Diode	MUR1560 (schotkky diodes)

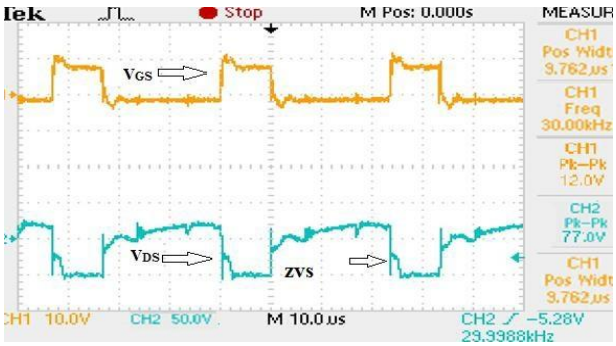


(a)



(b)

(c)



(d)

Fig 8(a) Hardware test bench (b)Gate pulses of High Side and Low side switches with fs=30kHz (c) Current across the resonant inductor (d) Waveform showing ZV switching across diode.

Figure 8(b) shows the pulses generated with the dead time. From the figure it could be seen that switches S1 and S2 are working with a phase shift of 180^0 between them. Dead time of 770ns can be observed between S_1 and S_2.The resonant inductor current is shown in Fig.8(c) and Fig 8(d) depicts the zero voltage switching of diode. The output voltage as well as the current through the resistive load is shown in Fig.9 with a magnitude of 41.3V and 1.64A.

Fig 9 The load voltage and load current waveforms

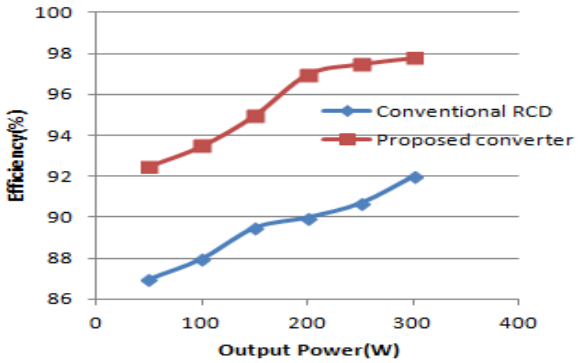


Fig.10 Efficiency Vs Output power

VIII. CONCLUSION

This paper discusses a hybrid switching resonant converter for hybrid electric vehicle applications. Zero Voltage Switching of inverter switches can be attained at variable load condition. Thus the soft switching technique eliminates the overall converter losses. The transformer leakage inductance is used to resonate along with output inductor and hence a supplementary component for achieving resonance is reduced. Since the hybrid switching resonant converter is operated at higher switching frequencies, the cost and size of the passive circuit elements are reduced. This results in overall converter size reduction and increases the converter efficiency.

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