A ZVS DC-DC Converter with Specific Voltage Gain for 3-Phase Load (Induction Motor) Operation

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***Abstract* – In this paper high (specific) voltage gain Zero-voltage-switching (ZVS) dc–dc converter is proposed. The specific voltage is use for inverter operation for supplying 3-phase constant load *i.e.* domestic load, small industrial operation, small entrepreneur etc... It is the combination of boost converter and half-bridge converter and two converters are merged into a single one. The ZVS operation of the power switches and continuous input current is provided by ZVS boost converter and half-bridge converter high voltage gain. The principle of operation and system analysis are presented. The specific voltage is used to obtain a 400 volts Line to Line RMS AC supply through PWM inverter operation. The supply is then used for a 3-Phase load like operation and speed control of 3-Phase Squirrel cage Induction motor as well as single phase load.**

***Index Terms* - Zero voltage switching, Coupled inductor, Specific voltage gain, Soft switching technique, Inverter.**

I. INTRODUCTION

DC–DC converters with high voltage gain are required in many industrial applications such as the front-end stage for the renewable and green energy sources including the solar arrays and the fuel cells, the power systems based on battery sources and super capacitors [1]–[6]. IN dc–dc converters with high voltage gain, there are several requirements such as high voltage gain [2], [3], [5], [6], low reverse-recovery loss [7], [8], soft-switching characteristic [16], low-voltage stress across the switches, electrical isolation, continuous input current, and high efficiency. In order to meet these requirements, various topologies are introduced. In order to extend the voltage gain, the boost converters with coupled inductors are proposed in [9] and [10].

The voltage gain is extended but continuous input current characteristic is lost and the efficiency is degraded due to hard switching of power switches. In [11], a step-up converter based on a charge pump and coupled inductor is suggested. Its voltage gain is around 10 but its efficiency is not high enough due to the switching loss. In [12], a high-step-up converter with coupled inductors is suggested to provide high voltage gain and a continuous input current. However, its operating frequency is limited due to the hard switching of the switches. The converters suggested in [13]–[15] have a similar draw back. Their switching frequencies are limited due to the hard-switching operation. In order to increase the efficiency and power density, soft switching technique is required in dc–dc converters. In various soft-switching techniques are suggested. Generally, there is a trade-off between soft-switching characteristic and high voltage gain. It is because an inductor that is related with soft switching limits the voltage gain. In order to solve these problems, a zero-voltage-switching (ZVS) dc–dc converter with high voltage gain is proposed.

As shown in Fig. 1, it consists of a ZVS boost converter stage to make the input current continuous and provide ZVS functions and a ZVS half-bridge converter stage to provide high voltage gain. Since single power processing stage can be a more efficient and cost-effective solution, both stages are merged and share power switches to increase the system efficiency and simplify the structure. Since both stages have the ZVS function, ZVS operation of the power switches can be obtained with wider load variation. Moreover, due to the ZVS function of the boost converter stage, the design of the half-bridge converter stage can be focused on high voltage gain. Therefore, high voltage gain is easily obtained. ZVS operation of the power switches reduces the switching loss during the switching transition and improves the overall efficiency. The theoretical analysis is verified by an experimental prototype.

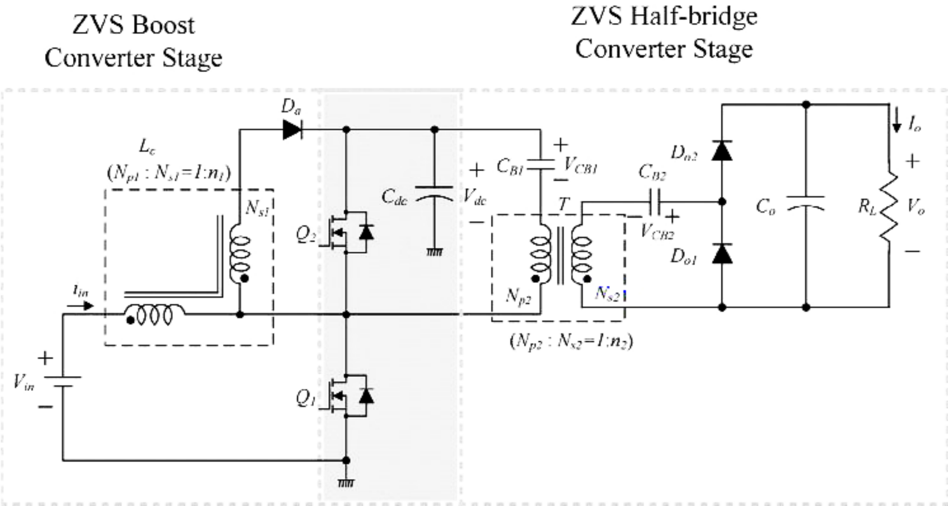

Fig. 1 Proposed Circuit for ZVS dc-dc conversion with specific voltage gain

II. ANALYSIS OF THE PROPOSED CONVERTER

Fig. 2 shows the equivalent circuit of the proposed converter. The ZVS boost converter stage consists of the lower switch Q1, the higher switch Q2, a coupled inductor Lc, Da the auxiliary diode, and the dc-link capacitor C. The intrinsic body diodes of Q1 and Q2 can be represented by DQ1 and DQ2 respectively. C_{Q1} and C_{Q2} represent the parasitic o/p capacitances of Q1 and Q2 respectively. The coupled inductor Lc is designed and takes as the combination of magnetizing inductance Lm1, the leakage inductance Lk1. It has a turn ratio of 1:n1 (n1 = Ns1/Np1). A transformer T, the dc blocking capacitors CB1 and CB2, and the output capacitor Co, the switches Q1 and Q2, the output diodes Do1 and Do2 represents the ZVS half bridge converter stage. The transformer T is designed and takes as the combination of magnetizing inductance Lm2, the leakage inductance Lk2. It has a turn ratio of 1:n2 (n2 = Ns2/Np2). The duty ratio D is based on the switch Q1 and the switches Q1 and Q2 are operated asymmetrically.

The operation period Ts of the proposed converter is divided into seven modes as describe below. The auxiliary diode Da, the upper switch Q2, and the output diode Do1 are conducting before Mode 1. i_{Da}, the auxiliary diode current arrives at its maximum value I_{Da} and i_{m1} the magnetizing current of Lc arrives at its minimum value I_{m12} also i_L has its max. Value I_{L1}.

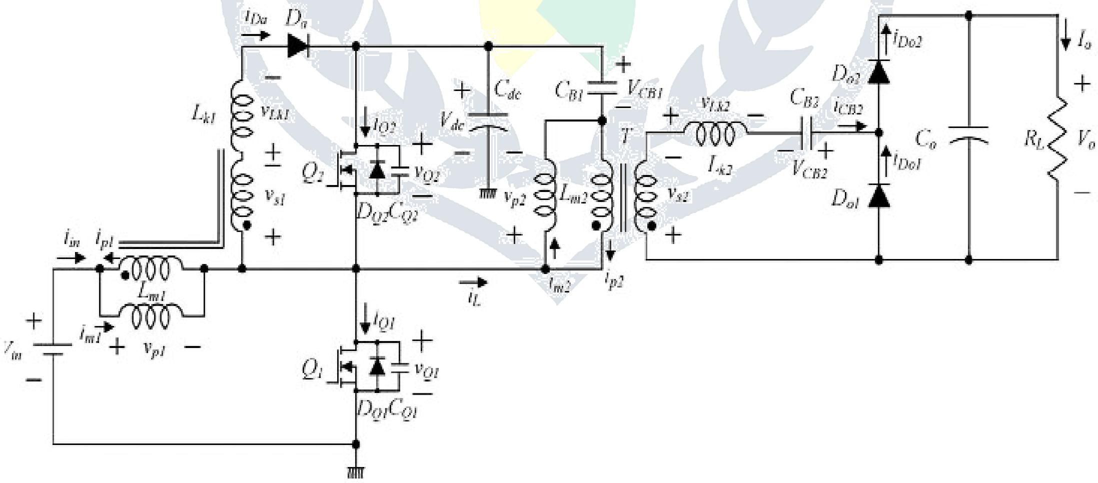

Fig. 2 Equivalent Circuit for ZVS operation

III. MODES OF OPERATION

Mode I [t0, t1]: Upper switch Q2 is turned OFF at time t0. The voltage v_{Q2} across Q2 increases toward Vdc as the capacitor CQ2 starts to be charged. Simultaneously, the voltage vQ1 across Q1 decreases toward zero as the capacitor CQ1 is discharged. With an assumption that capacitances across the switches (CQ1 and CQ2) are very small and all the inductor currents are constant, Tt1 the transition time interval can be considered as follows:

$$Tt1 = t1 - to = \frac{(CQ1 + CQ2)Vdc}{IL1 + (n1 + 1)IDa - Im12}$$

Mode II [t1, t2]: body diode DQ1 is turned ON and, the voltage v_{Q1} becomes zero across the lower switch Q1 at time t1. Then, the gate pulse is applied to Q1. The zero-voltage turn-on of Q1 is achieved as the voltage v_{Q1} is maintained as zero before the switch Q1 is turned ON and current has already flown through the body diode DQ1. Since V_{in} is the voltage v_{p1} across the magnetizing inductance L_{m1}, the magnetizing current i_{m1} increases linearly from I_{m2} its minimum value as follows:

$$\mathrm{im1(t)} = Im12 + \frac{Vin\,(t - t1)}{Lm1}$$

Since – (n1Vin+Vdc) is the voltage v_{Lk1} across the leakage inductance Lk1, linearly decreases in auxiliary diode current i_{Da} from its maximum value I_{Da} is as follows:

$$iDa\,(t) = IDa - \frac{n1\,Vin\,(t - t1)}{Lk1}$$

Mode III [t2, t3]: Current i_{CB2} changes its direction at t2. The diode Do1 is turned OFF as the output diode current i_{Do1} decreases to zero. Then current increases linearly in the output diode Do2 as it turned ON. Reverse-recovery problem of Do1 is alleviated as the current changing rate through it is controlled by the leakage inductance Lk2 of the transformer T. Since v_{p2} is the same as in Mode II and $n2V_{in}+V_{CB2}$–Vo is v_{Lk2}, the diode current i_{Do2} and the current i_{m2} are given by

$$im2\,(t) = Im2\,(t2) - \frac{Vin\,(t - t2)}{Lm2}$$

$$iDo2\,(t) = iCB2\,(t) = \frac{(VCB2 + n2\,Vin - Vo)\,(t - t2)}{Lk2}$$

From Mode II currents i_{p2}, i_L, and i_{Q1} can be obtained. The voltages v_{p1} and v_{Lk} are same in this mode as same to those in Mode II. Therefore, as in Mode II the input current i_{in}, the magnetizing current i_{m1} and the auxiliary current i_{Da} change with the same slopes.

Mode IV [t3, t4]: The auxiliary diode current i_{Da} decreases to zero and the diode Da is turned OFF at t3. Reverse-recovery problem of diode Da alleviated since the changing rate of the diode current i_{Da} is controlled by the leakage inductance Lk1 of the coupled inductor Lc. As the input voltage V_{in} is nothing but the voltage v_{p1} across the magnetizing inductance Lm1, as in Modes II and III the magnetizing current i_{m1} increases linearly with the same slope as given bellow:

$$im1\,(t) = Im1\,(t3) + \frac{Vin\,(t - t3)}{Lm1}$$

The input current i_{in} is equal to the magnetizing current i_{m1} as the auxiliary diode current i_{Da} is zero. The magnetizing current i_{m1} arrives at its maximum value I_{m11} at the end of this mode. The slopes of the current i_{m2}, i_{CB2}, and i_L are not changed because the voltages v_{p2} and v_{Lk2} are not changed in this mode.

Mode V [t4, t5]: Lower switch Q1 is turned OFF at t4. Then, the voltage v_{Q1} across Q1 increases toward V as the capacitor CQ1 starts to be charged. Simultaneously, the voltage v_{Q2} across Q2 decreases toward zero as the capacitor CQ2 is discharged. With the same assumption as in Mode I, Tt2 the transition time interval can be determined as follows:

$$Tt2 = t5 - t4 = \frac{(CQ1 + CQ2)Vdc}{IL2 + Im11}$$

Mode VI [t5, t6]: Voltage v_{Q2} across the upper switch Q2 becomes zero and the body diode DQ2 is turned ON at t5. Then, the gate pulses are applied to the switch Q2. The zero-voltage turn- ON of Q2 is achieved as the voltage v_{Q2} is maintained as zero before the turn-on of the switch Q2 and current has already flown through the body diode DQ2. Since – (Vdc–Vin) is the voltage v_{p1} across the magnetizing inductance Lm1, linearly decreases from its maximum value I_{m11} of the magnetizing current i_{m1} is as follows:

$$im1\,(t) = Im11 - \frac{(Vdc - Vin)\,(t - t5)}{Lm1}$$

Since n1 (Vdc–Vin) is nothing but the voltage v_{Lk1} across the leakage inductance Lk, $i_{Da,}$ the auxiliary diode current linearly increases from zero as follows:

$$iDa\,(t) = \frac{n1\,(Vdc - Vin)(t - t5)}{Lk1}$$

Mode VII [t6, t7]: The current i_{CB2} changes its direction at time t6. The output diode Do2 is turned OFF as current i_{Do2} decreases to zero through it. Then the output diode Do1 is switched ON and its current passing through it increases linearly. Similar to Do1, reverse recovery problem is alleviated for Do2 as the current changing rate of Do2 is controlled by the leakage inductance Lk2 of the transformer T. Since v_{p2} is the same as in Mode VI and v_{Lk2} is V_{CB2}– n2DVin/ (1–D), the current i_{m2} and the diode current i_{Do2} are given by

$$im2\ (t) = Im2\ (t6) + \frac{DVin\ (t - t6)}{Lm2\ (1 - D)}$$

$$iDo1\ (t) = -iCB2\ (t) = \frac{\left(VCB2 - \frac{n2\ DVin}{(1 - D)}\right)(t - t6)}{Lk2}$$

From the same relations in Mode II the currents i_{p2} and i_L can be obtained. In this mode, the voltages v_{p1} and v_{Lk1} are same as to those in Mode VI. Therefore, the auxiliary current i_{Da}, the magnetizing current i_{m1}, and the input current i_{in} change with the same slopes as in Mode II. Fig.6 shows the final output of combined stage operation.

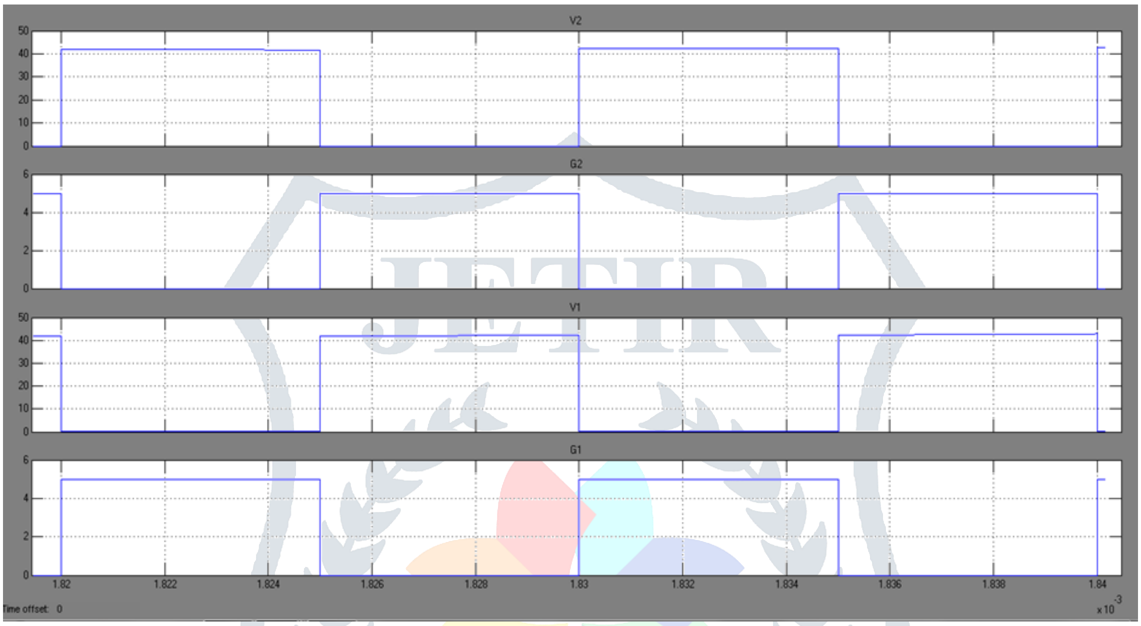

Fig. 3 Waveform of Switching of MOSFET at 100 kHz Zero voltage Switching

IV. INVERTER OPERATION AND ANALYSIS

The adoption of AC power has created a trend where most devices adapt AC power from an outlet into DC power for use by the device. However, AC power is not always available and the need for mobility and simplicity has given batteries an advantage in portable power. Thus, for portable AC power, inverters are needed. Inverters take a DC voltage from a battery or a solar panel as input, and convert it into an AC voltage output. There are types of DC/AC inverters available which are classified by their output type. Pure sine wave inverters offer more accuracy and less unused harmonic energy delivered to a load, but they are more complex in design and more expensive.

Pure sine wave inverters will power devices with more accuracy, less power loss, and less heat generation. Pure sine wave inversion is accomplished by taking a DC voltage source and switching it across a load using an H-bridge. If this voltage needs to be boosted from the DC source, it can be accomplished either before the AC stage by using DC-DC boost converter, or after the AC stage by using a boost transformer. The inverted signal itself is composed of a pulse-width-modulated (PWM) signal which encodes a sine wave shown in figure 4. The duty cycle of the output is changed such that the power transmitted is exactly that of a sine-wave. This output can be used as-is or, alternatively, can be filtered easily into a pure sine wave. This report documents the design of a true sine wave inverter, focusing on the inversion of a DC high-voltage source. It therefore assumes the creation of a DC-DC boost phase.

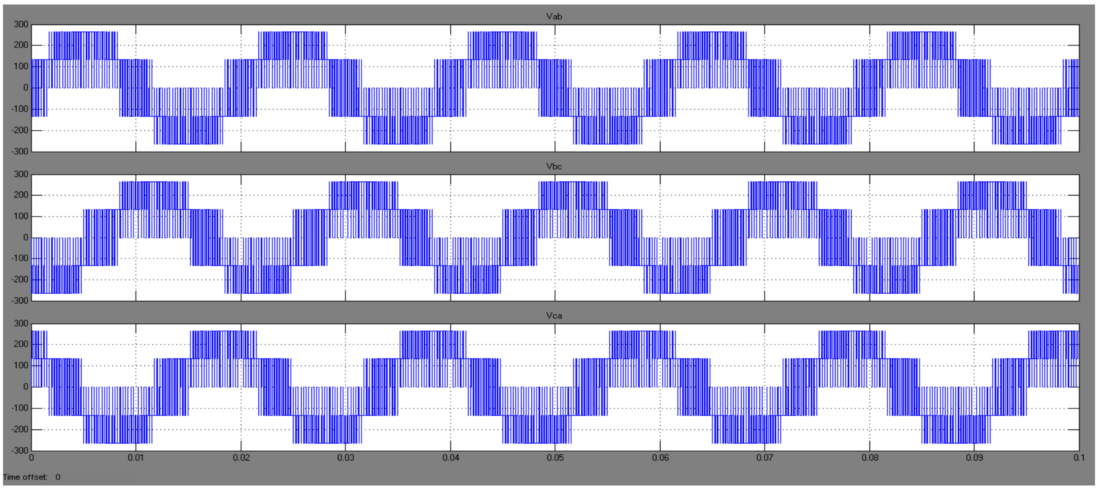

Fig. 4 PWM Inverter Output for 400v dc Input

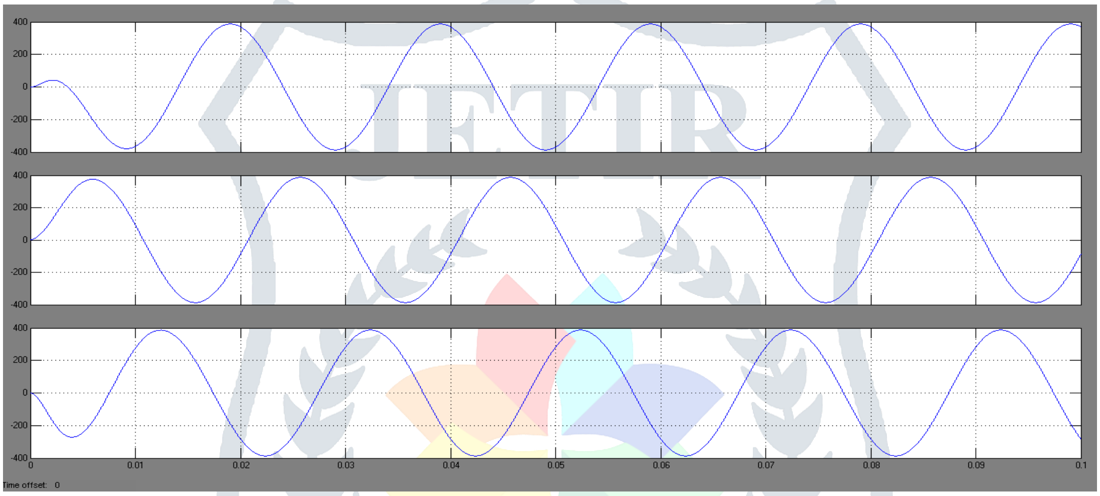

Fig. 5 Line to Line Output Voltage of Inverter

V. INDUCTION LOAD

A 3-Phase Induction motor is used as test load in this module. The figure 5 shows the stator current Iabc, speed, torque waveform of the 3-phase Induction motor respectively from total module output and following are 3-phase Induction motor parameters for that:

3-PHASE INDUCTION MOTOR PARAMETER

Rated power	5X746 = 3730 watts (5 HP)
Speed	153.12 rad./sec
Full load torque	24.35 N-m.
Rated Voltage	400 line-to-line(rms)
Frequency	50 Hz

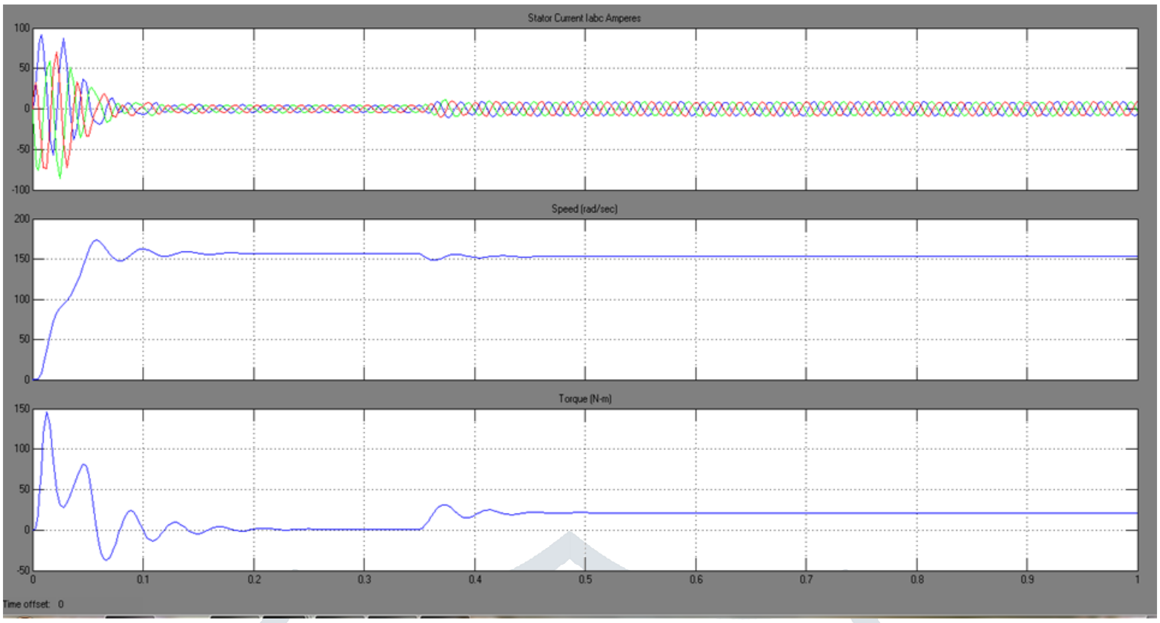

Fig .5 Final O/P waveform of Induction Motor stator current Iabc, speed & torque respectively

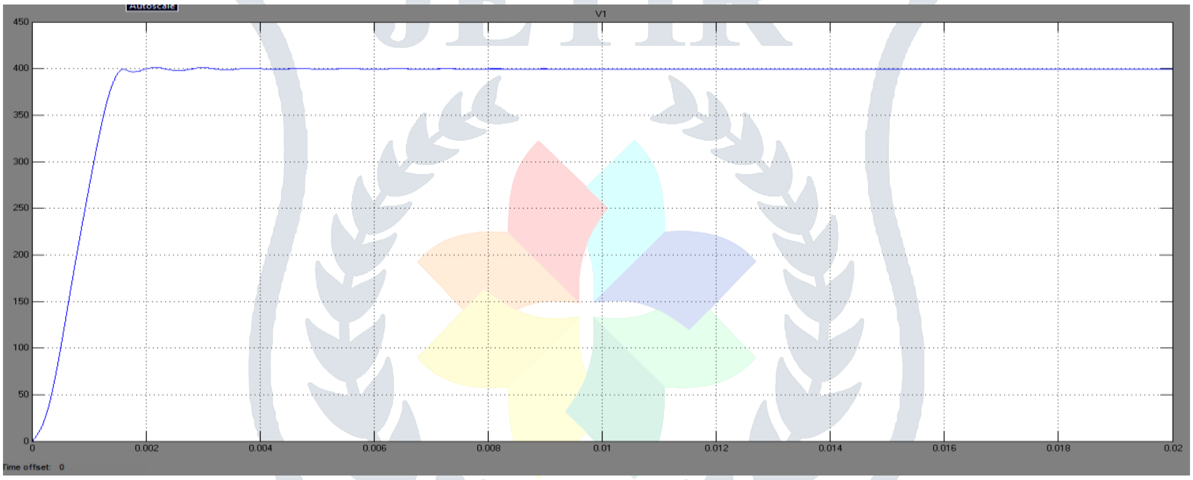

Fig.6 Constant Output Voltage from DC-DC Converter

VI. Acknowledgement

A ZVS dc–dc converter with high voltage gain was suggested. It can achieve ZVS turn - ON of two power switches while maintaining CCM. In addition, the reverse-recovery characteristics of the output diodes were significantly improved by controlling the current changing rate with the use of the leakage inductance of the transformer. The proposed converter presents a higher efficiency and a wider ZVS region compared to other soft switching converters due to the ZVS boost converter stage. . The output of the ZVS boost converter is fed to the inverter for AC output at final stage.. The AC supply is fed to a 3-Phase Squirrel cage Induction motor for operation and speed control to obtain a constant speed and torque.

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