RESONANT PWM ZVZCS DC TO DC CONVERTERS FOR RENEWABLE ENERGY APPLICATIONS

B.Karthikeyan [#1], Prof.Jebasalma [*2]

[#1] PG Scholar, Alagappa chettiar college of engineering and technology, karaikudi-4.India.

[1]karthi2balu@gmail.com.

[*2] Professor and head of the department, Alagappa chettiar college of engineering and technology, karaikudi-4.India.

Abstract— **When the conventional boost converter employed for high power application, it should operate at high duty cycle in order to achieve high-output voltage. To increase output voltage in high range, higher rating of power semiconductor switches should be used which increases the dominating conduction loss. The diode must sustain a short pulse current with high amplitude, resulting in severe reverse recovery problem as well as high electromagnetic interference problems. Moreover, a high duty cycle may lead to poor dynamic responses to line and load variations. In order to overcome these problems, non isolated high step-up dc–dc converters with soft switching technique is proposed in this project. The main drawback of pulse width modulation is high turn-off switch losses. To reduce the turn off switching losses, an improved switching method, called resonant PWM (RPWM) is proposed for the soft-switched non isolated high step-up dc–dc converters in order to reduce the turn-off switching losses.The proposed converter shows zero-voltage switching turn-on of the switches in continuous conduction mode as well as reduced turnoff switching losses. Also, as a result of the proposed switching method, the switching losses associated with diode reverse recovery become negligible even in the small duty cycle. The duty cycle loss is further reduced resulting in increased step-up ratio. Since the RPWM is performed by utilizing Lr–Cr resonance in the auxiliary circuit, the capacitance is significantly reduced compared to the pulse width modulation method. Because of reduced switching losses, diode reverse recovery problems and increased step up ratio, the proposed converter is used in dc backup energy systems for uninterruptible power supply (UPS),photovoltaic systems, fuel cell systems, and hybrid electric vehicles.**

Keywords— **High step-up, high-voltage gain, nonisolated,soft-switched.**

I. Introduction

The demand for non isolated high step-up dc–dc converters has been gradually increasing in accordance with the growth in dc backup energy systems for uninterruptible power system (UPS), photovoltaic systems, fuel cell systems, and hybrid electric vehicles. Since the general boost converter should operate at high duty cycle in order to achieve high-output voltage, the rectifier diode must sustain a short pulse current with high amplitude, resulting in severe reverse recovery as well as high electromagnetic interference problems. Also, as output voltage is increased, the switch voltage rating is increased, which increases the dominating conduction loss. Moreover, a high duty cycle may lead to poor dynamic responses to line and load variations.

Various types of non isolated high step-up dc–dc converters have been presented to overcome the aforementioned problem. Converters with coupled inductors can provide high output voltage without using high duty cycle and yet reduce the switch-voltage stress. The reverse recovery problem associated with rectifier diode is also alleviated. However, they have large input current ripple and are not suitable for high-power applications since the capacity of the magnetic core is considerable. The switched-capacitor converter does not employ an inductor making it feasible to achieve high-power density. However, the efficiency could be reduced to allow output voltage regulation. The major drawback of theses topologies is that attainable voltage gains and power levels without degrading system performances are restricted.

Most of the coupled-inductor and switched-capacitor converters are hard switched. The hard-switched CCM boost converter suffers from severe diode reverse-recovery problem in high-current high-power applications. That is, when the main switch is turned on, a shoot through of the output capacitor to ground due to the diode reverse recovery causes a large current spike through the diode and main switch. This not only incurs significant turn-off loss of the diode and turn-on loss of the main switch, but also causes severe electromagnetic interference (EMI) emission. The effect of the reverse-recovery-related problems becomes more significant for high switching frequency at high power level. Therefore, the hard-switched CCM boost converter is not capable to achieve high efficiency and high power density at high power level. Therefore, they are not suitable for high efficiency and high-power applications. Some soft-switched interleaved high step-up converter topologies have been proposed to achieve high efficiency at desired level of volume and power level. Among them, the soft-switched continuous conduction mode (CCM) boost converter demonstrated reduced voltage stresses of switches and diodes and zero-voltage switching (ZVS) turn-on of the switches in CCM and zero-current switching (ZCS) turn-off of the diodes. However, a drawback of this pulse width modulation (PWM) converter is high turn-off switch losses. In this project, an improved switching method, called resonant PWM (RPWM) is proposed for the soft-switched CCM boost converter in order to reduce the turn-off switching losses.

Since the RPWM is performed by utilizing *Lr–Cr* resonance in the auxiliary circuit, the capacitance is significantly reduced. Also, because of the proposed RPWM operation, the switching losses associated with diode reverse recovery become negligible even in the small duty cycle and the duty cycle loss is further reduced resulting in increased step-up ratio.

II. PROPOSED CONVERTER

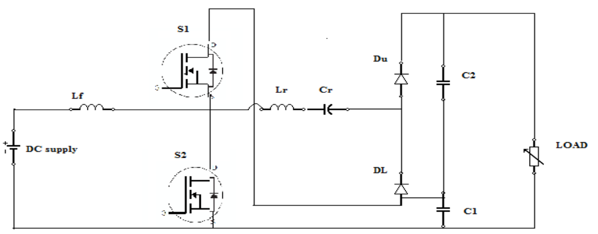


Fig 1.Proposed converter.

Fig. 1 shows the circuit diagram of the proposed converter which has the same circuit topology as the PWM method proposed converter. The proposed converter consists of a general boost converter as the main circuit and an auxiliary circuit which includes capacitor *Cr* inductor *Lr*, and two diodes *DL* and *DU* . Two switches are operated with asymmetrical complementary switching to regulate the output voltage. Owing to the auxiliary circuit, not only output voltage is raised but ZVS turn-on of two switches can naturally be achieved in CCM by using energy stored in filter inductor *Lf* and auxiliary inductor *Lr*. Unlike PWM method in which the switches are turned OFF with high peak current, the proposed converter utilizes *Lr–Cr* resonance of auxiliary circuit, thereby reducing the turn-off current of switches. Furthermore, for resonance operation, the capacitance of *Cr* is reduced by at least 20-fold, resulting in reduced volume. Also, switching losses associated with diode-reverse recovery of the proposed RPWM converter are significantly reduced.

MODE OF OPERATION:

Mode 1:

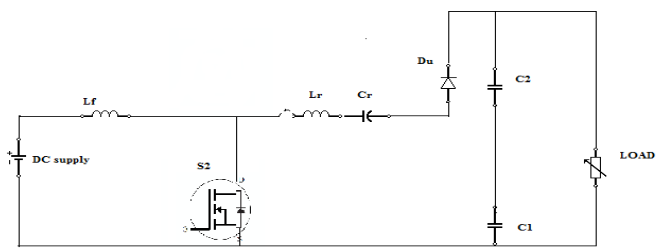


Fig. 2 operating mode 1

This mode begins when upper switch SU which was carrying the current of difference between iLf andiLr is turned OFF. SL can be turned ON with ZVS if gate signal for SL is applied before the current direction of SL is reversed. Filter inductor current iLf and auxiliary current iLr starts to linearly increase and decrease, respectively, as follows.

$$iLf = \frac{Vi}{Lf}(t - to) + iLf(to) \quad (1)$$

$$iLr = \frac{Vcr,min - Vo}{Lr}(t - to) + iLr(to) \quad (2)$$

This mode ends when decreasing current iLr changes its direction of flow. Then DU is turned OFF under ZCS condition.

Mode 2:

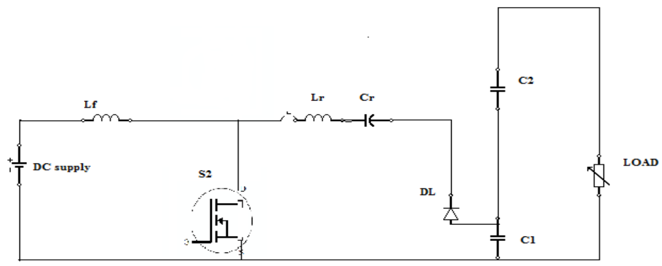


Fig 3.operating mode 2

This mode begins with Lr–Cr resonance of the auxiliary circuit. Fig 3 shows equivalent circuit of this resonant mode. Current iLf is still linearly increasing.

The voltage and current of resonant components are determined, respectively, as follows:

$$iLr = -iCr = \frac{Vr2}{Z}\sin(\omega r(t - t1)) \quad (3)$$

$$vCr(t) = Vr,2\left[\cos\left(\omega r(t - t1)\right) - 1\right] + vCr(t1)$$
(4)

$$Vr,2 = Vcr,min - Vc1, Z = \sqrt{Lr/_{Cr}}\ and\ \omega r = 1/\sqrt{LrC} \quad (5)$$

This resonance mode ends when iLr reaches to zero. Note that DL is turned OFF under ZCS condition

Mode 3:

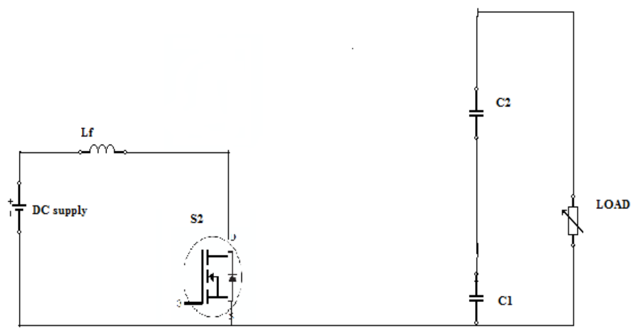


Fig 4.operating mode 3

There is no current path through the auxiliary circuit during this mode. Output capacitors supply the load. At the end of this mode the turn-off signal of SLis applied. It is noted that the turn-off current of SL,ISL,offis limited to filter inductor current at t3, ILf,max, which is much smaller than that of PWM method.

Mode 4:

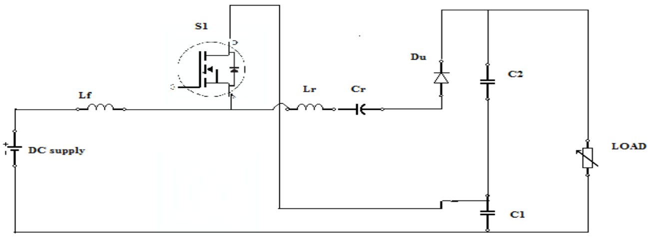


Fig 5.operating mode 4

This mode begins when lower switch *SL* is turned OFF. *S*U can be turned ONwith ZVS if gate signal for *SU* is applied before the current direction of *SU* is reversed. Filter inductor current iLfstarts to linearly decrease since voltage VLf becomes negative Like Mode 2, the other *Lr*–*Cr* resonance of auxiliary circuit is started, and *DU* starts conducting. Equivalent circuit of resonant mode is shown in fig. The voltage and current of resonant components are determined, respectively, as follows:

$$iLf = \frac{Vi - Vc1}{Lf}(t - t3) + iLf(t3) \qquad (6)$$

$$iLr = -iCr = \frac{Vr4}{Z}\sin(\omega r(t - t3)) \qquad (7)$$

$$Vr4 = Vcr,max - Vc2 \qquad (8)$$

This mode ends when iLr is equal to iLf.

Mode 5:

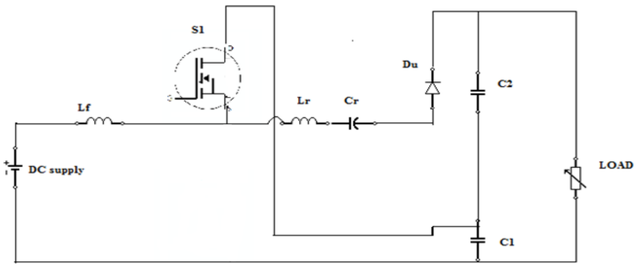


Fig 6.operating mode 5

After *iLr*equals *iLf*,*iSU*changes its direction, then this mode begins. At the end of this mode, turn-off signal of *SU* is applied and this mode ends.

III CONTROL TECHNIQUE

PWM Method Vs RPWM Method:

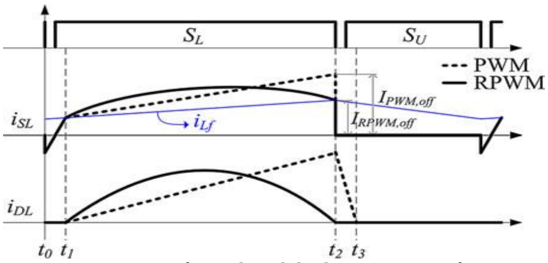


Fig 7.Comparision of switch and diode current waveforms.

Fig.7 shows the current waveforms of lower switch SL and lower diode DL of the converter illustrating the effectiveness of the proposed RPWM. As shown in Fig. 7, IRPWM,off, switch turn-off current of the proposed RPWM method, is smaller than IPWM,off, switch turn off current of PWM method which is the sum of input inductor current iLf and auxiliary inductor current iLr at turn off instant. For the proposed RPWM operation, resonant capacitor Cr is reduced by at least 20-fold compared to the auxiliary capacitor of the PWM operation which should be large enough to act as a voltage source. Furthermore, the turn-off losses associated with diode reverse recovery of the proposed RPWM converter are negligible while that of the PWM method could be somewhat considerable, especially at operation with small duty, due to high turn-off current and di/dt, as shown in Fig. 7.

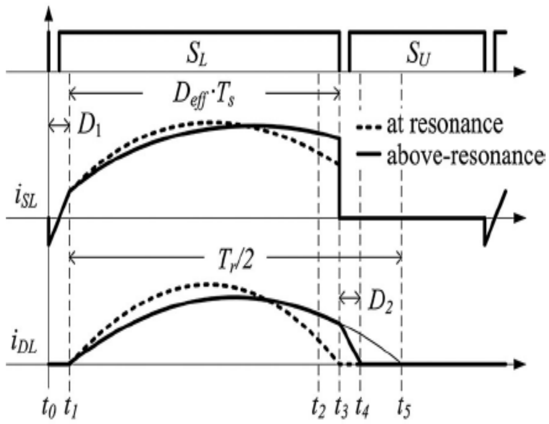


Fig 8. Comparision of switch and diode current waveforms. Of two resonant condition.

Above-Resonance Operation Vs Below-Resonance Operation:

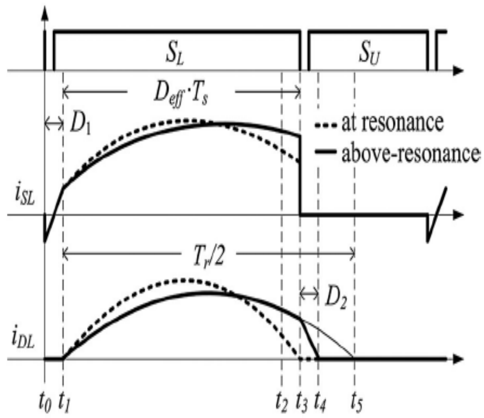


Fig 9.Comparision of switch and diode current waveforms of two resonant condition.

$$fr = 1/2\pi\sqrt{LrCr} \qquad (9)$$

It can be seen from Fig. 8 that the below-resonance operation has advantages over the above-resonance operation. First, the total switching losses are smaller for the below-resonance operation since both switch turn-off current and diode *di/dt*are smaller. Second, duty loss *D*1 of the below-resonance operation is smaller than duty loss *D*1+*D*2 of the above-resonance operation, as shown in Fig. 8. The below-resonance operation is chosen for the proposed RPWM method. For below-resonance operation, half of the resonant period (t1–t2) should be shorter than DeffTs(t1–t3). Therefore, the resonant frequency can be determined by

$$fr > (fs/2)Deff \quad (10)$$

Deffis effective duty cycle *D*–*D*1 considering the duty loss.

IV VOLTAGE CONVERSION RATIO:

To obtain the voltage gain of the proposed converter, it is assumed that the voltage across C1 and C2 are constant during the switching period Ts . The output voltage is given by

$$Vo = Vc1 + Vc2 \quad (11)$$

$$Vo = \frac{N+1}{1-D}Vi - \Delta V \quad (12)$$

$$Vo = \frac{2}{1-Deff}Vi = \frac{2}{1-D}Vi - \Delta V \quad (13)$$

where effective duty *D*effand voltage drop Δ*V* are expressed using duty loss Δ*D*

$$Deff = D - \Delta D \quad (14)$$

$$\Delta V = \frac{2\Delta DVi}{(1-D)(1-Deff)} \quad (15)$$

VC 1 that is the same as output voltage of the general boost converter can be expressed as

$$Vc1 = \frac{1}{1-D}Vi$$

(16)

$$Vc2 = \frac{1}{1-D}Vi - \Delta V$$

(17)

$$IDL,av = \frac{V0}{Ro} = \left|\frac{2}{Ts}\int_0^{\frac{Tr}{4}}(Vcrmin - Vc1)\frac{\sqrt{Cr}}{\sqrt{Lr}}\sin(\omega rt).dt\right| \quad (18)$$

$$IDU = \frac{Vo}{P.Ro} = \frac{1}{2}(1 - D - \Delta D).IL2,peak \quad (19)$$

$$IDU = \frac{Vo}{P.Ro} = \frac{1}{2}(D + \Delta D).IL2,peak \quad (20)$$

VCr,minand *VCr*,maxof capacitor voltage *VCr*can be approximated by

$$Vcrmin \approx Vc1 - \frac{V0}{2CrRofs} \quad (21)$$

$$Vcrmax \approx Vc1 + \frac{V0}{2CrRofs} \quad (22)$$

Below-resonance operation (*D* >*fs*/2*fr*)*:*

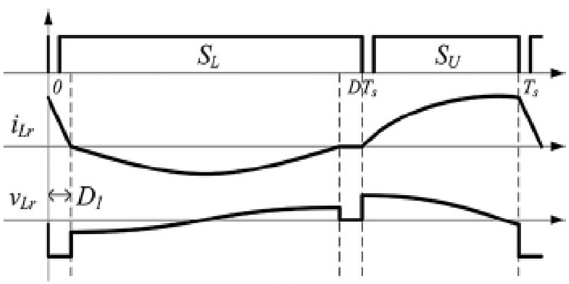


Fig 10 . below resonant operation with D>fs/2fr

In this mode, the duty loss is the same as *D*1, as shown in Fig.10 The steady-state inductor voltage equation on inductor *Lr*during *D*1*Ts* gives

$$Vcrmin - Vo = Lr\frac{\sqrt{\frac{Cr}{Lr}}(Vcrmax - Vc2)}{D1Ts} \quad (23)$$

the duty loss can be obtained by

$$\Delta D = D1 = \frac{(1-D)(fs/\omega r)\sin((1-D)\omega r/fs)}{2CrRofs\left(\frac{Vi}{Vo}\right)+(1-D)} \quad (24)$$

the voltage gain in this mode can be obtained by

$$M = \frac{D'(1-A)+\sqrt{[D'(A-1)^2+4AD'(D'+B)]}}{D'(D'+B)} \quad (25)$$

Where $D' = 1 - D, A = CrRofs, and\ B = \frac{fs}{\omega r}\sin(\frac{D'\omega r}{fs})$

Above-resonance operation (1− (fs/2fr) < D <(fs/2fr)):

In this mode, the duty loss is *D*1+*D*2 as shown in Fig.15In a similar way, the duty loss can be obtained as follows:

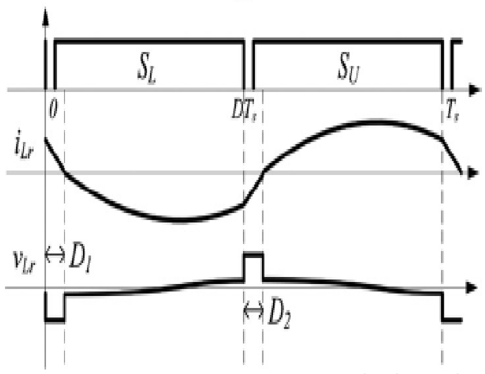


Fig 11 . Above resonance operation with 1-(fs/2fr)>D>fs/2fr

Δ*D* = *D*1 + *D*2

Voltage gain can be obtained by,

$$M = \frac{D'(1-A)+\sqrt{[D'(A-1)^2+4AD'(D'+C)]}}{D'(D'+C)} \quad (26)$$

Where

$$C = 2\frac{fs}{\omega r}\sin(\frac{\omega r}{2fs})\cos(\frac{(D-0.5)\omega r}{fs})$$

***Above-resonance operation* (*D* <1 − (*fs/*2*fr*)):**

In this mode, the duty loss is *D*2 , as shown in Fig.16. Ina similar way, the duty loss can be obtained as follows:

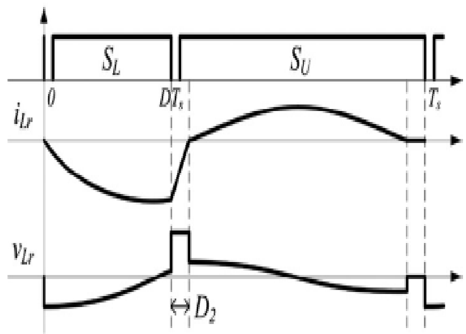


Fig 12 . Above resonance operation with D<1-(fs/2fr)

$$\Delta D = D2 = \frac{(1-D)(fs/\omega r)\sin(D\omega r/fs)}{2crRofs\left(\frac{Vi}{Vo}\right)+(1-D)} \quad (27)$$

Voltage gain of this mode can be obtained by,

$$M = \frac{D^{'}(1-A)+\sqrt{[D^{'}(A-1)^{2}+4AD^{'}(D^{'}+E)]}}{D^{'}(D^{'}+E)} \quad (28)$$

Where $E = \frac{fs}{\omega r}\sin(D\omega r/fs)$

SIMULATION RESULTS:

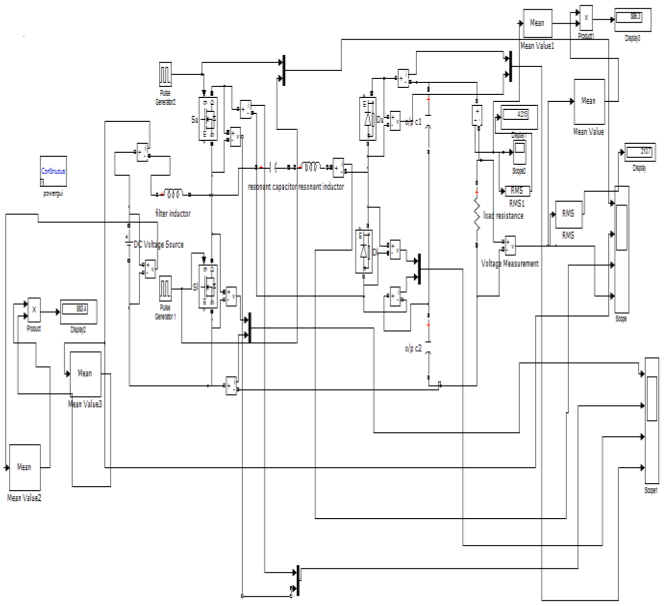
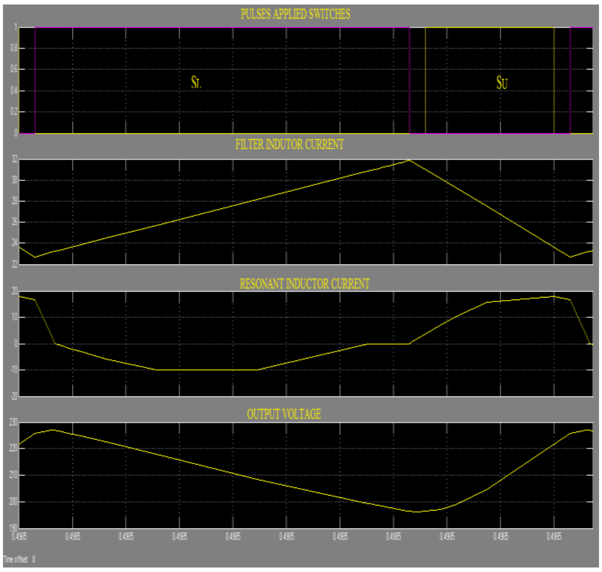


Fig 13. Simulation diagram of proposed converter.



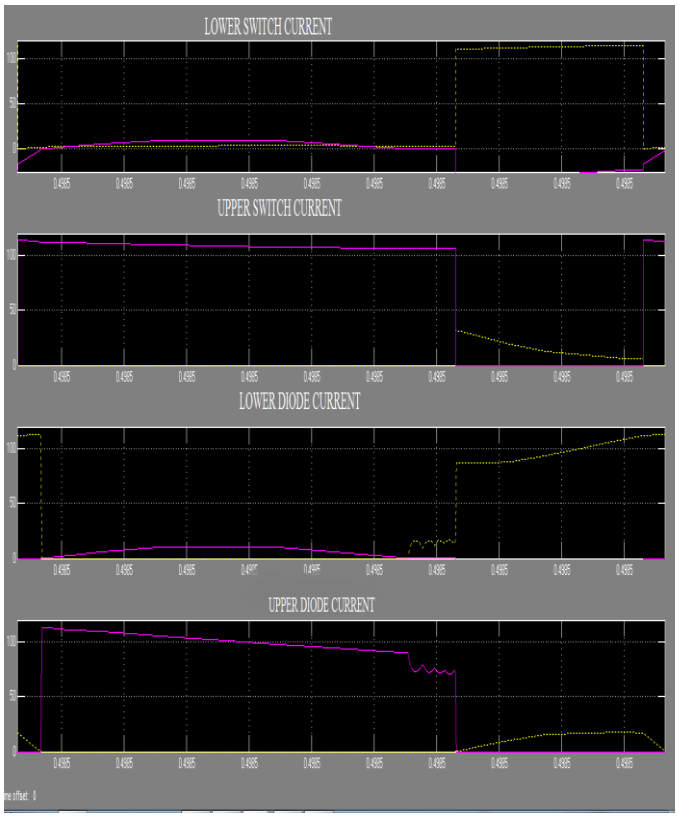


Fig 14. Output waveforms of proposed converter

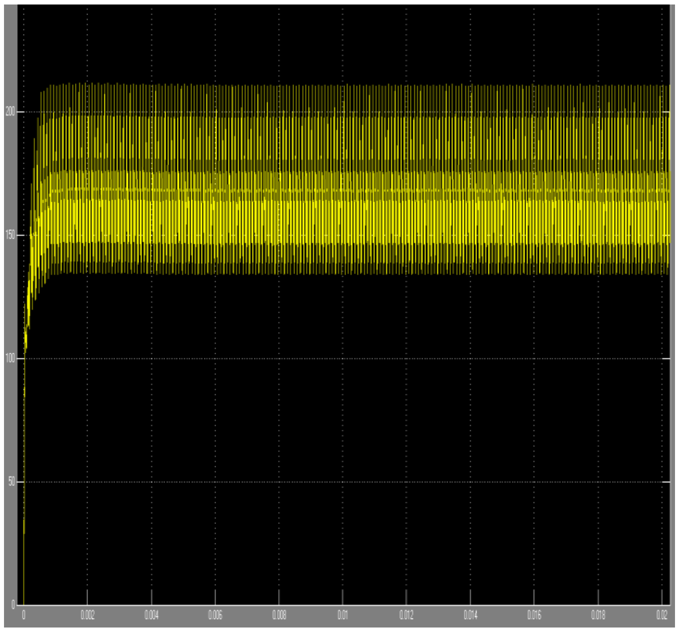


Fig 15. Output voltage waveform.

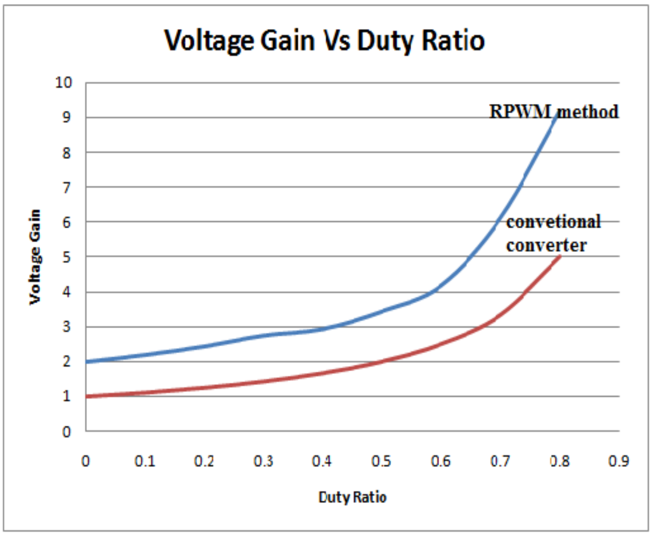


Fig 18. Comparison of voltage gain

PERFORMANCE CURVE:

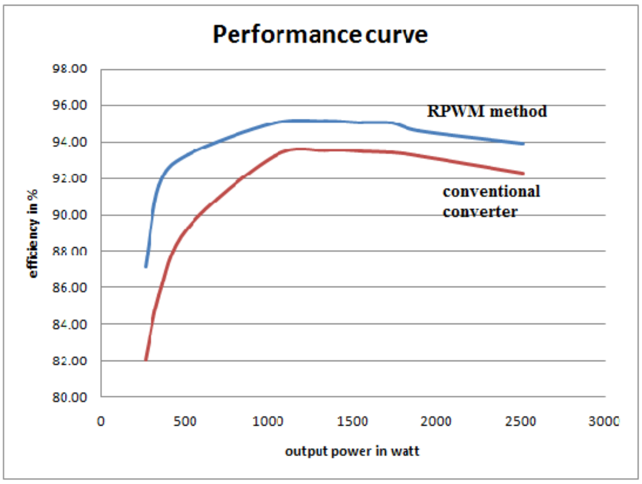


Fig 19. Comparison of performance curve.

V. EXTENSION OF THE PROPOSED CONCEPT:

Fig.20 shows the basic cell used as the building block to build the proposed converter. The basic cell consists of an input filter inductor, a switch leg and a diode leg and an auxiliary inductor and capacitor .

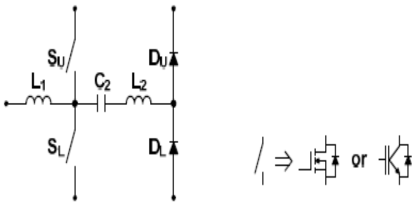


Fig 20. Basic cell

TABLE I

Comparison Table for RPWM and Conventional PWM:

Duty cycle	Proposed RPWM method	Conventional converter
0	2	1
0.1	2.1997	1.1111
0.2	2.4444	1.25
0.3	2.7518	1.4286
0.4	2.9382	1.6667
0.5	3.4519	2
0.6	4.1923	2.5
0.7	6.1403	3.3333
0.8	9.2068	5

Table 1. comparison of voltage gain between conventional converter and proposed converter

N could be increased to get higher output voltage while P could be increased to get higher output power. where N is the number of output series connected basic cell and P is the number of output parallel connected basic cell respectively

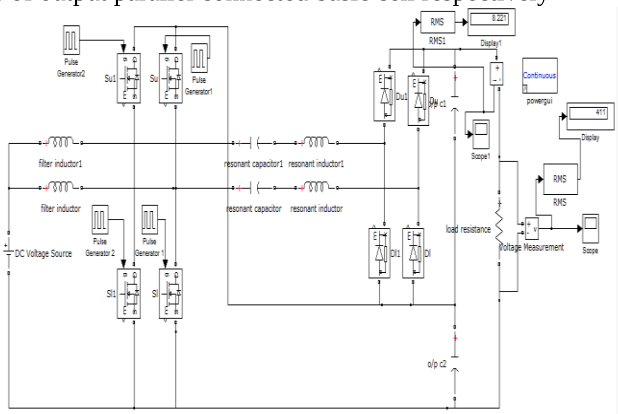


Fig 21.extension of the proposed converter.

VI. CONCLUSION

In this project resonant PWM (RPWM) is proposed for the soft-switched non isolated high step-up dc–dc converters. The following improvements over the PWM method have been achieved: 1) the turn-off losses of the switch are significantly reduced due to reduced turn-off current.2)the switching losses associated with diode-reverse recovery become negligible even in the small duty cycle. 3) The auxiliary capacitor is reduced by 20-fold.4) The duty cycle loss is much reduced resulting in increased step-up ratio.5) The maximum efficiency of RPWM method is 95.3% at 1400W load. The maximum efficiency of PWM method is 94.3% at 1200W load. The efficiency of the proposed RPWM method is approximately 1% higher than that of the PWM method.

REFERENCES:

[1] K. Hirachi, M. Yamanaka, K. Kajiyama, and S. Isokane, "Circuit configuration of bidirectional DC/DC converter specific for small scale load leveling system," in *Proc. IEE Power Convers. Conf.*, Apr. 2002, vol. 2, pp. 603–609.

[2] Q. Zhao and F. C. Lee, "High-efficiency, high step-upDC–DC converters," *IEEE Trans. Power Electron.*, vol. 18, no. 1, pp. 65–73, Jan. 2003.

[3] T. J. Liang and K. C. Tseng, "Analysis of integrated boost-flyback step-up converter," *Proc. IEE Electr. Power Appl.*, vol. 152, no. 2, pp. 217–225, Mar. 2005.

[4] R. J. Wai and R. Y. Duan, "High-efficiency DC/DC converter with high voltage gain," *Proc. IEE Electr. Power Appl.*, vol. 152, no. 4, pp. 793–802, Jul. 2005.

[5] R. J. Wai and R. Y. Duan, "High-efficiency DC/DC converter with high voltage gain," *Proc. IEE Electr. Power Appl.*, vol. 152, no. 4, pp. 793–802, Jul. 2005.

[6] J. U. Duncombe, "Infrared navigation—Part I: An assessment of feasibility," *IEEE Trans. Electron Devices*, vol. ED-11, no. 1, pp. 34–39, Jan. 1959.

[7] B. Axelrod, Y. Berkovich, and A. Ioinovici, "Switched coupled-inductor cell forDC-DC converterswith very large conversion ratio," in *Proc. IEEE32ndAnnu. Ind. Electron. Conf.*, Nov. 2006, pp. 2366–2371E. H. Miller, "A note on reflector arrays," *IEEE Trans.Antennas Propagat*., to be published.

[8] M. S. Makowski, "Realizability conditions and bounds on synthesis of switched-capacitor dc–dc voltage multiplier circuits," *IEEE Trans. CircuitsSyst. I. Fundam. Theory Appl.*, vol. 44, no. 8, pp. 684–691, Aug. 1997.

[9] O. C.Mak,Y. C.Wong, andA. Ioinovici, "Step-upDC power supply based on a switched-capacitor circuit," *IEEE Trans. Ind. Electron.*, vol. 42, no. 1, pp. 90–97, Feb. 1995.

[10] F. L. Luo and H. Ye, "Positive output multiple-lift push-pull switchedcapacitorLuo-converters," *IEEE Trans. Ind. Electron.*, vol. 51, no. 3, pp. 594–602, Jun. 2004..

[11] F. H. Khan and L. M. Tolbert, "A multilevel modular capacitor-clamped DC–DC converter," *IEEE Trans. Ind. Appl*, vol. 43, no. 6, pp. 1628–1638, Nov./Dec. 2007.

[12] D. Cao and F. Z. Peng, "Multiphase multilevel modular DC–DC converter for high-current high-gain TEG application," *IEEE Trans. Ind. Appl.*, vol. 47, no. 3, pp. 1400–1408, May/Jun. 2011.

[13] W. Li and X. He, "High step-up soft switching interleaved boost converters with cross-winding-coupled inductors and reduced auxiliary switch number," *IET Power Electron.*, vol. 2, no. 2, pp. 125–133, Mar. 2009.

[14] W. Li, Y. Deng, R. Xie, J. Shi, and X. He, "Interleaved ZVT boost converters with winding-coupled inductors and built-in LC low pass output filter suitable for distributed fuel cell generation system," in *Proc. IEEEPower Electron. Spec. Conf.*, Jun. 2007, pp. 697–701.

[15] M. Prudente, L. L. Pfitscher, G. Emmendoerfer, E. F. Romaneli, and R. Gules, "Voltage multiplier cells applied to non-isolated DC–DC converters," *IEEE Trans. Power Electron.*, vol. 23, no. 2, pp. 871–887, Mar. 2008.

[16] S. Park and S. Choi, "Soft-switched CCM boost converters with high voltage gain for high-power applications," *IEEE Trans. Power Electron.*, vol. 25, no. 5, pp. 1211–1217, May 2010.

[17] S. L. Talleen. (1996,Apr.). The Intranet Archi-tecture: Man a g in ginfor ma tio n in the n ew paradigm.AmdahlCorp., CA.

[18] S. Park, Y. Park, S. Choi,W. Choi, and K. Lee, "Soft-switched interleaved boost converters for high step-up and high power applications,"*IEEETrans. Power Electron.*, vol. 26, no. 10, pp. 2906–2914, Oct. 2011.