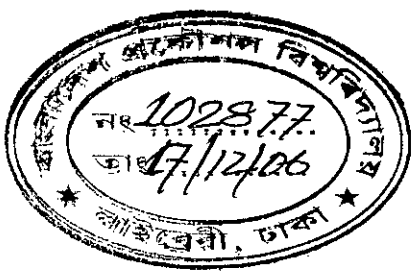
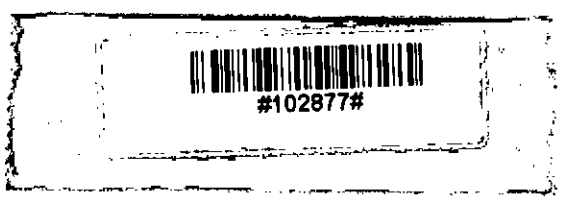
Improvement of Input Side Currents of a Three Phase Rectifier Combining Active and Passive Filters

Amina Hasan Abedin





DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
BANGLADESH UNIVERSITY OF ENGINEERING AND TECHNOLOGY
2006

Improvement of Input Side Currents of a Three Phase Rectifier Combining Active and Passive Filters

A thesis submitted to
the department of Electrical and electronic engineering of
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for the degree of

MASTER OF SCIENCE IN ELETRICAL AND ELECTRONIC ENGINEERING

By

Amina Hasan Abedin

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The thesis entitled "**Improvement of Input Side Currents of a Three Phase Rectifier Combining Active and Passive Filters**" submitted by Amina Hasan Abedin, Roll No: 040206113P, Session: April 2002 has been accepted as satisfactory in partial fulfillment of the requirement for the degree of MASTER OF SCIENCE IN ELECTRICAL AND ELECTRONIC ENGINEERING on June, 2006.

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Dept. of Electrical and Electronic Engineering
BUET, Dhaka-1000, Bangladesh

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Dr. Kazi Khairul Islam — **Member**
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Dept. of Electrical and Electronic Engineering
IUT, Board Bazar, Gazipur, Bangladesh.

Declaration

It is hereby declared that this thesis or any part of it has not been submitted elsewhere for the award of any degree or diploma.

Signature of the candidate

Amina Hasan Abedin

(Amina Hasan Abedin)

Dedication

To my Country
The People's Republic of Bangladesh

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Abstract

In order to reduce the Total Harmonic Distortion (THD) in a three phase boost rectifier, active switching and passive filters are incorporated in this work. A constant frequency switching is used for active filtering and pulse width modulation is used to regulate the output voltage. An input Electro Magnetic Interference (EMI) filter is used to suppress the high frequency component generated by the active switching. Moreover, a series LC filter resonating with the supply frequency is also used to suppress the low frequency component that may pass to the input. With addition of this series LC filter, the THD value could be made less than three percentages, which is a great improvement over the earlier rectifiers that have only EMI filter. In the earlier types of rectifiers, the THD value was as high as 17%.

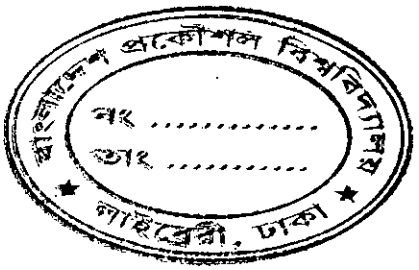
The efficiency of the module is also studied. As the output voltage has nonlinear relation with duty cycle, the efficiency is also nonlinear with output voltage variation. But up to certain range of duty cycle it could be made linear in nature with output voltage.

The input LC series filter and the EMI filter design are discussed. The efficiency versus output voltage, THD versus duty cycle curve for the proposed rectifier circuit is given for clear understanding of the model.

ORCAD 9.1 release Version, very powerful Electrical and Electronics design software is used to design the rectifier circuits. A lot of efforts had to be done to get the ultimate results by simulation.

Chapter 1

INTRODUCTION



1.1 Introduction

Stable and regulated dc power supplies for the electrical appliances are provided by ac-dc conversion from the mains. The most common method of conversion from ac to dc is to use diode bridge rectifiers followed by input capacitors as power stage. The operation of the diode bridge rectifier is nonlinear and the capacitor used to reduce the ripple of the output voltage causes input current to be drawn from the line over very short period of time and hence to be pulsating in nature. As a result, input currents are non-sinusoidal in nature and contain harmonics [1]. Neither Harmonic nor reactive current flowing through a system provides useful power. The power infrastructure has to carry these currents that create a number of problems [1-3] in the sensitive electronic equipment and in the power distribution network which include

a) Heat loss due to I^2R drop in wire and higher flux in transformer iron,
b) Input ac mains voltage distortion because of the associated high peak currents especially with the soft sources,
c) Lower rectifier efficiency for the large r.m.s values of input current,
d) Lower power conversion reliability,
e) Excitation of system resonances,
f) Increasing Volt-Ampere ratings of the utility equipments, such as generator, transmission lines and transformers and
g) Malfunctioning of the sensitive electronic equipments.

Economic and safety concerns along with new regulations to maintain the integrity of the power distribution system has created an acute interest in power quality strategies. The recommended practice, IEEE-519 and IEC 1000-3 have evolved to maintain utility power quality at acceptable levels. In order to meet these standards a cost effective and economical solution to mitigate harmonics generated by power electronic equipment is

currently of high interest. Various active and passive techniques are being carried out to meet the standers. Passive techniques, which introduce a filtering stage consisting of inductor and/or capacitors, to reduce low frequency harmonics, are attractive for their simplicity and reliability. But size and weight are their major drawbacks [4-5]. Active techniques on the other hand, use a high switching frequency converter that shapes the input currents almost sinusoidal with small harmonic contents [6]. Although various research efforts have produced several power circuit configurations for Power Factor Corrections (PFC) and Total Harmonics Distortions (THD) has alleviated the problem to some extent, they greatly increased the cost and complexity comparing to conventional low quality rectifiers. Thus the research is still being carried out to optimize cost and complexity.

1.2 Background

Research works on PFC with Reduces THD is currently of high interest. Many solutions have been proposed using passive or active approaches, these last employing high or low frequency commutation.

Passive filters, besides do not allow the regulation of the output voltage, decrease its value in comparison with the unfiltered rectifier. Taking the harmonic limits as a quality index, the resulting inductors are typically larger than the ones used in high quality rectifiers using active circuits [7].

That is why the use of active filter is getting increasing interest. One approach is to use three single-phase power factor corrected rectifiers in cascade [8]. The main advantage of this configuration is that a well-known single-phase power factor correction (PFC) technique can be used in three phase applications. But this increases component count and complicates the input synchronization logic. The switching frequency is load dependent and owing to the variation in power circuit control parameters among the three individual converters a complete triplen harmonic elimination from the input line current cannot be achieved. Moreover, the main advantage of using a three phase inverter and transformer for better transformer core and copper utilization cannot be achieved.

Another simple approach to reduce THD is to use harmonic injection method [9-10]. The injected signal modifies the duty cycle of the rectifier switch to meet IEC1000-3 requirement. Nevertheless if the phase of the injected signal is not well synchronized with the fifth order harmonic of the input current, the expected values may not be possible to achieve.

Another approach is to use boost topology [1] in PFC applications provided that dc output voltage is close to but slightly greater than the peak ac input voltage [11]. A boost converter designed for universal input PFC applications is heavily oversized compared to converters designed for a narrow range of input line voltages [12].

Another boost topology is to use six switches full bridge rectifier [13-14], which gives continuous input current, excellent power factor and low switch current ratings. However this circuit is very complicated in power stage and control and too expensive for medium power level applications. The active semiconductor utilization is also low [15].

Another boost PFC method implies the switch to be turned on and off only twice per line period and these are called line frequency commutated rectifiers [16-18]. This result in limited di/dt and dv/dt and switch losses, allow use of slow recovery diodes and avoids the need for heavy EMI filters. As certain amount of boost effect is present in low frequency topologies where the energy processed during the switch on interval is transferred to the output capacitance partially compensating for the input inductance voltage drop. But the maximum allowed switch on time is limited to keep the switch current stress at an acceptable level. As a consequence, only a limited output voltage regulation can be achieved.

One more approach for improving THD of input currents involves controlling the average current to a constant level in the boost diode [19, 20]. In Average Current Mode Control (ACMC), the duty cycle must be modulated over the line cycle that results an improved

input current waveform. Advantages of ACMC includes large noise margin, no requirement for additional slope compensation, easy current limit implementation, excellent voltage and current regulation, good behavior in both continuous and discontinuous inductor current modes and inherent input and output voltage feed-forward properties. But the drawback of this method is the extra current sensor required to control the average boost diode current.

Other techniques like varying duty cycle of PWM also called divider approach, multiplier approach and operation at Continuous Conduction Mode (CCM) and Discontinuous Conduction mode (DCM) boundary that use variable switching frequency use input voltage sensing. These can give high quality input current at the cost of either complicating the control circuitry and/or using mathematical operations such as division, multiplication and square root. These make the analog implementation complicated and costly [20].

That is why, single switch boost rectifier that uses six diodes and a single switch are preferred for simple and low cost applications. Moreover efficiency of this topology is very high. However, if a discontinuous conduction mode (DCM), pulse width modulated (PWM) [1,9] rectifier is implemented with conventional constant frequency low bandwidth output voltage feedback control the input current exhibits a relatively large fifth harmonic. This leads to operate the single switch boost rectifier in critical mode [21, 22] where the switching frequency is variable to reduce THD of input current .The drawback of this circuit is the wide range switching frequency variation that depends upon both load and input voltage limits. That is why in our thesis we choose constant frequency PWM in single switch boost rectifier and to solve the harmonics problem along with active approach, we had also introduced passive filters in the circuit design to get approximately sinusoidal input current wave shapes.

1.3 Objective of the work

The objective of this thesis is to make the input current of a three-phase rectifier circuit to be nearly sinusoidal and at the same time in phase with the supply potentials. This eliminates the harmonic contents in the input current and improves the power factor of the circuit. In doing so, the efficiency of the module is also improved. As a whole, the size and ratings of the equipments has been minimized. The aim of the study is to investigate an integrated power quality improvement by simultaneous active and passive filtering technique.

In this work, the appropriate switching frequency and solid-state switch has been selected to shape the input current. But this leads to introduce very high frequency harmonics in the wave. To eliminate these high frequency components in the input, a simple filter with very small rating inductor and capacitor is used to bypass these high frequency components. However, even after use of input filter low frequency components still exists in the input current.

That is why another series LC filter has been set up at the input side. The component values of this input filter is modeled in such a way that they would resonate at the supply frequency so that the power factor of the circuit remains near unity. All of this modeling has been done by using the simulation and the mathematical expression for the acceptable range of values of the components are found out.

1.4 Thesis Layout

Organization of this thesis includes five chapters. Chapter one gives a general introduction followed by background and objective of the work.

Chapter two reviews different concepts that are needed for active and passive control of three-phase boost rectifier circuit, to enhance the knowledge.

In chapter three, different rectifier circuits with passive and active filters are analyzed to find out a perfect solution of our research. Simulation of each circuit is also provided.

Chapter four includes the proposed scheme that has to be implemented. It has four stages. First stage is a resonating low frequency filter; second stage has high frequency filter designed by high frequency transfer function, third stage is an IGBT switch that is driven by a separate switching module and the last and forth stage is load with capacitive filter for reducing the output voltage ripples.

Conclusive discussions and remarks are drawn in chapter five. Some suggestions leading to future scope of work are also presented here.

An Appendix is included at the end as chapter six. It describes a simple method of steady state analysis of PWM converters.

Chapter 2

Basics of Three-Phase Rectifier

2.1 Introduction

When both voltage and current are ideal sinusoidal in nature then the power factor,

$$PF = \cos\theta \qquad 2.1$$

where θ is the angle between voltage and current waveforms. However, most power supplies draw a non-sinusoidal current. When the current is not sinusoidal and the voltage is sinusoidal, the power factor consists of two factors:

a) The displacement factor related to the phase angle and
b) The distortion factor related to wave shape.

The following equation represents the relationship of the displacement and distortion factor as it pertains to the power factor.

$$P.F = \frac{I_{rms\,(1)}}{I_{rms}} \cos\theta \qquad 2.2$$

where, $I_{rms(1)}$ is the fundamental component of current and I_{rms} is the rms value of current.

Therefore the purpose of the PF correction circuit is to minimize the input current distortion and make the current in phase with the voltage. When the P.F is not equal to unity, the current waveform does not follow the voltage waveform. This result not only in power losses but also cause harmonics that travel down the neutral line and disrupt other

devices connected to the line. The closer the power factor is to unity the closer the current harmonics will be to zero since all the power is contained in the fundamental frequency.

2.2 Problem with SMPS

One problem with Switch Mode Power Supplies (SMPS) is that they do not use any Power Factor (PF) correction and that the input capacitor C_{in} will only charge when V_{in} is closer to V_{PEAK} or when V_{in} is greater than the capacitor voltage V_{cin} as shown in Figure 2.1. If C_{in} is designed using the input voltage frequency, the current will look much closer to the input waveform (load dependent), however any little interruption on the mainline will cause the entire system to react negatively. As a result in designing SMPS the hold up time for C_{in} is designed to be greater than the frequency of V_{in}, so that if there is a glitch in V_{in} and a few cycles are missed C_{in} will have enough energy stored to continue to power its load.

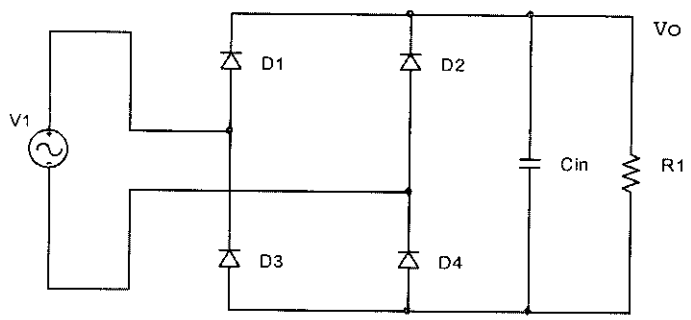


Figure 2.1: SMPS without PFC

Theoretically $V_{cin}(t)$ will retain its voltage if connected to light loads. As load increases, there will be more droop from $V_{cin}(t)$ between subsequent peaks, but only a small percentage with respect to the overall V_{in} . This is shown in Figure 2.2.

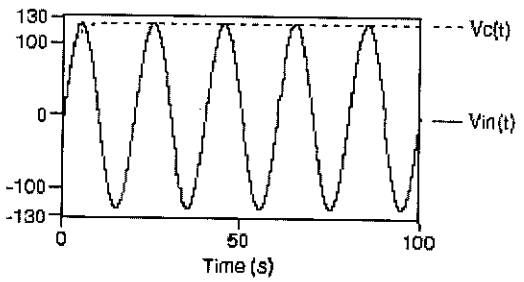


Figure 2.2: V_{in} with charging C_{in}

As stated before, C_{in} will only charge when V_{in} is greater than its stored voltage, means a non PFC circuit will only charge C_{in} a small percentage of the overall cycle time. After 90^{o}, the half cycle from the bridge drops below the capacitor voltage V_{cin}, which back biases the bridge, inhibiting current flow into the capacitor via V_{in}. As a result input current spikes are generated in the inductor. This is shown in Figure 2.3.

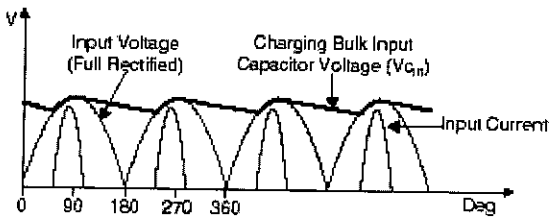


Figure 2.3: Voltage ands current waveforms in a simple rectifier circuit

All the circuitry in the supply chain must be capable of carrying this huge peak current. During these short periods C_{in} must be fully charged, therefore large pulses of current for a short duration are drawn from V_{in}. There is a way to average this spike out so it can use the rest of the cycle to accumulate energy, in essence smoothing out the huge peak current, by using power factor correction.

In order to follow V_{in} more closely and not have these high amplitude current pulses, C_{in} must charge over the entire cycle rather than just a small portion of it. Today's nonlinear loads make it impossible to know when a large surge of current will be required, so keeping the inrush to the capacitor constant over the entire cycle is beneficial and allows much smaller capacitance to be used. This method is well known as power factor correction.

2.3 Use of Boost converter for PFC

For active power factor correction boost converter topology is widely used in both continuous and discontinuous mode. From Figure 2.4, 2.5, 2.6 and 2.7 the operation of boost converter is described below. Initially the inductor is assumed to be uncharged, so that $V_o = V_{cin} = V_{in}$. When the switch closes the current gradually increases linearly since

$$I_L = \frac{1}{L}\int V_L dt \qquad 2.3$$

Voltage across inductor increases exponentially until it comes to V_{in} .The current will flow in reverse direction and decrease from I_{max} to zero as the switch is OFF and thus change the polarity across the inductor.

$$V_L = L\frac{di}{dt} \cong L\frac{\Delta i}{\Delta t} \qquad 2.4$$

If the inductor would be ideal then the voltage will approach negatively infinity. Actual voltage value has limited by the series connected resistance with the inductor. With the switch open and the inductor discharging, the voltage across it reverses and becomes additive with the source voltage V_{in}. A diode and capacitor connected to the output would charge to this high voltage. This is how the converter boosts the voltage.

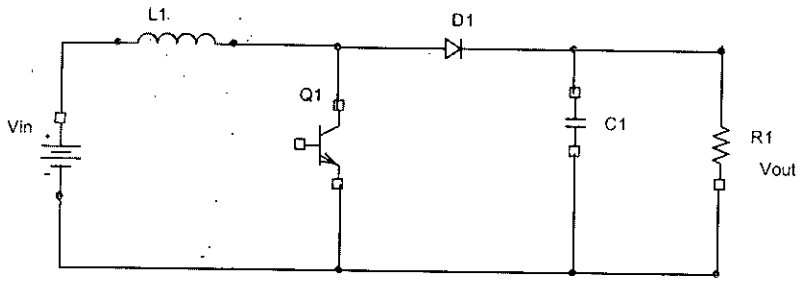


Figure 2.4: Boost converter circuit with BJT switch

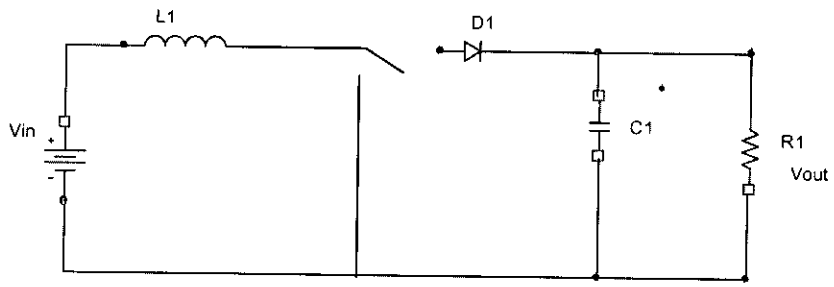
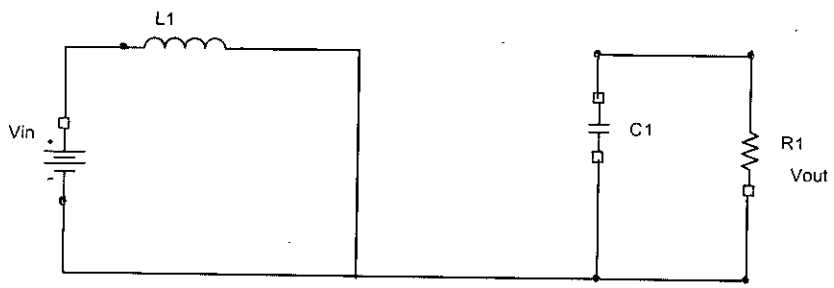


Figure 2.5: Boost converter circuit with SPDT switch

At $T=T_{ON}$



At $T=T_{OFF}$

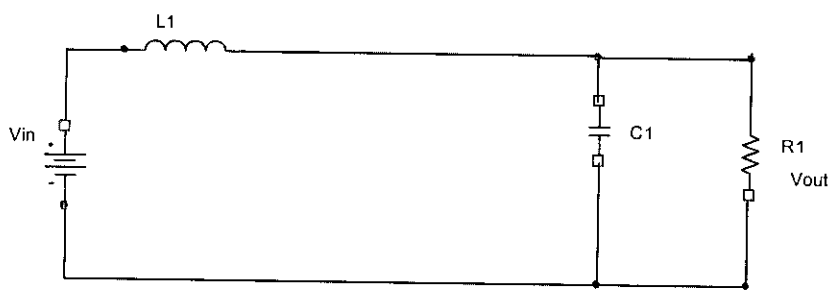


Figure 2.6: Equivalent circuit of boost SMPS at ON and OFF time of the switch

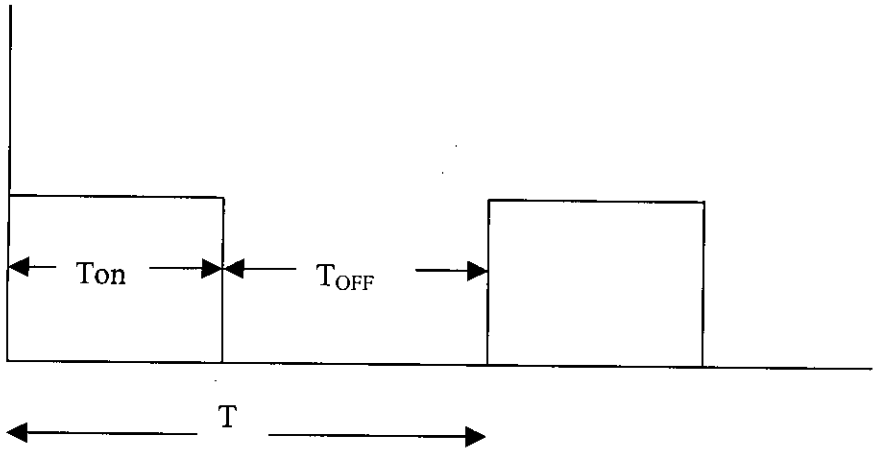


Figure 2.7: Switching time period, $T=T_{ON}+T_{OFF}$

The boost converter must meet two simultaneous conditions:

1) The output voltage of the boost converter must be set higher than the peak value of the line voltage
2) The current drawn from the line at any given instance must be proportional to the line voltage.

Having the boost pre-converter voltage higher than the input voltage forces the load to draw current in phase with the ac main line voltage that in turns rid harmonic emissions.

2.4 Modes of PFC Operation

There are two modes of PFC operation:

a) Discontinuous: In this mode the boost converter's switch is turned ON when the inductor current reaches zero, and turned OFF when the inductor current meets the desired input reference voltage. In this way the input current waveform follows that of the input voltage, therefore attaining a power factor close to unity.

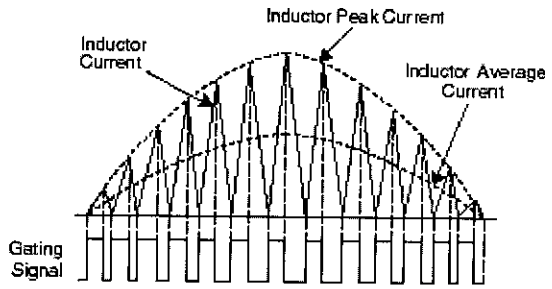


Figure 2.8: Discontinuous mode of operation

For low power level application like less than 300W this mode is preferable for SMPS. In comparison with continuous mode devices, discontinuous ones use larger cores and have higher I^2R and skin effect losses due to the large inductor current swings. As the swing increases, a larger input filter is also required. However, it also offers a number of advantages over its CCM counterpart such as simpler transfer function, ease of control, zero current turn on, minimum diode reverse recovery current and reduced inductor size. For this reason, less expensive diodes can be used. Hence DCM converters are increasingly used in applications such as UPS and battery charger and in applications where size, weight and cost are major concerns [20].

How switching reduces the harmonics content

Since the boost rectifier is operated in Discontinuous Conduction Mode (DCM) with constant frequency and constant duty cycle, all three phase currents i_a, i_b and i_c are zero at the end of a switching period immediately before boost switch is turned ON. Four operating regions can be identified in a switching period as shown in Figure 2.9 and 2.10.

1st operating region: The power switch is turned on to linearly charge the input inductors according to the phase voltage that is applied across each one.

2nd operating region: The power switch is turned OFF to reset the inductors, the inductor with the lowest peak current reset first.

3rd operating region: The two remaining inductor currents are reset to zero at the same rate.

4th operating region: Once the reset interval is finished, the output load is supplied by the energy stored in the output capacitor until the next switching period restart.

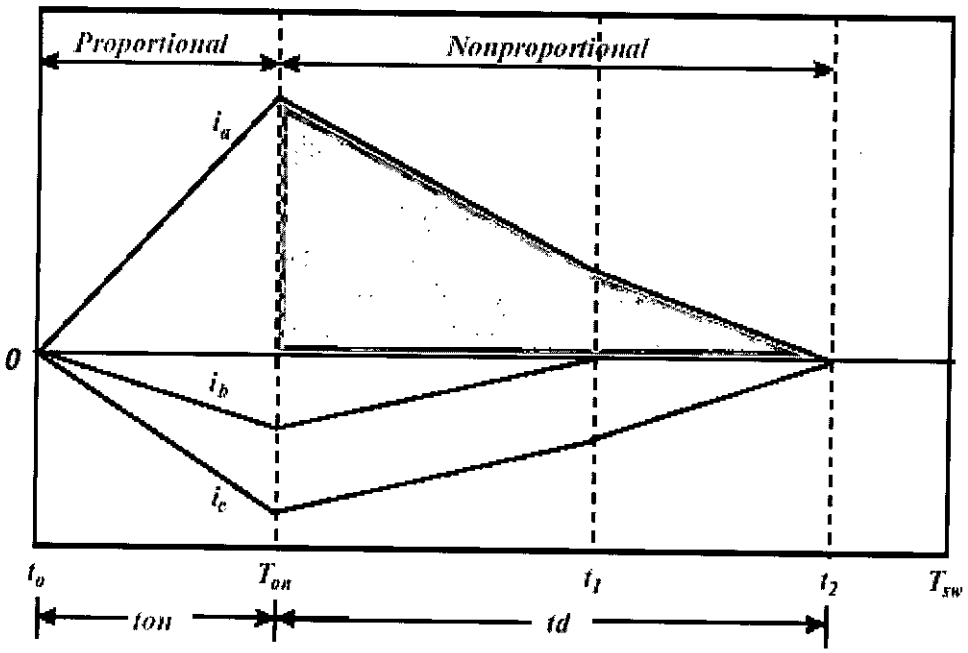


Figure 2.9: The details input inductor currents over one switching cycle

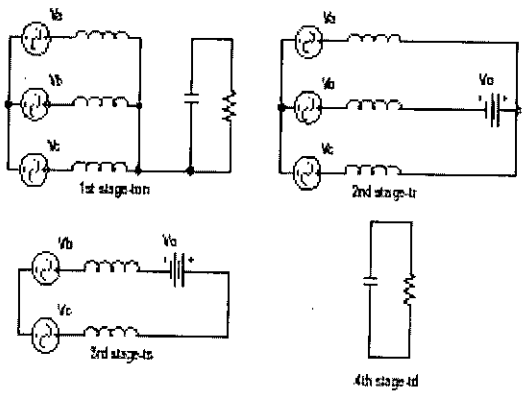


Figure 2.10: Operating stages of the Rectifier

The switching frequency is much higher than the line frequency. The well-balanced and undistorted three phase input voltages can be constant within each switching cycle. Therefore, by averaging the input current in a switching period, its higher orders harmonics are filtered out.

b) Continuous: This mode is preferred for power levels greater than 300W.The semi-conductor switch does not switch on when the boost inductor is at zero current, instead the current in the energy transfer inductor never reaches zero during the switching cycle.

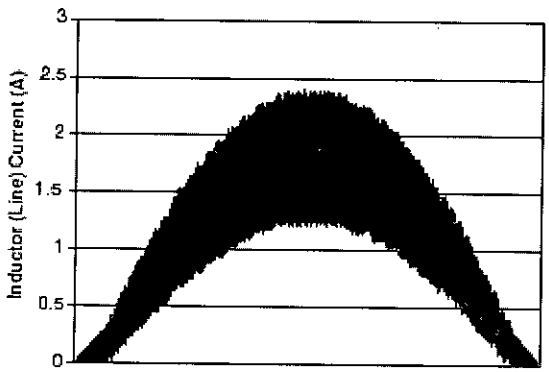


Figure 2.11: Continuous mode of operation

The voltage swing is less than in discontinuous conduction mode, resulting in lower I^2R losses, lower ripple current and hence lower inductor core losses. Less voltage swing also reduces EMI and allows for a smaller input filter to be used. Since the switch is not being turned ON when the boost inductor's current is at zero, a very fast reverse recovery diode is required to keep losses to a minimum.

2.5 Placement of Inductor

Single switch rectifier can place the inductor on the ac side, "ac-link inductor", or the dc side, "dc-link inductor", of the rectifier. With the exception of high per unit current demands and low output voltages, the performance of the rectifier is not affected by placing the inductor on either side of the rectifier. For a rectifier using dc-link inductor, positive unipolar PWM is used in both half cycles of the ac supply. For a rectifier using ac-link inductor, positive unipolar PWM is used in the positive half cycle of the ac supply and negative unipolar PWM is used in the negative half cycle.

2.6 Frequency Modulation Techniques

The switching frequency used can be constant or variable.

Constant Frequency operation: The switching frequency is kept constant and the ON time is varied. The width of the pulse is varied and this type of control is known as pulse-width-modulation (PWM) control.

Variable Frequency operation: If variable, the frequency can be controlled or to be set free to vary within set limits. A circuit using variable switching frequencies can result in lower EMI and lower power losses but these types of control would generate harmonics at unpredictable frequencies and hence the filter design would be difficult and the topology is harder to analyze.

2.7 Switch Stress

The total switch stress is defined as the product of the switch blocking voltages and peak current, summed over all active switches in the converter. The switch stress is a measure of the total active silicon area required for realization of the converter. And silicon

utilization is defined as the converter output power divided by the total switch stresses [15].

A six-switch bridge network leads to poor silicon utilization since the semiconductor devices are effectively utilized only near the peaks of the applied ac phase current waveforms.

The single switch approach utilizes the active semiconductor devices more effectively. Zero current switching of the active semiconductor devices and zero voltages switching of diodes can be obtained.

For example, if the six IGBT devices of a three phase bridge module were reconnected in parallel to form a single device and then operated in an equivalent single switch Zero Current Switching (ZCS) rectifier, then the peak current density in each silicon device would be reduced by a factor of one-third.

Again increasing the switching frequency leads to higher switching losses and electromagnetic interference noises. In order to improve the efficiency of the PFC circuit many efforts have been done on the soft switching converter. Soft switching techniques allow operation with much reduced switching losses and stresses enabling high switching frequency operation for improved power density with high efficiency. In general, the soft switching approaches can be classified into two groups:

(a) Zero Voltage Switching (ZVS) approach
(b) Zero Current Switching (ZCS) approach

The choice depends on the semiconductor devices to be used. The ZVS approaches are generally recommended for MOSFETs. On the other hand, ZCS approaches are effective for IGBTs since the turn off switching losses caused by the tail current is the major part of the total switching losses [23-25].

2.8 Preference of IGBT over MOSFET for switching

When using constant current ripple, increasing the switching frequency allows a reduction in the value of the boost inductor. However, increasing the switching frequency will lead to increase switching losses in the power semiconductors. In the standard boost PFC circuits, conduction losses in the power switch will be lower than the switching losses, and consequently the switching frequency will be limited by the switching losses of the chosen power transistor and the recovery losses of the boost diode.

In recent years IGBTs are preferred for high power applications, since IGBTs have a higher voltage rating, higher power density and lower cost compared with MOSFETs. However IGBTs are relatively slow in switching speed, so the switching losses and the high frequency operation are two well-known problems [23].

2.9 Efficiency of the Module

AC-DC Converters or rectifiers are basically posed with two control objectives:

(1) Tight regulation of output DC voltage and
(2) Regulation of input power factor to unity.

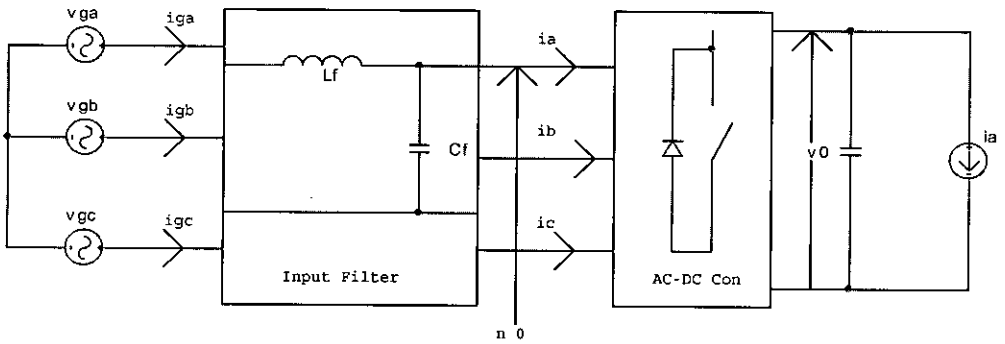


Figure 2.12: Schematic of 3-Φ rectifier with input filter

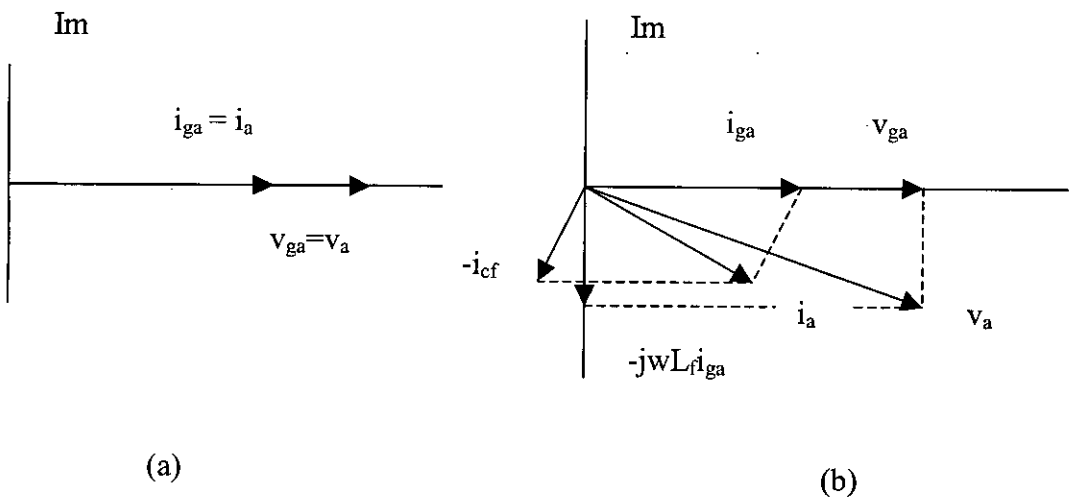


Figure 2.13: Phasor diagram of phase current and voltage (a) without input voltage (b) with input filter

The power drawn by the converter from the ac source is purely real and is equal to the power required by the load on the DC side. A generic schematic of a three phase rectifier supplied from an AC source through an input filter is shown in the Figure 2.12 .In the absence of an input filter the voltage $v_{ga}=v_a$ and the current $i_{ga}=i_a$. With the power factor regulated to unity, the current i_{ga} and the voltage v_{ga} are in phase. The phasor diagram showing the voltage and current at sinusoidal steady state is shown in Figure 2.13. With

the inclusion of the input filter, if the input current is required to be in phase with the voltage as shown in Figure 2.13(b), the converter must be forced to draw a current that is not in phase with the voltage at the input. Thus some **reactive power** has to be circulated between the converter and the input filter to achieve unity power factor at the source. As there is always a loss in power due to reactive component the efficiency can never be possible to achieve at 100%.

Chapter 3

Improvement of THD in Rectifier

3.1 Analysis of Three-Phase rectifier Circuit

Three phase diode bridge rectifier circuits are extensively used in many high power low cost applications leading the degradation in the power quality due to the current distortion. A simple three phase rectifier circuit and its input current and harmonics are shown in Figure 3.1, 3.2 and 3.3 respectively.

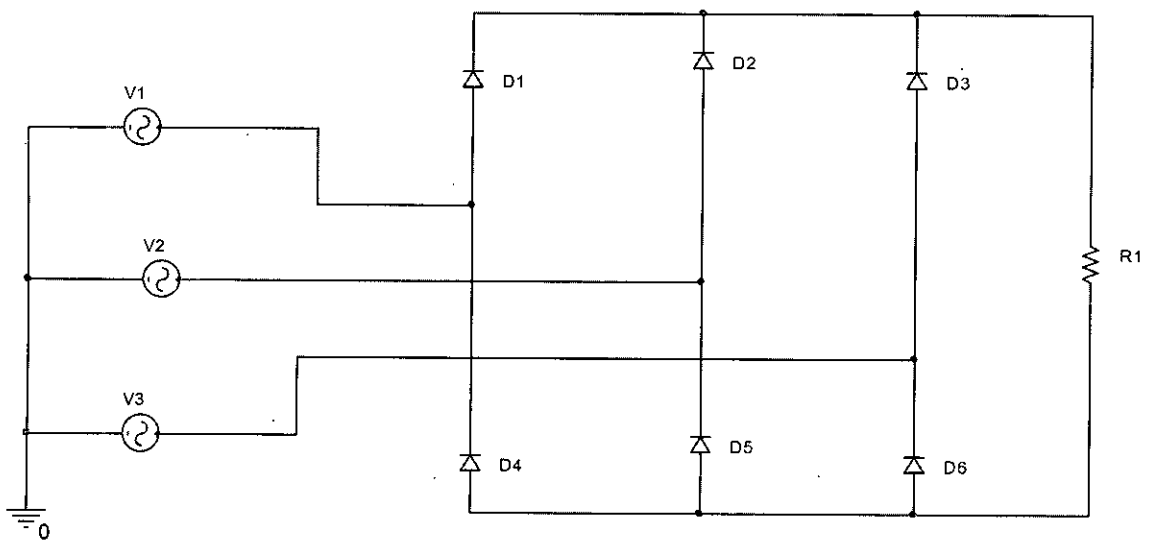


Figure 3.1: A simple diode bridge rectifier without capacitor

Here the input sinusoidal voltages V1, V2 and V3 have peak amplitude of 300Volts with phase difference of 120°. The frequency is 50Hz. The value of R1 is 100 ohm.

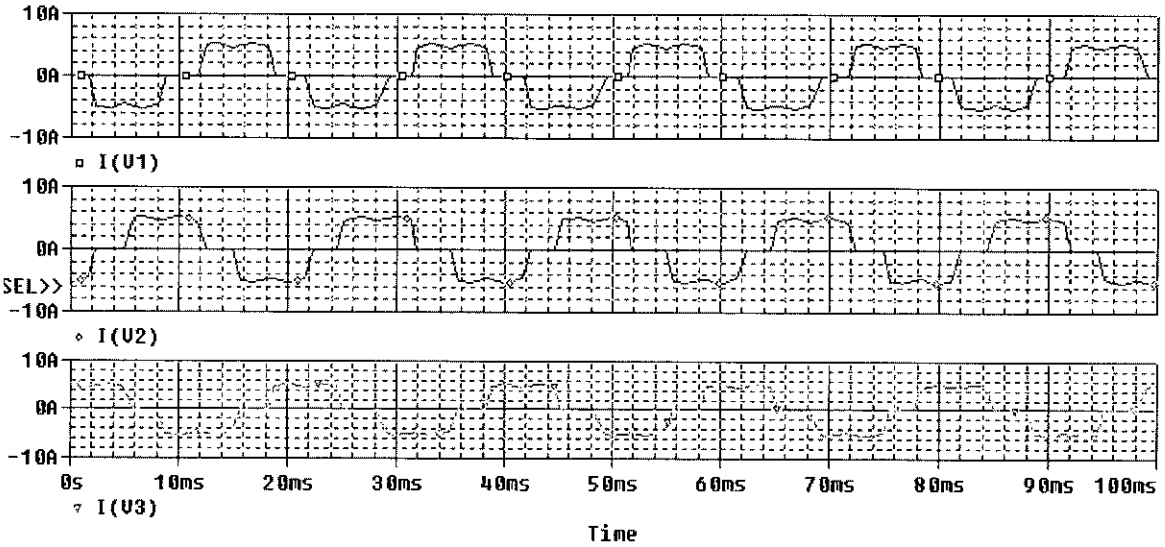


Figure 3.2: Input currents of the three phases of the rectifier for Phase A.B and C

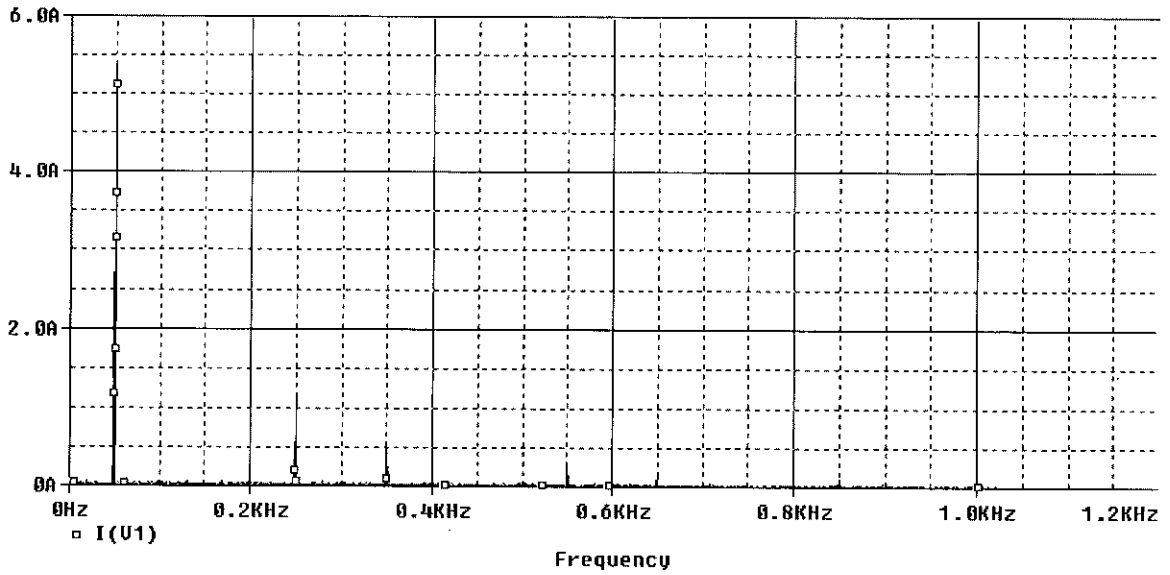


Figure 3.3: Harmonic content of phase A.

Table 3.1: Harmonic content of Current in Phase A without capacitor

Harmonics	Values(mA)	Harmonics	Values(mA)
I_1(50Hz)	5500	I_{11} (550Hz)	280
I_2(100Hz)	32	I_{12} (600Hz)	17
I_3 (150Hz)	5	I_{13}(650Hz)	160
I_4 (200Hz)	32	I_{14}(700Hz)	4
I_5(250Hz)	1100	I_{15}(750Hz)	4
I_6 (300Hz)	32	I_{16}(800Hz)	7
I_7 (350Hz)	500	I_{17}(850Hz)	71
I_8 (400Hz)	19	I_{18}(900Hz)	4
I_9 (450Hz)	3.5	I_{19}(950Hz)	40
I_{10}(500Hz)	20	I_{20}(1000Hz)	5

$$THD\% = \frac{\sqrt{\sum_{h=2}^{h=\infty} (M_h)^2}}{M_1} \times 100\%$$

Where M_h is the magnitude of either voltage or current harmonic component and M_1 is the magnitude of either the fundamental voltage or current.

Putting the values in the equation we have got the THD values for a simple rectifier is 22.83%.

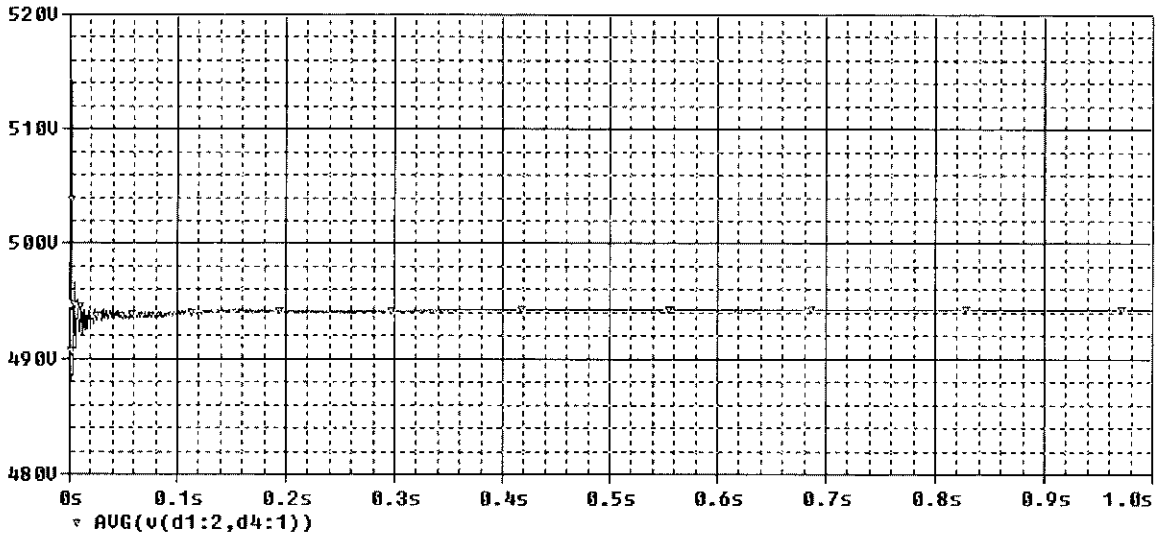


Figure 3.4: Output voltage of this rectifier without capacitor

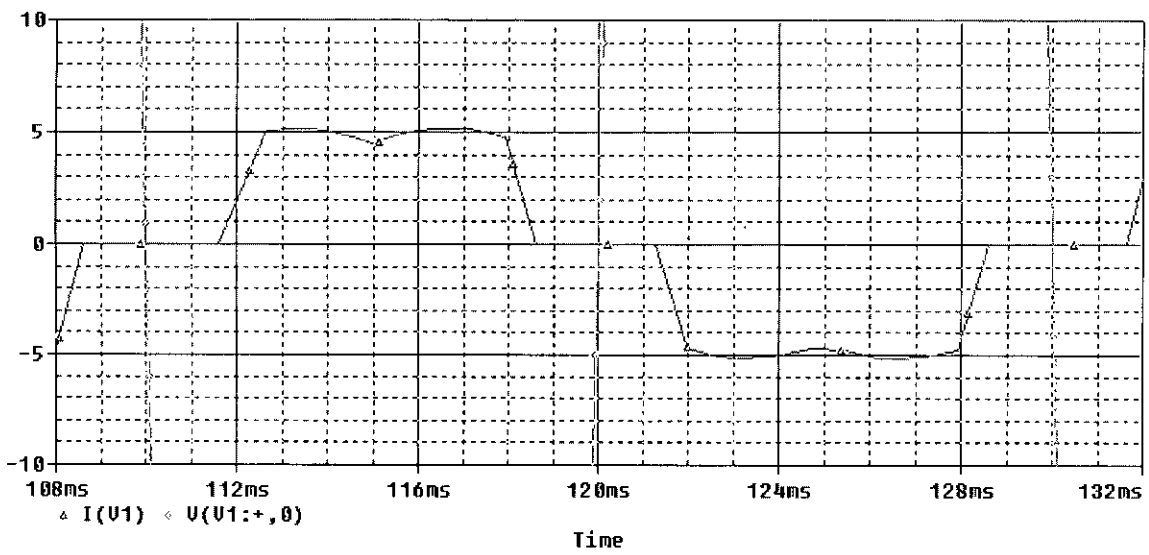


Figure 3.5: Phase relation of input current and voltage without capacitor

A three phase diode bridge rectifier with capacitor:

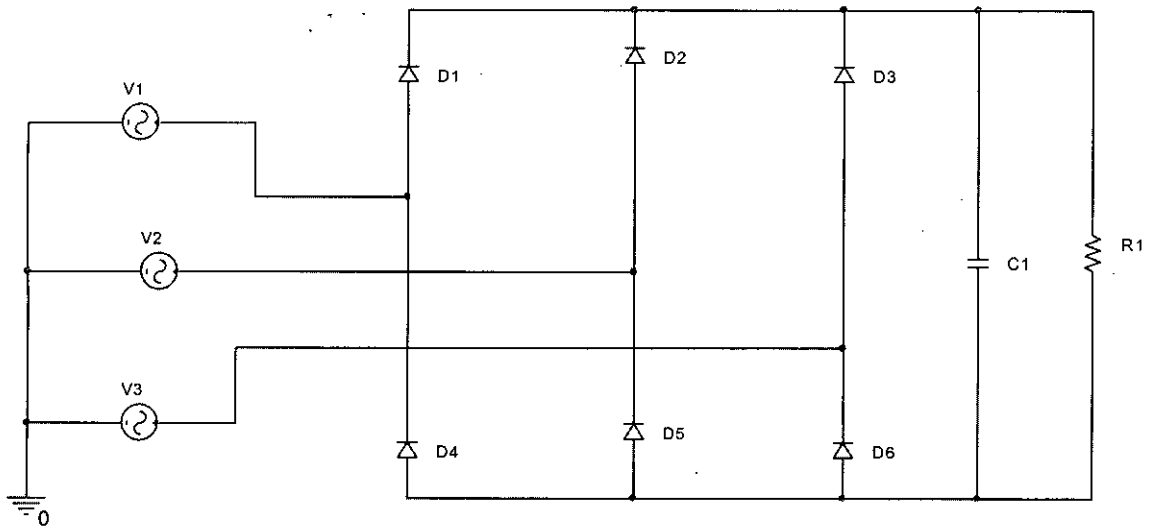


Figure 3.6: A simple diode bridge rectifier with capacitor (C1=100uF)

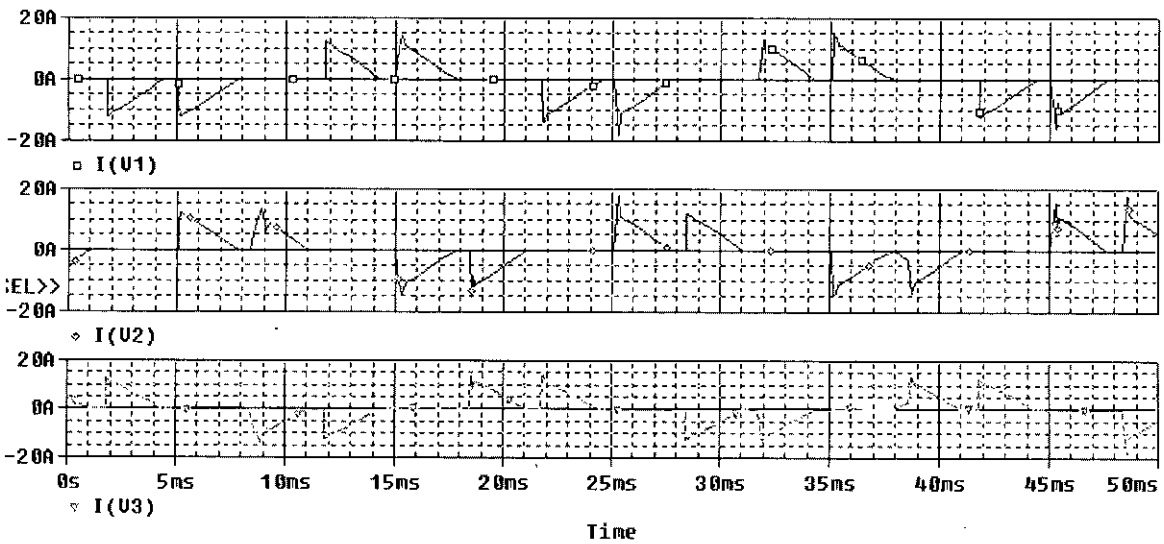


Figure 3.7: Input currents for the three phases of the rectifier with capacitor.

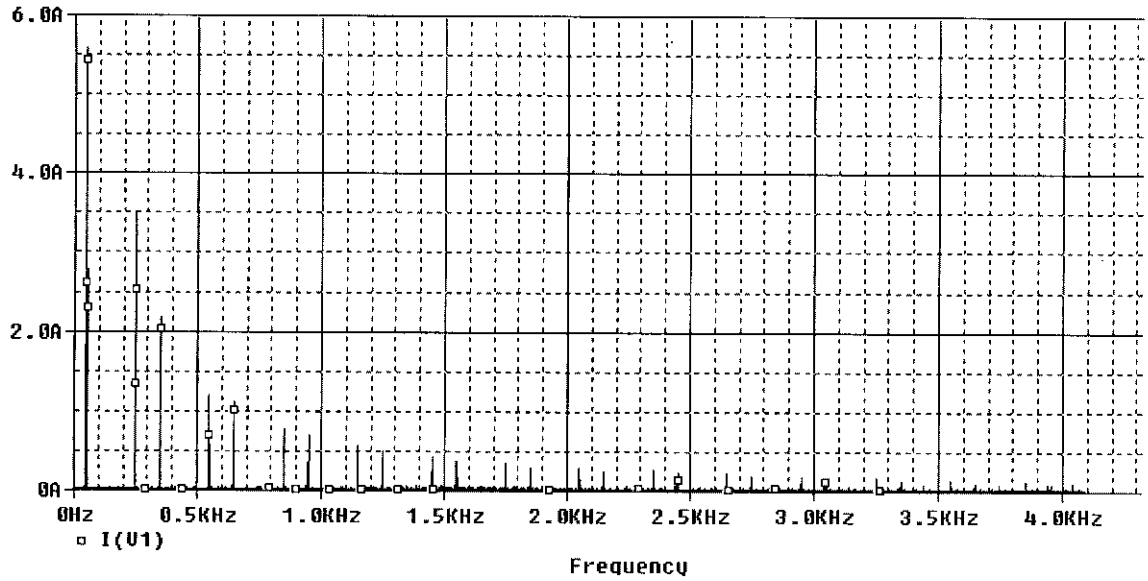


Figure 3.8: Harmonic content for phase A for rectifier with capacitor

Table3.2: Harmonic content of Current in Phase A with capacitor

Harmonics	Values(mA)	Harmonics	Values(mA)
I_1(50Hz)	5500	I_{11}(550Hz)	1200
I_2(100Hz)	25	I_{12}(600Hz)	20
I_3(150Hz)	19	I_{13}(650Hz)	1150
I_4(200Hz)	10	I_{14}(700Hz)	11
I_5(250Hz)	3500	I_{15}(750Hz)	11
I_6(300Hz)	20	I_{16}(800Hz)	10
I_7(350Hz)	2200	I_{17}(850Hz)	780
I_8(400Hz)	13	I_{18}(900Hz)	19
I_9(450Hz)	20	I_{19}(950Hz)	700
I_{10}(500Hz)	10	I_{20}(1000Hz)	10

The THD value obtained from this rectifier is 83.23%. This is an extremely large value. Because of the insertion of the capacitor at the output to make the output voltage ripple free the input current becomes too much distorted and harmonic content has increases a lot.

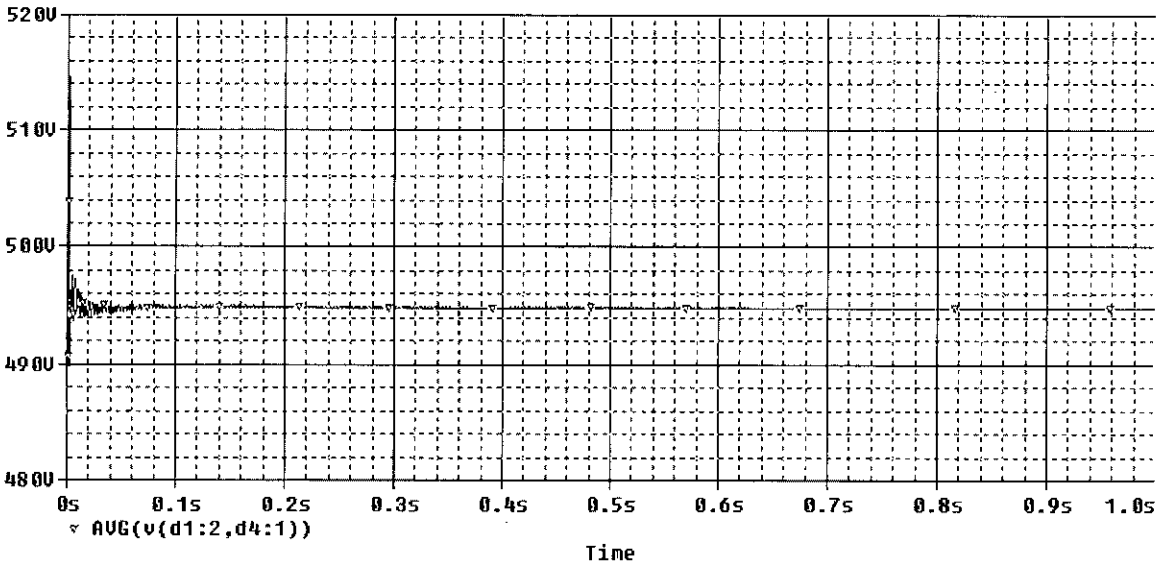


Figure 3.9: Output voltage of three phase rectifier with capacitor

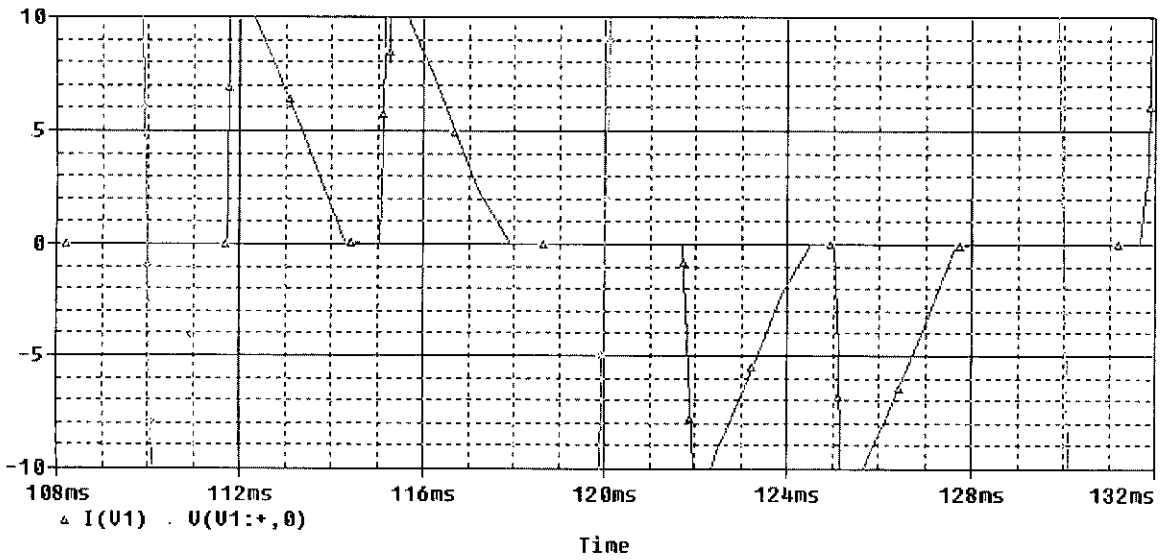


Figure 3.10: Phase relation of input current and voltage with capacitor

3.2 Harmonics reduction with passive filters

By observing the input current wave shapes of these filters we can say about the harmonic contents of them. There is no even harmonics as the waveforms are symmetrical about the X-axis. Another noteworthy fact is balanced three phase rectifier type loads do not produce a third harmonic component. Nor do they produce any triplen harmonic component. Again the 11^{th} harmonic and higher is a point where the magnitude diminishes to a very low level. Thus 5^{th} and 7^{th} orders are the problem child harmonics for AC drives.

Table 3.3 Harmonic spectrums Analysis

Harmonics	Peer unit value	Frequency
Fundamental	1	50Hz
5^{Th}	0.2	300Hz
7^{th}	0.14	350Hz
11^{th}	0.09	550Hz
13^{th}	0.07	650Hz
17^{th}	0.06	850Hz
19^{th}	0.05	950Hz

Passive filters may be used for reducing the harmonics content of the input currents. But they do not allow the regulation of the output voltage and also decreases the output voltage levels in comparison with the unfiltered rectifiers.

Taking the harmonic limits as a quality index, the resulting inductors are typically larger than the ones used in high quality rectifiers using active circuits [26].

3.3 Effect of input inductance

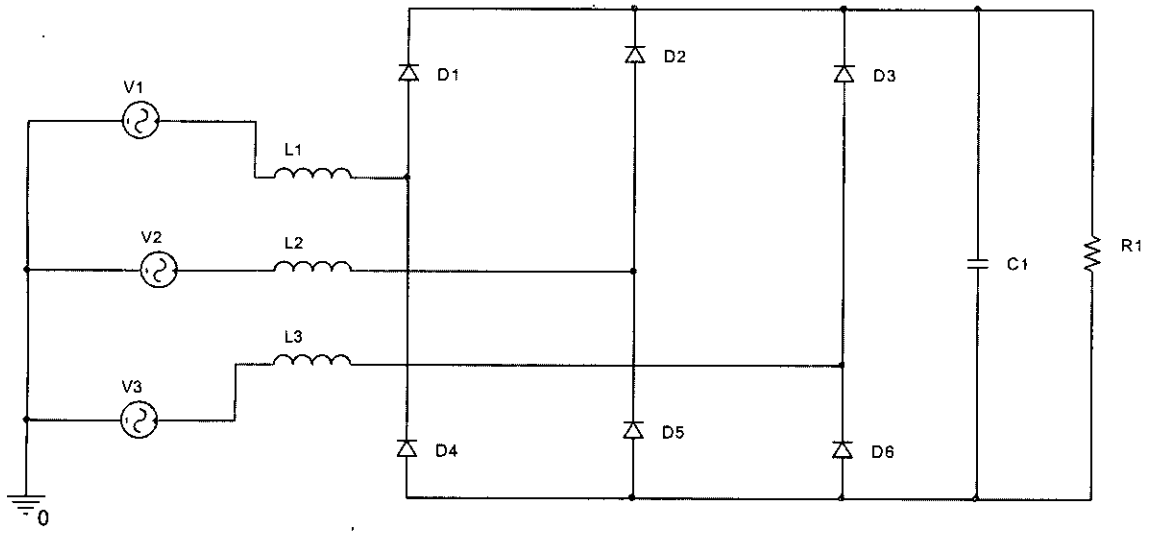


Figure 3.11: Three- phase rectifier with input inductance

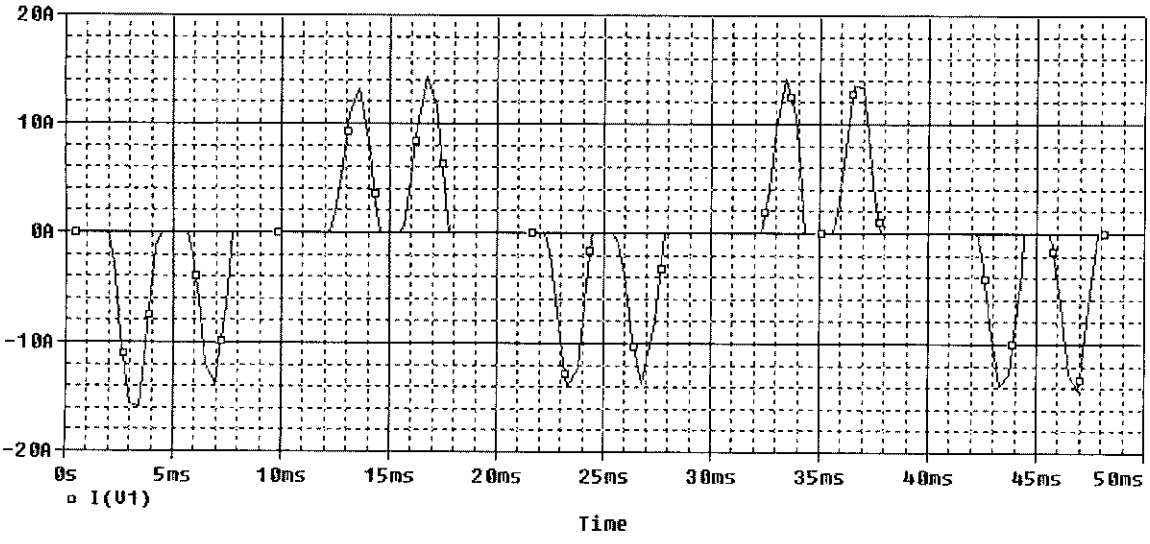


Figure 3.12: Input current with 1mH inductor

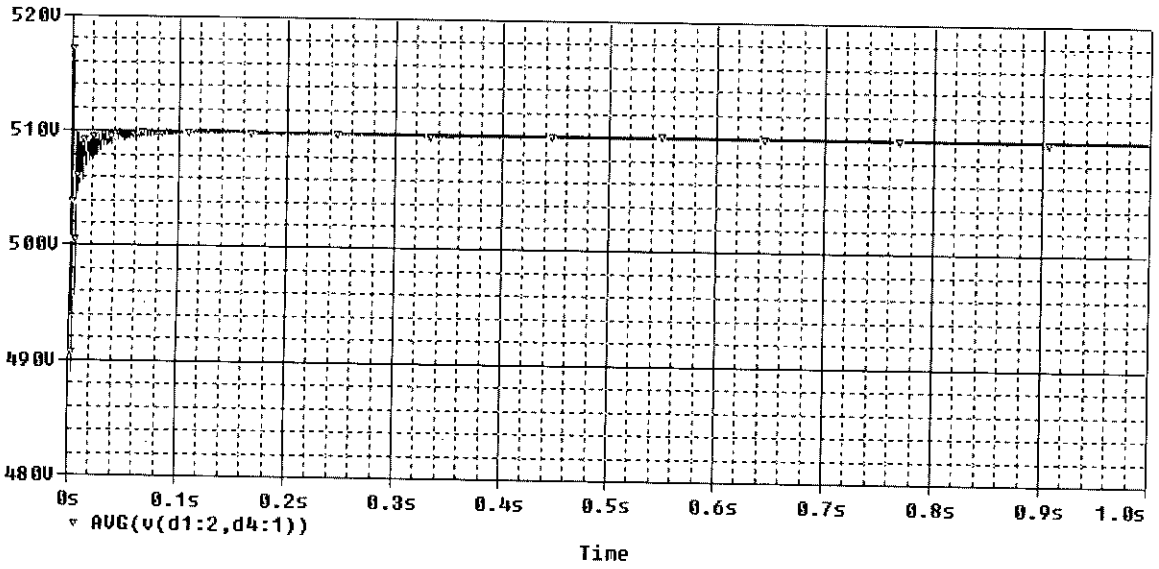


Figure 3.13: output voltage with 1mH inductor

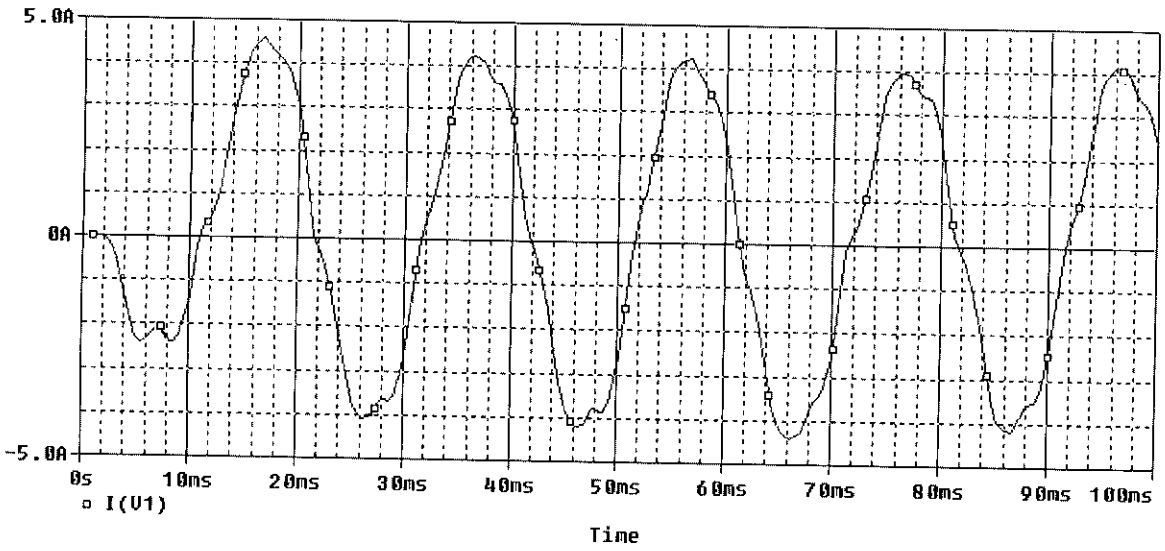


Figure 3.14: Input current with L=100mH

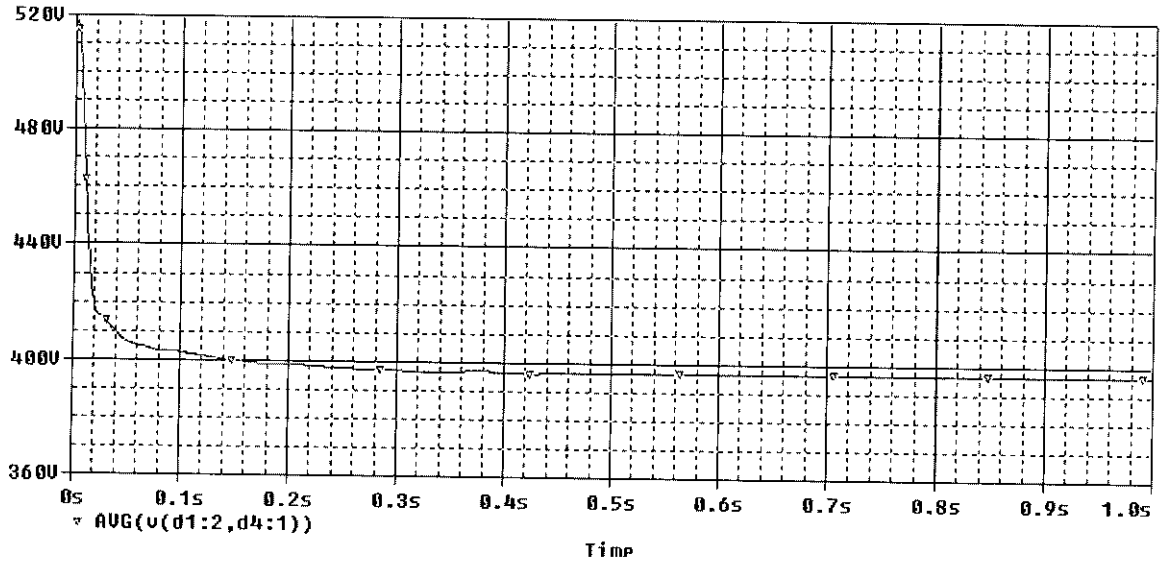


Figure 3.15: output voltage with L=100mH

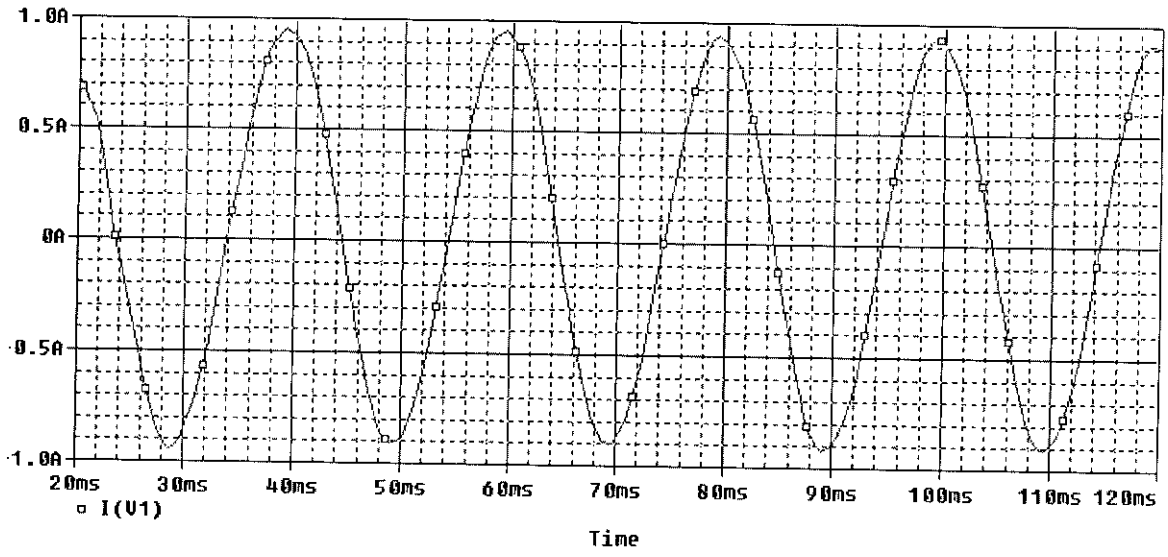


Figure 3.16: input current with L=1H

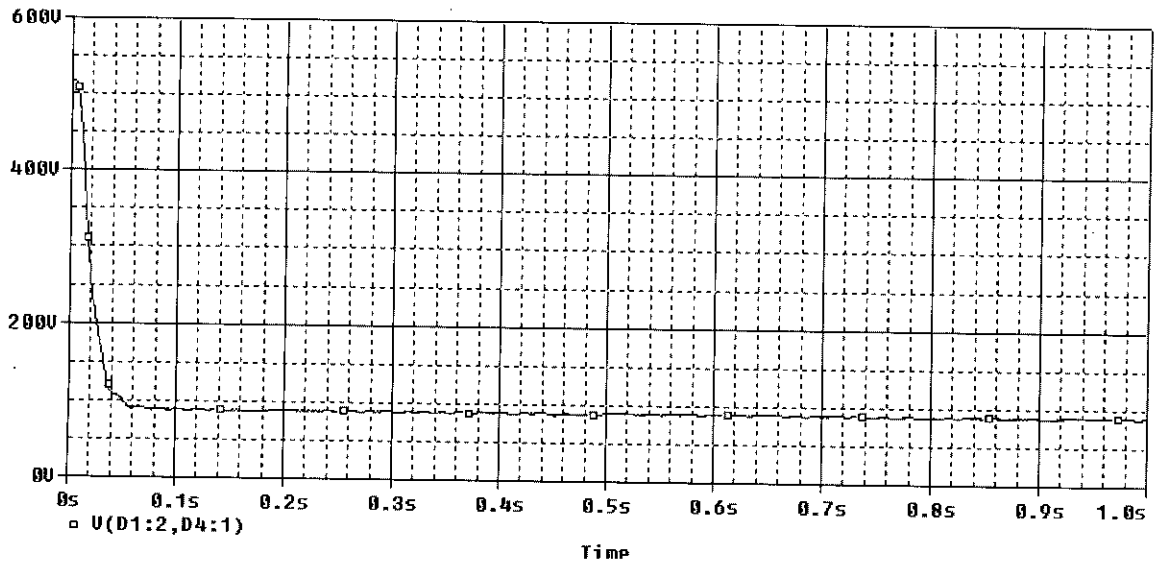


Figure 3.17: output voltage with L=1H

As we can see with increasing the inductive value of the input inductor the current wave shape improves but the output voltage level decreases.

3.4 Rectifier Analysis with passive filter

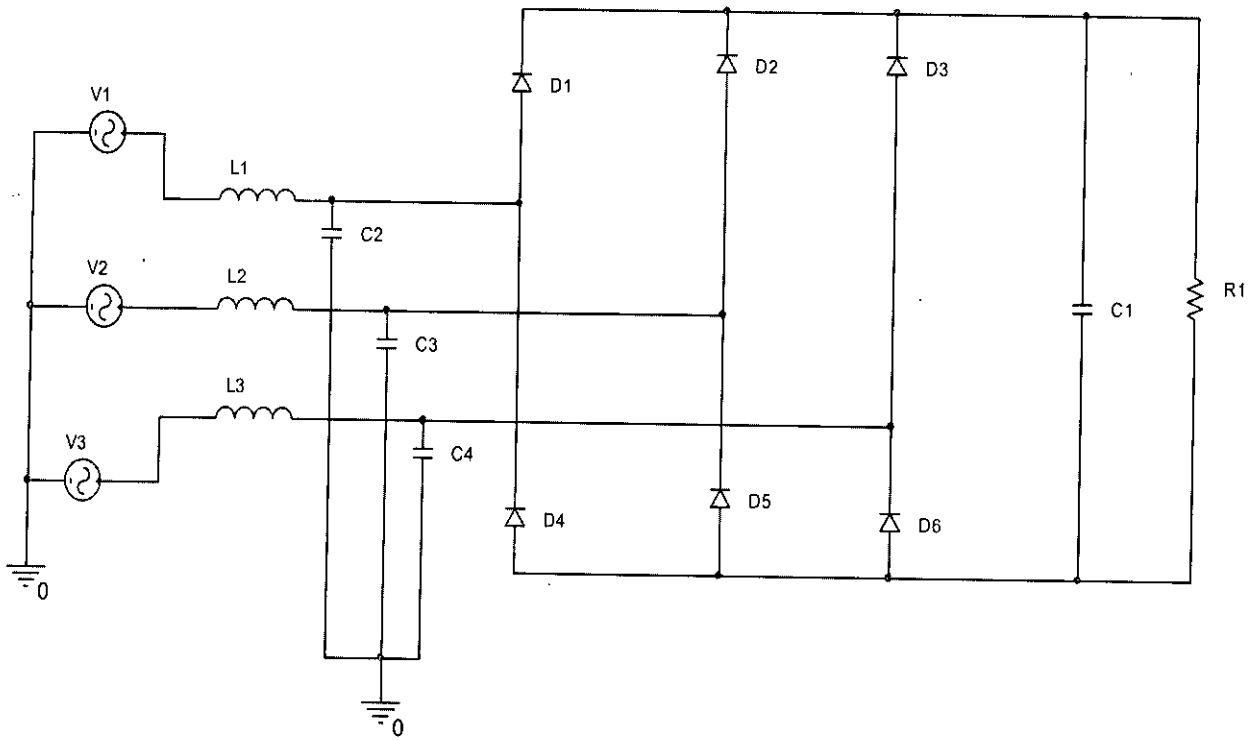


Figure 3.18: Rectifier with Input passive filter

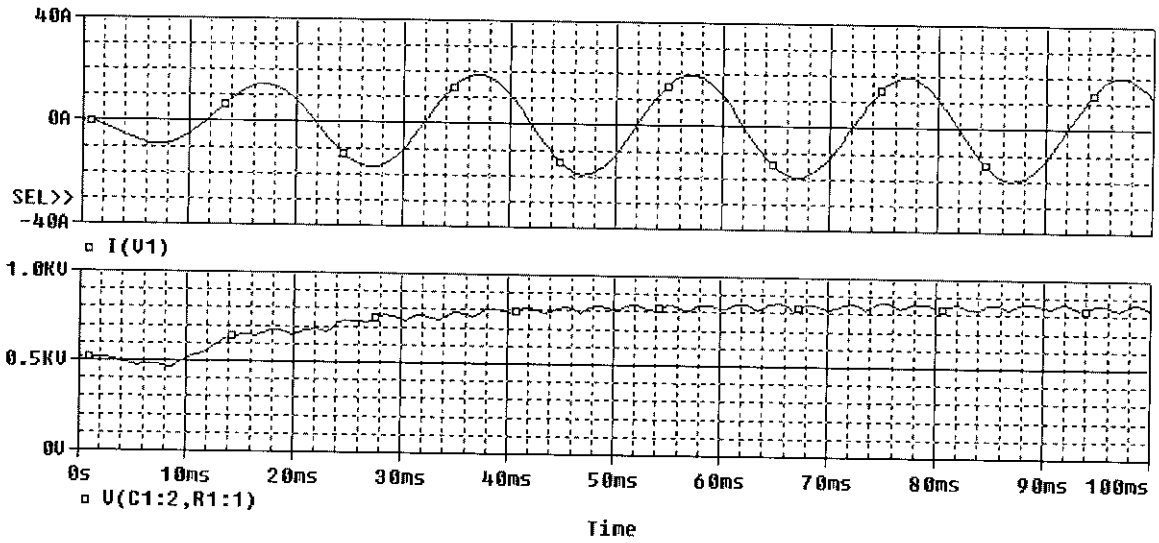


Figure 3.19: Input current and output voltage of rectifier with passive filter, L=100mH, C= 100uF

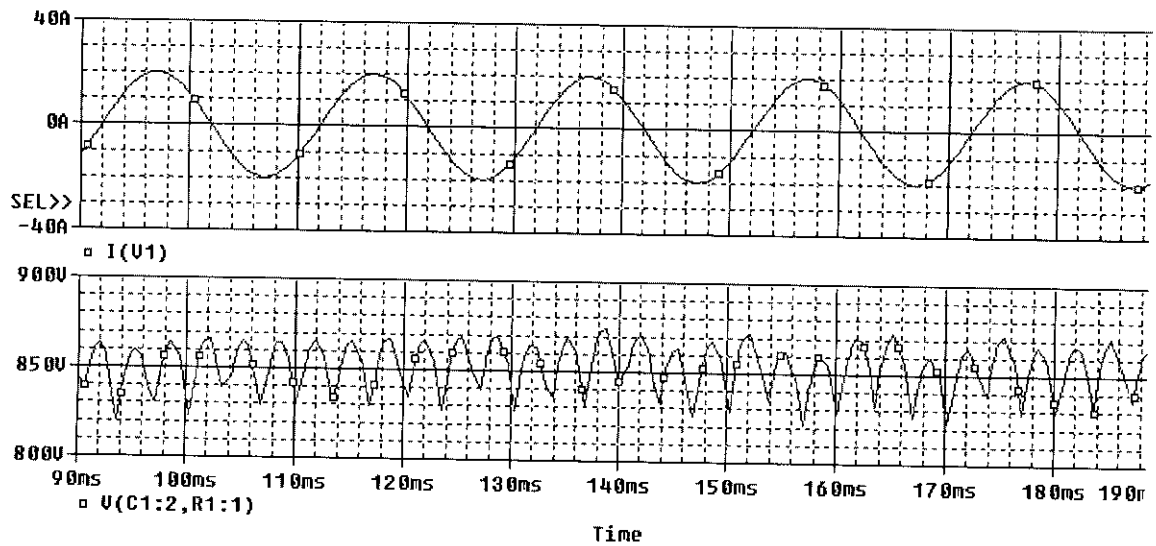


Figure 3.20: Input current and output voltage of rectifier with passive filter, L=100mH, C= 100uF

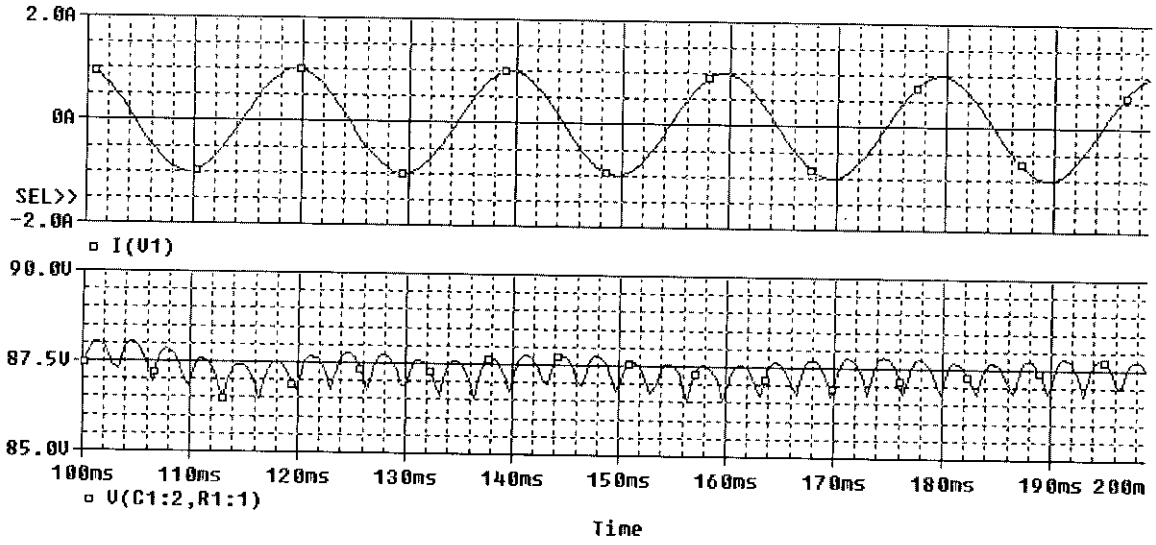


Figure 3.21: Input current and output voltage of rectifier with passive filter, L=1H, C= 10uF

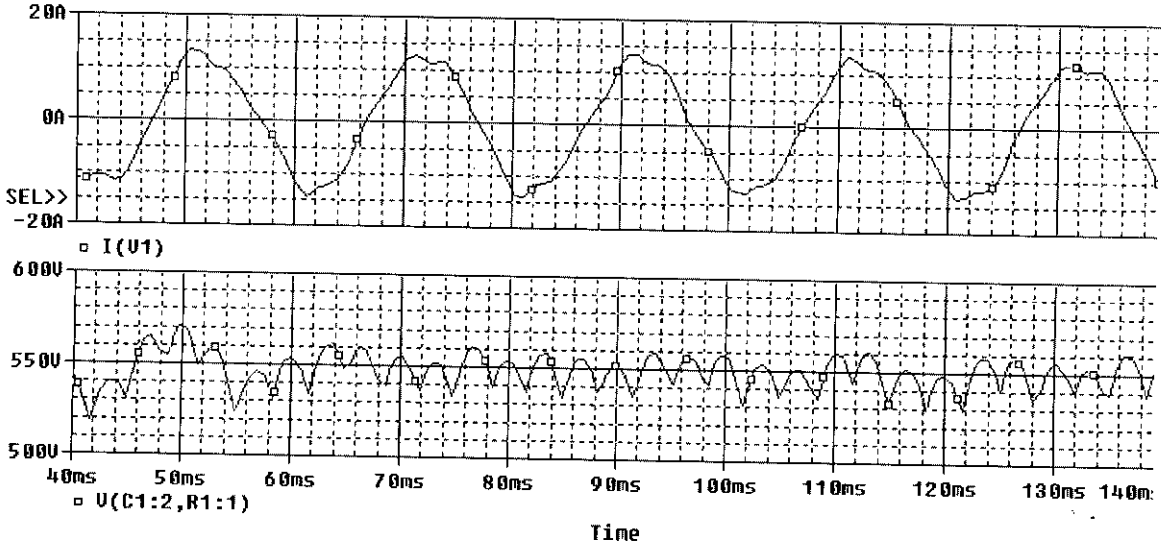


Figure 3.22: Input current and output voltage of rectifier with passive filter, L=10mH, C= 100uF

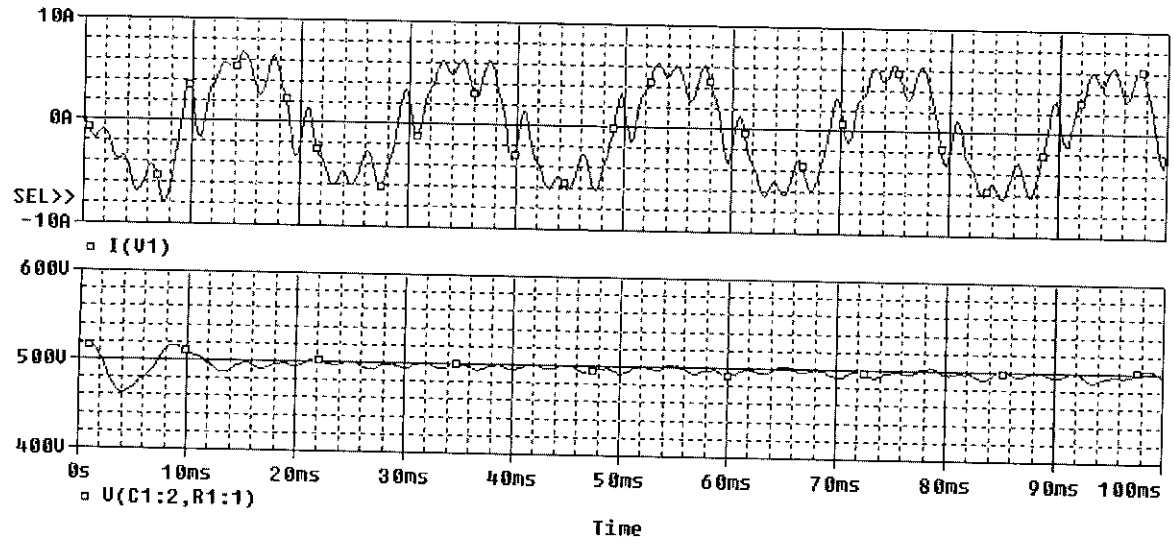


Figure 3.23: Input current and output voltage of rectifier with passive filter, L=10m, C=10uF

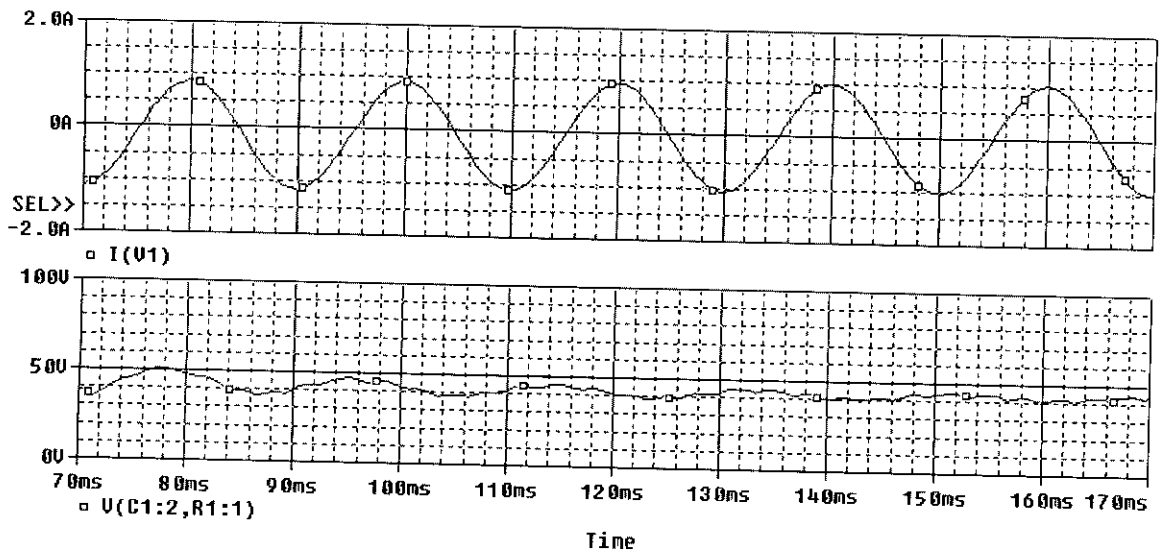


Figure 3.24: Input current and output voltage of rectifier with passive filter, L=1H, C=100uF

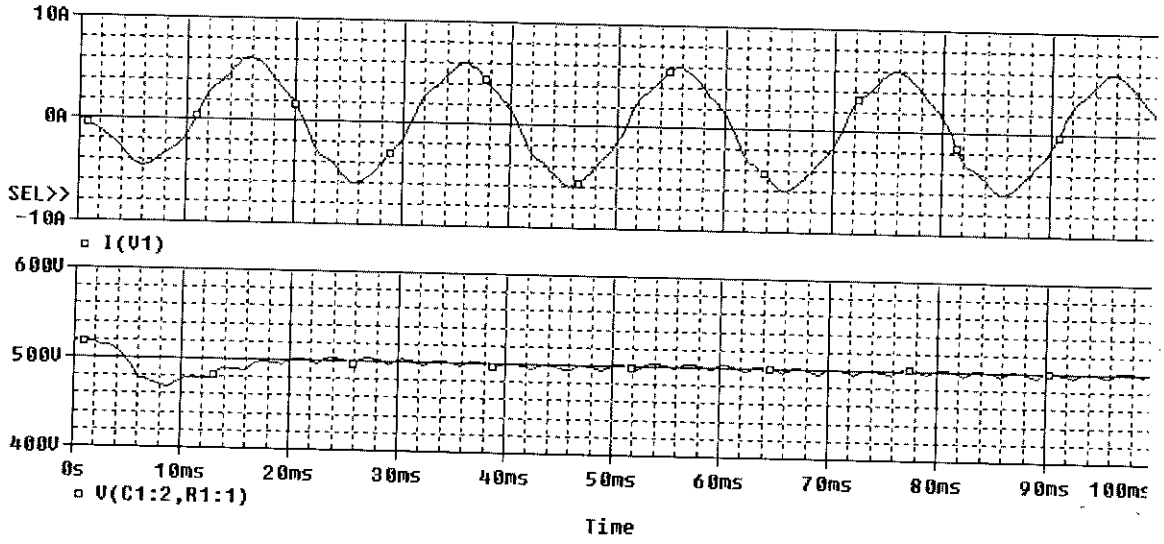


Figure 3.25: Input current and output voltage of rectifier with passive filter, L=100mH, C= 10uF

From the wave shapes shown above, we see that as the value of the inductor increases the input wave shapes is improving a lot but output voltage level decreases and decreasing the value of inductor gives more output voltage but input current distortion increases.

The best model for the passive filter has obtained when L=100mH and C=100uF. Actually the filter can be modeled by calculating the resonating values as $X_L=X_C$.

Calculation is done by considering the fundamental component as 50Hz. The product of LC should be 1×10^{-5}. So in a passive input filter the component values are very large and regulation of the output voltage is not possible.

3.5 Introduction of switching in boost Rectifier

Active wave shaping means to introduce switching in the rectifier circuit. In a boost rectifier if switching is introduced without having any input filter the input current start to conduct in Discontinuous Conduction Mode (DCM) and the wave shape follows the input voltage but have high frequency switching components. The output cannot reach to a desired level as expected by the boost converter. So, though the input current wave shape has improved a lot, the efficiency of this module is not good at all. In Figure 3.26 the input current and output voltage wave shapes are given that we got from simulation.

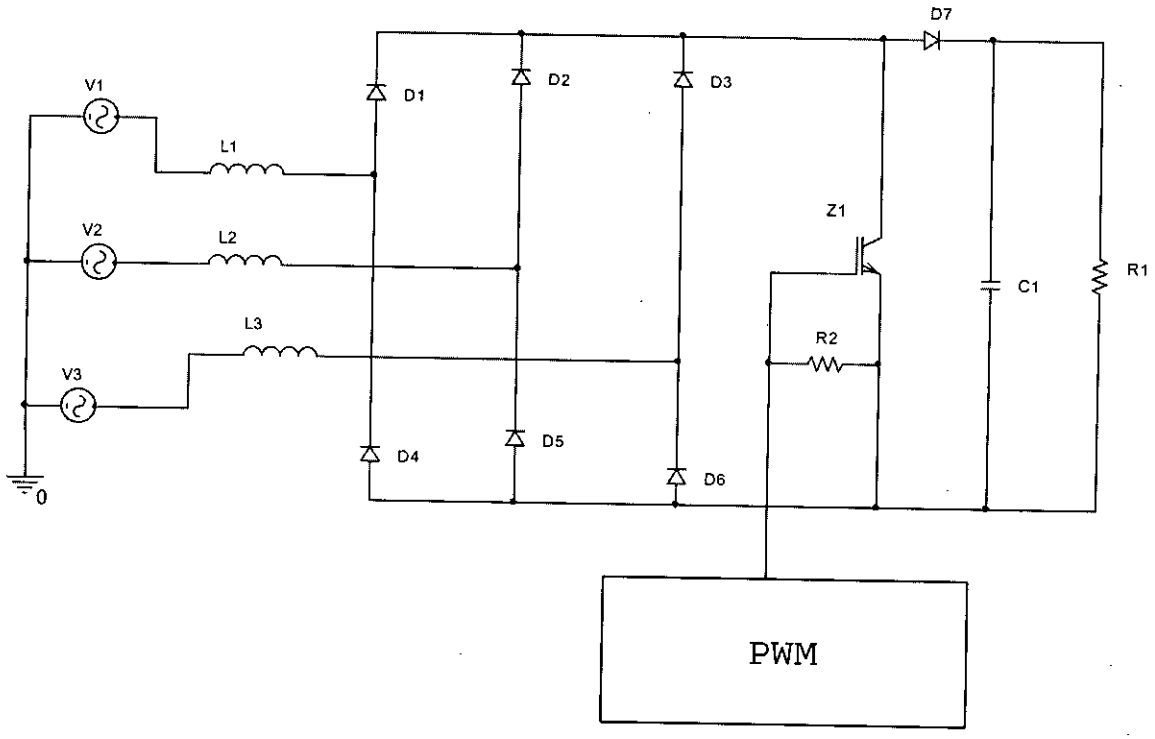


Figure 3.26: A single switch boost rectifier

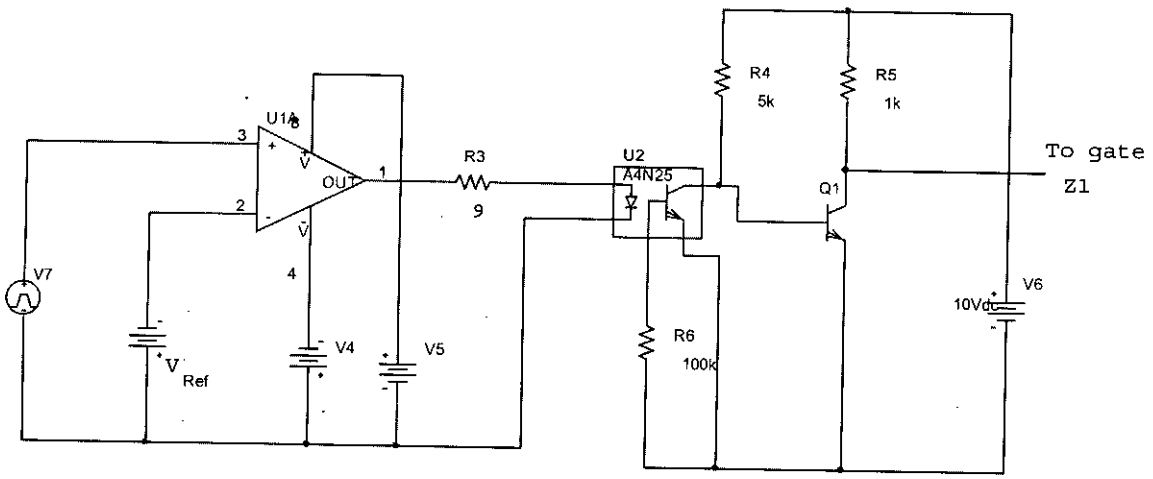


Figure 3.27: PWM circuit.

- Biasing voltages V4 and V5 are 0f 15 Volts here. V_{Ref}=12Volts.

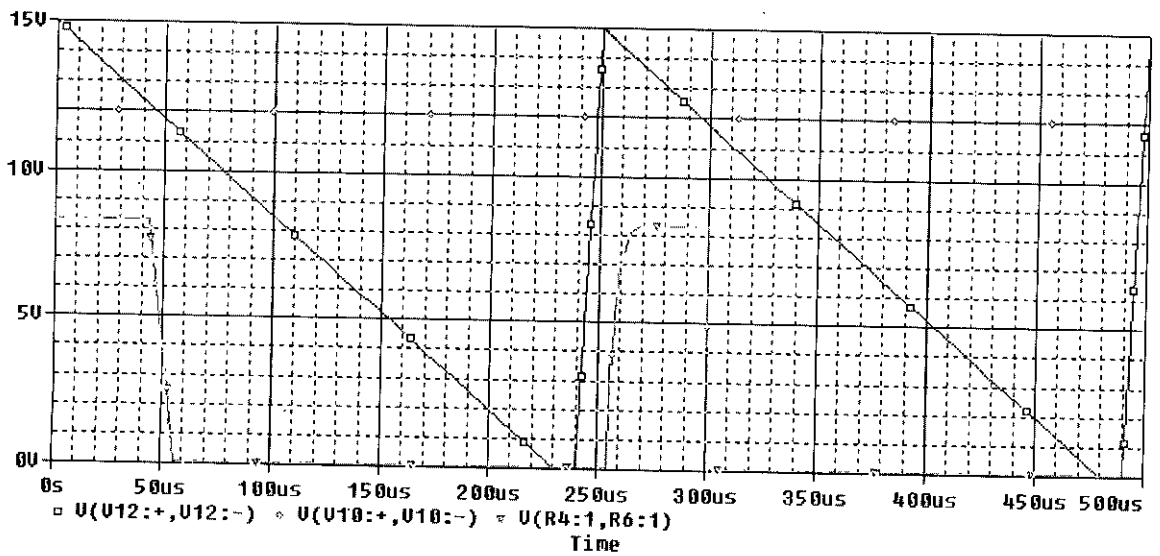


Figure 3.28: The input voltage (V7) of the opamp at pin #3

A 4 KHz saw tooth wave varying from 15 to zero volts.

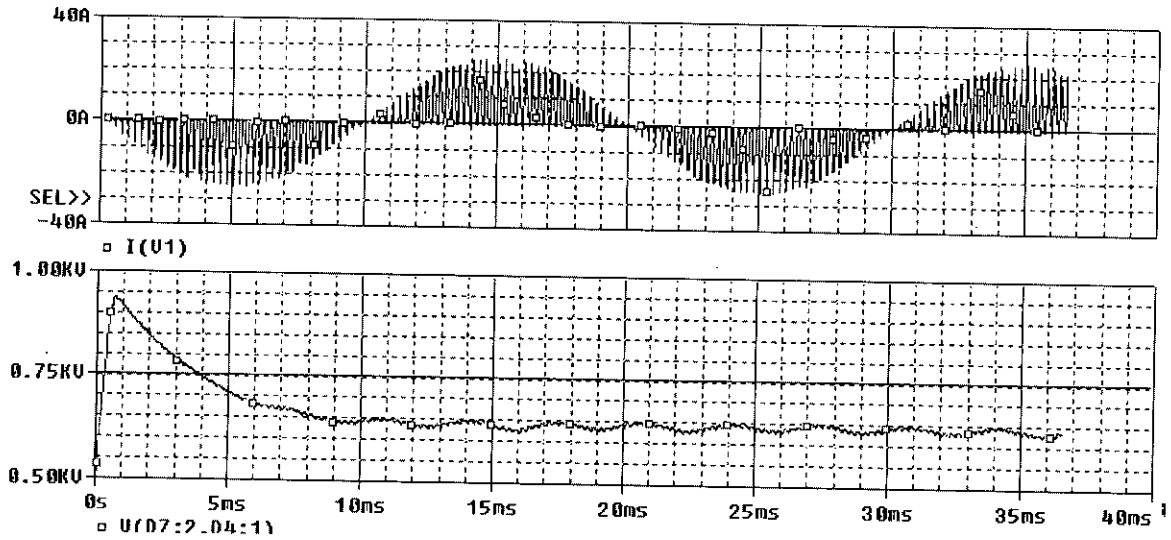


Figure 3.29: Input current and output voltage with switching

With the introduction of EMI filter at the input can make the performance better. Figure 3.26 shows the circuit model for switching with input filters where L1, L2 and L3 are of 500uH. Here we can see the output voltage has got to an acceptable high value as expected by boost rectifier. Input current wave shape is also shown in Figure 3.29.

Boost Rectifier with EMI filter and switching

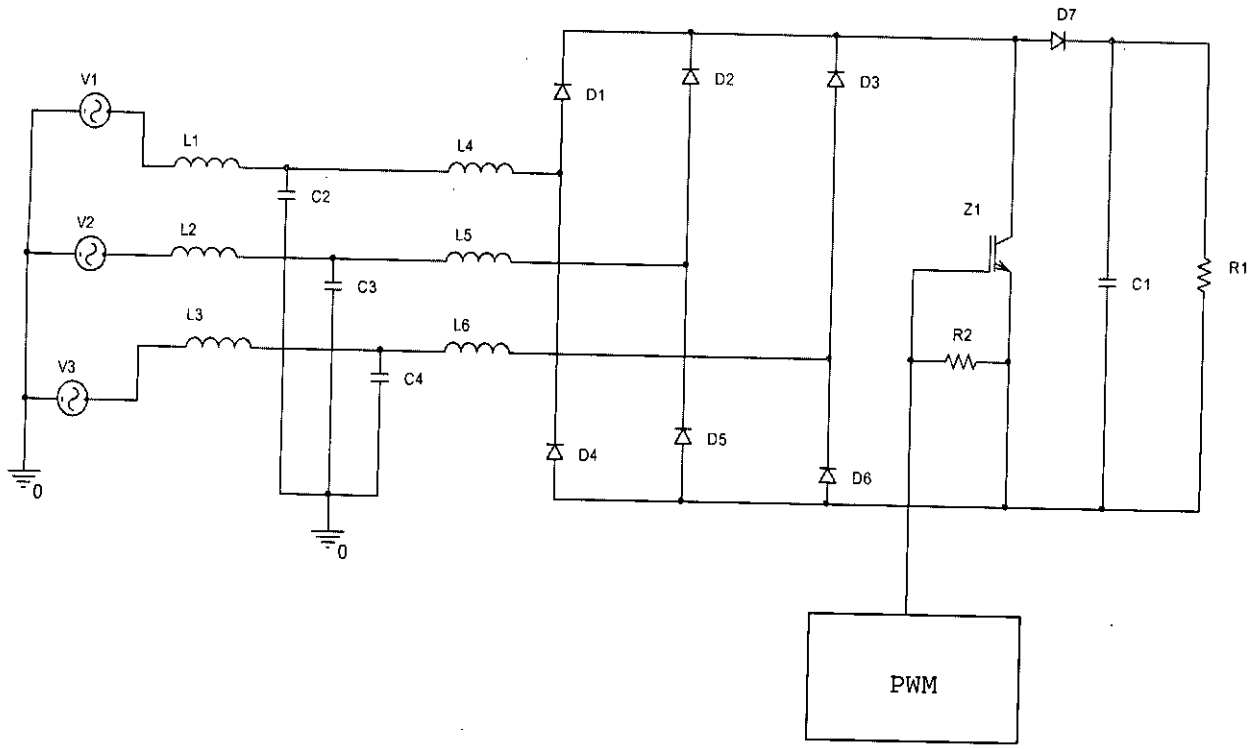


Figure 3.30: A three phase boost rectifier with single switch and input filter

L1.L2 and L3 10uH,

L4, L5 and L6 .1mH

C2, C3 and C4 1nF

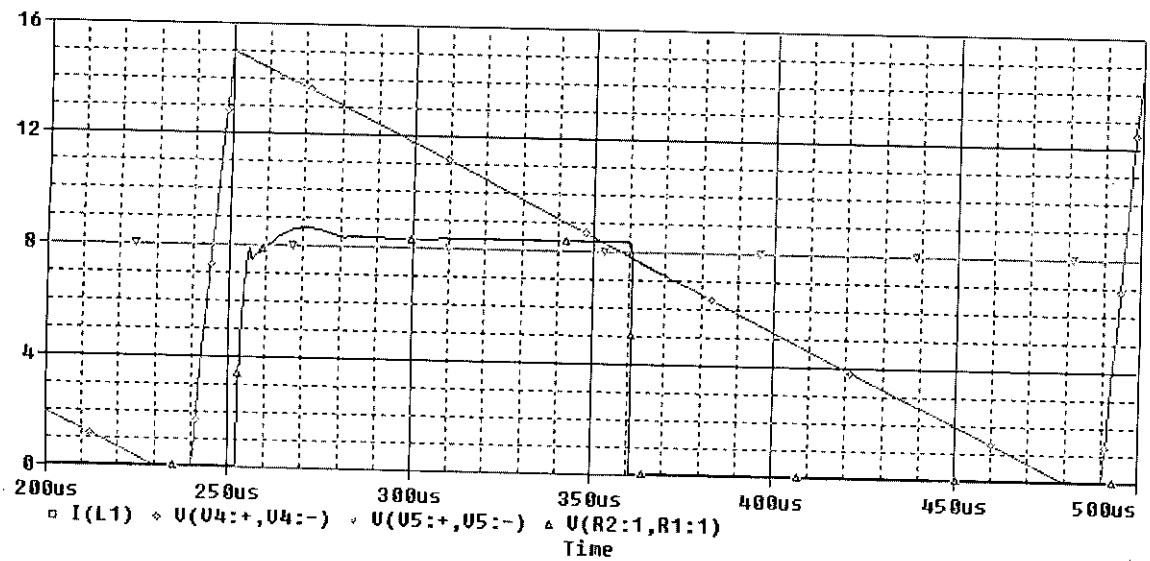


Figure 3.31: Duty Cycle of IGBT

Duty cycle: D=(361-252)/250=0.436

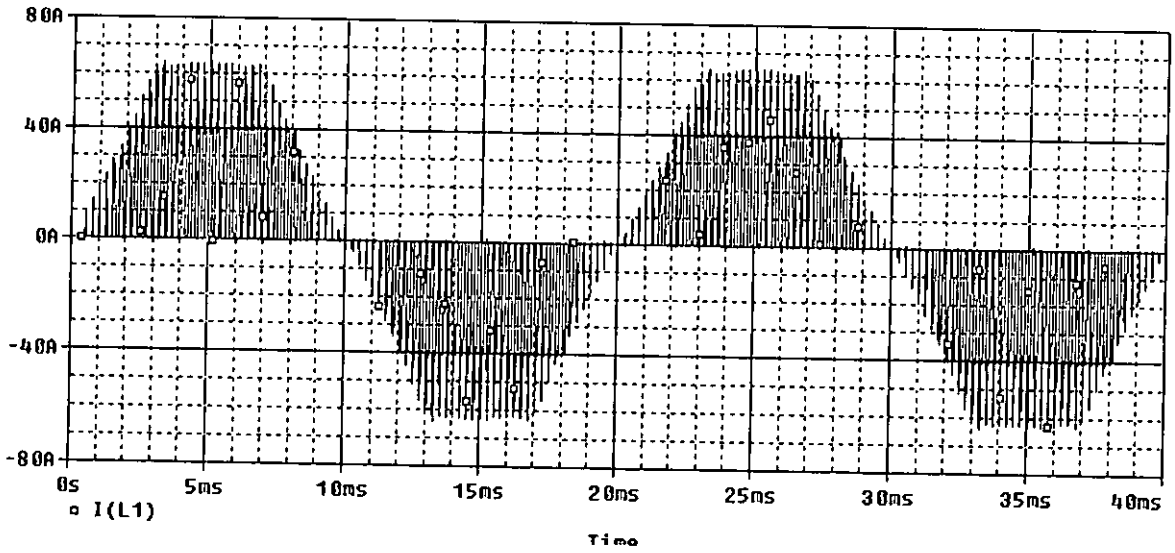


Figure 3.32: Input current for boost rectifier with input filter and switching

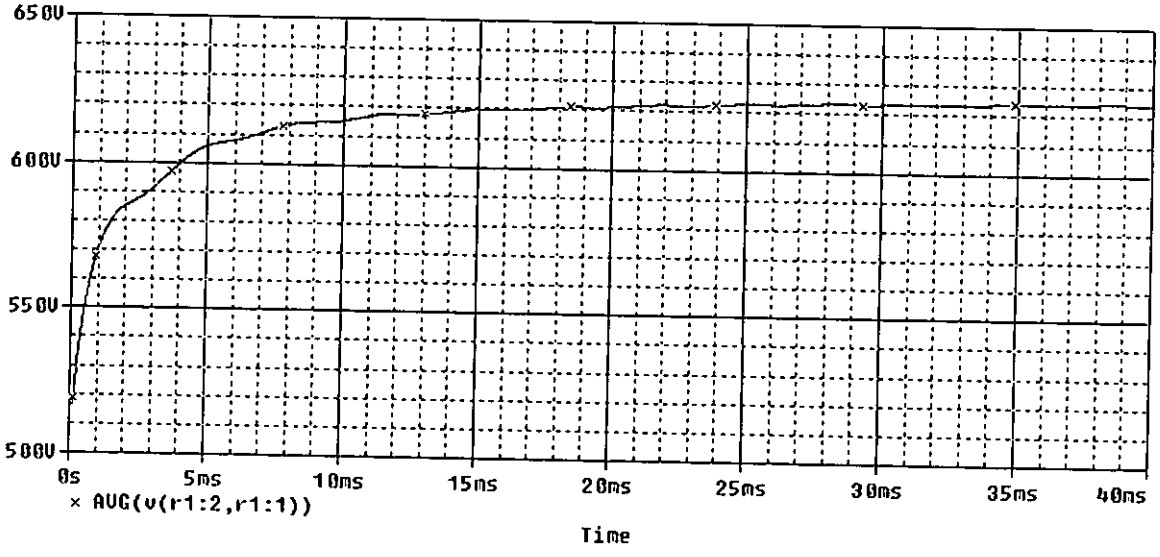


Figure 3.33: Out Voltage for boost rectifier with input filter and switching

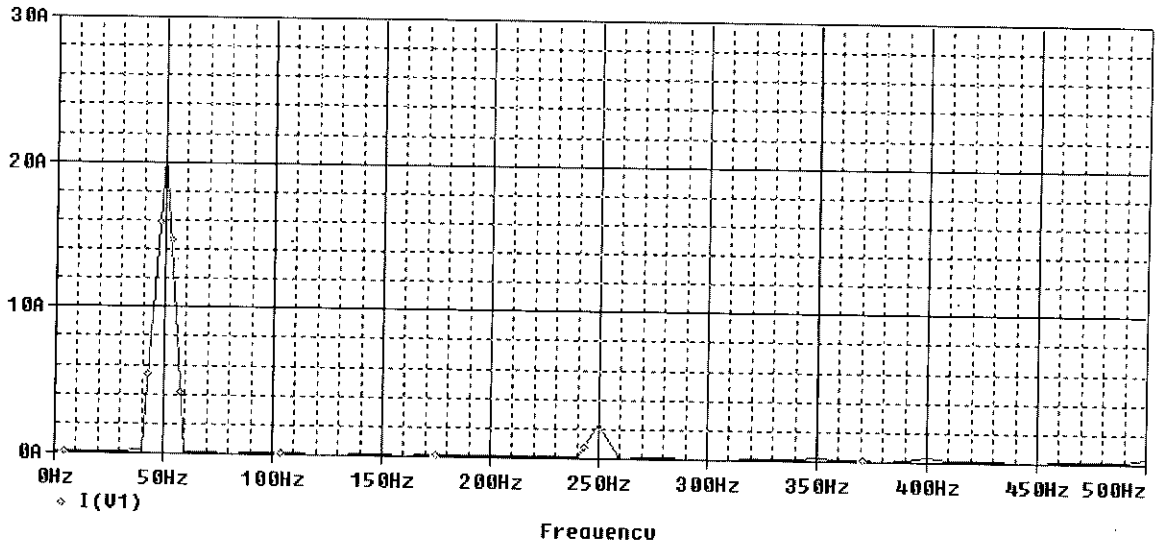


Figure 3.34: Input current Harmonics for boost rectifier with input filter and switching

Table3.4: Harmonic content of Current in Phase A with switching and EMI filter

Harmonics	Values(mA)	Harmonics	Values(mA)
I_1(50Hz)	22000	I_{11}(550Hz)	700
I_2(100Hz)	850	I_{12}(600Hz)	165
I_3(150Hz)	450	I_{13}(650Hz)	545
I_4(200Hz)	375	I_{14}(700Hz)	103
I_5(250Hz)	3400	I_{15}(750Hz)	136
I_6(300Hz)	380	I_{16}(800Hz)	132
I_7(350Hz)	1000	I_{17}(850Hz)	455
I_8(400Hz)	255	I_{18}(900Hz)	60
I_9(450Hz)	205	I_{19}(950Hz)	157
I_{10}(500Hz)	196	I_{20}(1000Hz)	50

At this point the THD value reduced to 17.61%

A THD value of a three phase rectifier with capacitor can be reduced from 83.23% to 17.61% with the introduction of active filter or switching in the rectifier. Even at this value of the THD is not acceptable as recommended by different regulations and standards. Some application requires the THD to be less than 10% and even some other application have set the limit to be less than 5%. That is why research is still going on improving the THD level. In our work, we tried to reduce the value to be less than 5%.

Chapter 4

Analysis of the Proposed Model

4.1 Introduction

In our model we have introduced both active and passive filtering in the rectifier circuit. We have chosen constant frequency for switching, as it is easy to analyze. The harmonics generated by constant frequency switching are known and so to reduce their effect is easy to implement. To regulate the output voltage at a certain level Pulse Width Modulation (PWM) has been introduced. With this introduction of switching very high frequency harmonics are generated at the input of the rectifier. To reduce these high frequency harmonics the circuit is analyzed in a simplified model and from the transfer function of the model the filter components (L and C) are calculated. At this point the input current still has the low frequency components and THD value cannot reduce to less than 17%. Here we had introduced another series LC filter so that the filter resonates to the supplied input voltage's frequency. As a result, the input current wave can appear to be nearly sinusoidal and the THD value calculated has been reduced to less than 3%. This is a great improvement over the previous models that do not have the input series LC filter.

4.2 Boost Inductor Design

Due to DCM operation, the boost inductor design in a single switch three phase boost rectifier is critical. If graphical method is used for the designing then the obtained values are not always give accurate results. Moreover, this is not a convenient method. In this section a simple and a straightforward approach is introduced to simplify the design procedure.

For DC/DC boost converter, the results shown in Appendix 6 can be directly used to calculate the boost inductor. In fact, if the boost inductor of the DC/DC converter is selected so that the critical power (P_c) is greater than the maximum delivered power, the DC/DC converter operates in DCM.

For a single switch three-phase boost rectifier the solution is not so straightforward. This is because the duty cycle D, critical power P_c and delivered power are functions of the instant of the sinusoidal input voltages. That is, over one line cycle the critical power P_c is time variant. At some instances, the critical power is going to be minimum. Hence, if these instants can be found and the inductor is designed at these worse cases, then the converter will operate at DCM over the whole line cycle.

Equations for balanced and undistorted three phase input voltages are

$$V_a = V_m \sin(wt) \qquad 4.1$$

$$V_b = V_m \sin(wt - \frac{2\pi}{3}) \qquad 4.2$$

$$V_c = V_m \sin(wt - \frac{4\pi}{3}) \qquad 4.3$$

The equivalent input voltages of the circuit for Figure 4.1 has to be:

$$V_{in\_e} = \max\left[|V_{ab}(wt)|, |V_{bc}(wt)|, |V_{ca}(wt)|\right], \qquad 4.4$$

Which repeats every 60° and reaches maximum at wt=n×60°, n=0, 1, 2 ----, where the equivalent inductor is L_e=2L. This is because at these moments one phase voltage is zero and other two-phase voltages applied to the input see two inductors in series.

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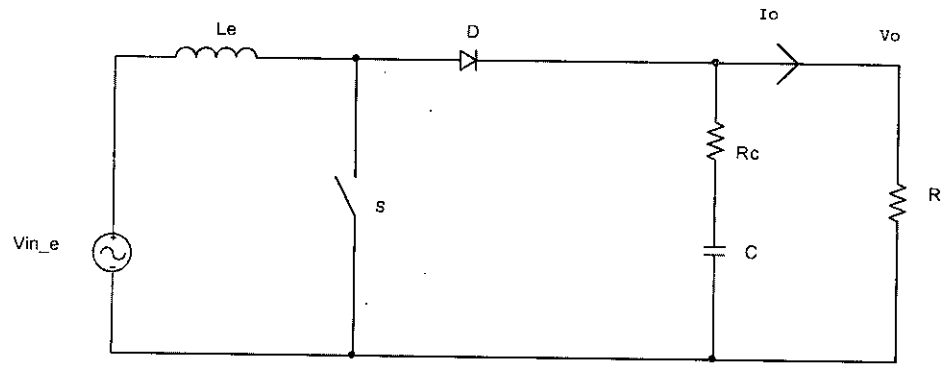


Figure 4.1: Equivalent circuit of the single switch three-phase boost rectifier

Once the input voltage reaches the maximum the duty cycle D is minimum and the critical power reaches the minimum. In other words, at these moments, the converter is going to operate to CCM the earliest as load is increased. For example, during $0° \leq wt \leq 60°$, wt=0°and wt=60° are the two moments at which the critical power reaches the minimum, and the converter is going to operate in CCM earlier than the other moments. So, as long as the inductors are selected to ensure the DCM operation at these two particular moments, as load is maximum the converter is going to operate in DCM in the whole load range and whole time periods. This is the inductor design guideline for the single switch three phase boost rectifier.

Typically, with the parameter in the designed rectifier the maximum equivalent input voltage is

$$V_{in\_e} = V_{bc}(wt = 0°) = \sqrt{3}V_m \cos 0° = 520V. \qquad 4.5$$

Here V_m=300V

The duty cycle in CCM is

$$D = 1 - \frac{V_{in\_e}}{V_o} = 0.4526, \qquad 4.6$$

Here V_0=950V

To ensure the DCM operation, the critical power at this moment should be equal to the maximum load [10], and then the equivalent boost inductor is obtained:

$$Le = \frac{V_0^{\ 2}}{2P_{o\_max}f} D(1-D)^2 = 15.29mH \qquad 4.7$$

Here, P_{0_max} =1kW, f=4 kHz. So the boost inductor L=$\frac{Le}{2} = 7.65mH$. With 20% margin the inductor could be selected as 6mH. In our model we have set the inductor value to 15mH.

4.3 The PWM module

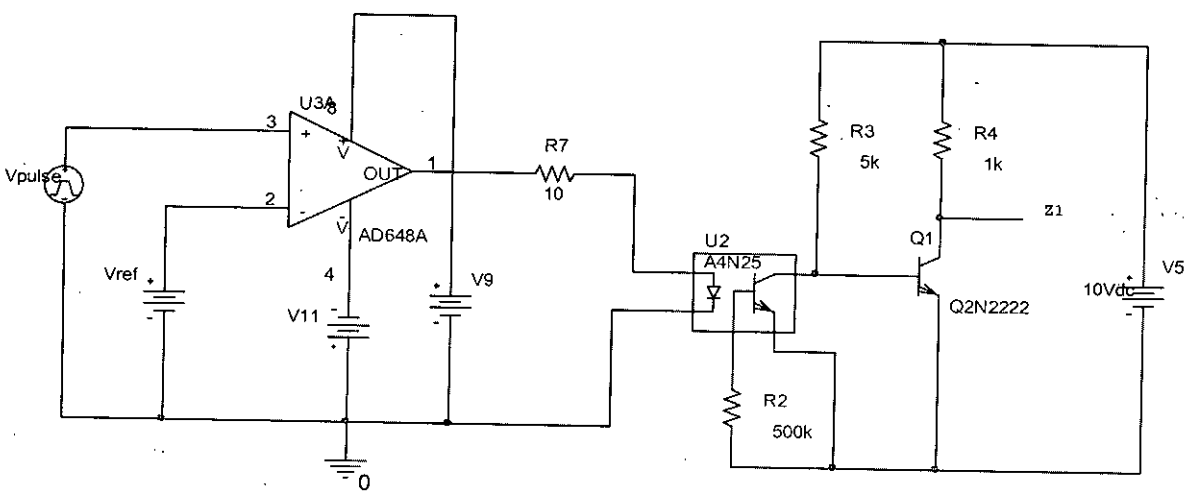
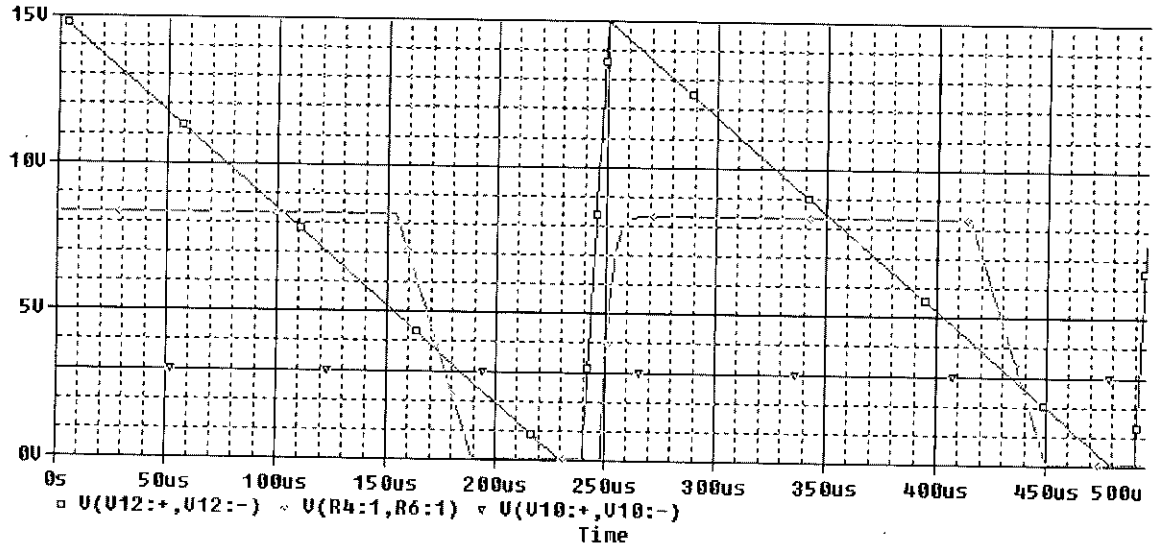


Figure 4.2: The PWM model

The PWM module consists of an opamp, an opto-coupler and BJT. Pin # 3 of the opamp is connected to a saw tooth wave varying from 15V to 0 Volts. For PWM, changing the reference voltage at pin # 2 will change the duty cycle. In Figure 4.3 we can see how the change in fixed reference at pin # 2 changes the duty cycle.



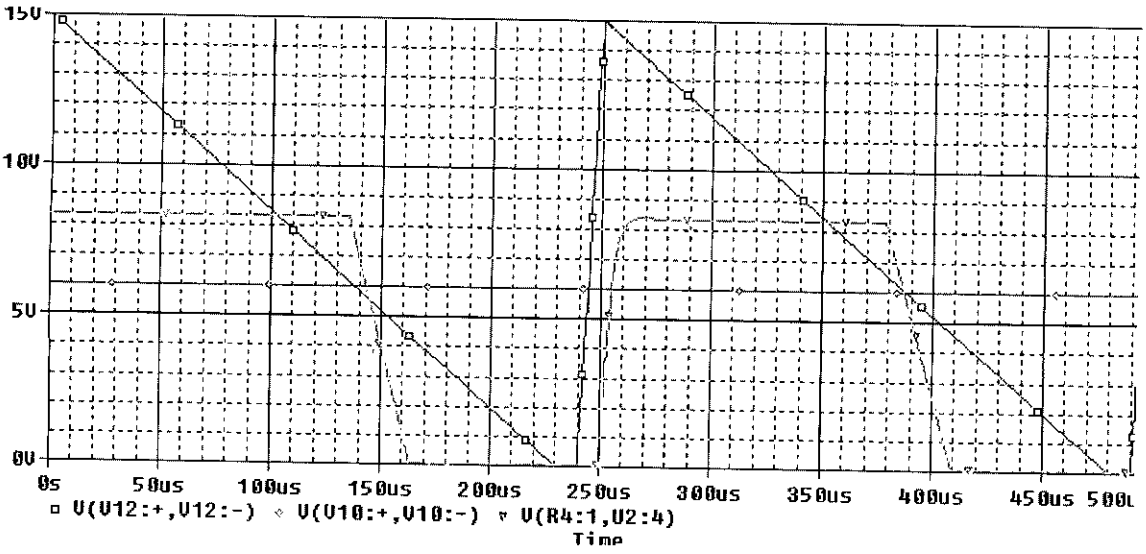


Figure 4.3: T_{on} time at output of opamp changes with change of reference voltage

4.4 Use of Opto-Coupler

An optocoupler, also called optoisolator, is an electronic component that transfers an electrical signal or voltage from one part of a circuit to another, or from one circuit to another, while electrically isolating the two circuits from each other. It consists of an infrared emitting LED chip, all enclosed in the same package. The silicon chip could be in the form of a photo diode, photo transistor, photo darlington, or photo SCR.

4.4.1 Functions of Opto-Coupler

- To isolate one section of a circuit from another, each section having different signal voltage levels to ensure compatibility between them.
- To prevent electrical noise or other voltage transients that may exist in a section of a circuit from interfering with another section when both sections have a common circuit reference. Noise or voltage transients can be caused by a poor printed circuit board layout.

4.4.2 Spectral Response of Silicon

Since silicon has a response to light (spectral response) that peaks at infrared wavelengths (between 800 and 950 nanometers), silicon devices are preferred as the photodetector section in optocouplers in conjunction with an infrared LED emitter (Figure 4.4). Matching the infrared LED to the silicon chip provides a maximum transfer of the desired electrical signal.

Different types of optocouplers have specific characteristics that determine suitability for each unique application. The simplest type is the optocoupler with a photo diode output section. The optocoupler output is often connected to an amplifier (or series of amplifiers) to change a low-level input voltage into an appropriate higher signal level.

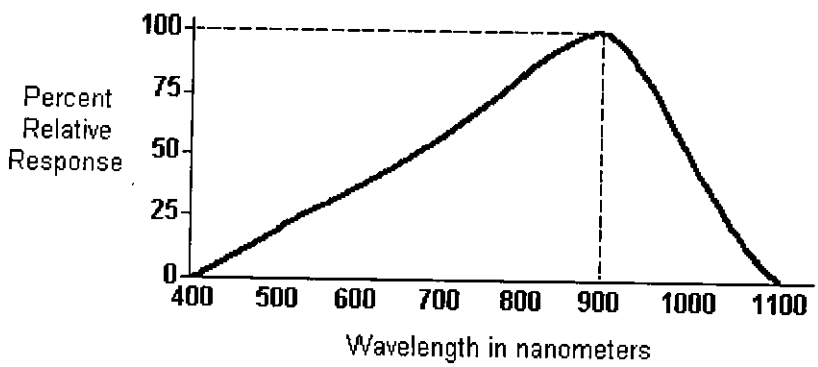


Figure 4.4: Spectral response curve of silicon

4.4.3 Principle of operation

When a forward bias voltage is applied to the input terminals of the LED (positive to the anode), an input current, I_{IN}, limited by the series resistor, Rs, will flow in the LED circuit. The current produces the infrared light emission at about 900 nanometers that impinges on the photosensitive silicon chip.

4.4.4 Photo Diode Output

• With light impinging on the silicon diode in Figure 4.5, its photovoltaic characteristic will create photo current, I_L or I_{OUT}, to flow in the silicon diode. With a load resistor, R_L, connected to the output terminals of the coupler, the photo current, I_{OUT}, will develop a voltage, V_L, across the load.

$$V_L = I_{OUT} \times R_L. \qquad 4.8$$

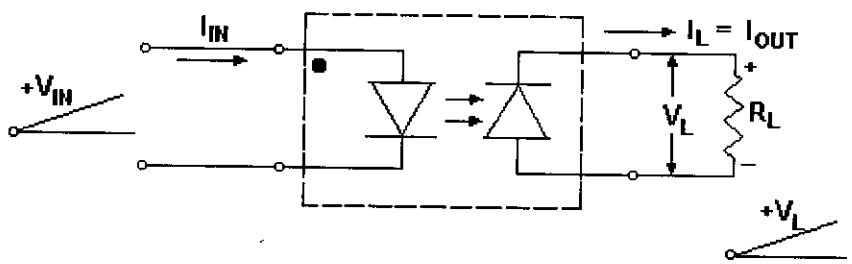


Figure 4.5: optocoupler with Photo Diode Output

• As the input signal, V_{IN}, varies; it will vary the intensity of the infrared light. The output current, I_{OUT}, will also change, causing the output voltage, V_L, to change in the same manner.

• As output current increases, output voltages will also increase, and vice-versa. A small change in input current will produce a proportionate change in output current. This characteristic of the optocoupler will act to couple low-level analog signals or small DC voltage variations with little or no distortion.

In the circuit of Figure 4.5, both signal coupling and input-to-output isolation is achieved, however, the current transfer ratio (CTR) of a diode output optocoupler is extremely low – about 10% to 15%. The term current transfer ratio (CTR) defines the relationship of output current, I_{OUT}, to input current, I_{IN}.

$$CTR = \frac{I_{OUT}}{I_{IN}} \qquad 4.9$$

The output voltage, V_L, can be coupled to the input of an amplifier to increase its amplitude to an appropriate level.

The input section of all optocouplers is an infrared LED, however, the output section can be different depending on the required application. The basic principle of operation is the same, regardless of the particular output section selected.

4.4.5 Photo Transistor output

Since the CTR of an optocoupler with a photo diode output is so low (10 to 15%) a preferred approach is to replace the diode chip with a silicon bipolar phototransistor (Figure 4.6).

The bipolar transistor, with its inherent current gain, β, will provide a considerably higher CTR (between 50% to 100%) depending on the beta of the phototransistor.

The base lead of the transistor can be reverse biased to reduce sensitivity, or forward biased to increase sensitivity, or left "floating" (disconnected).

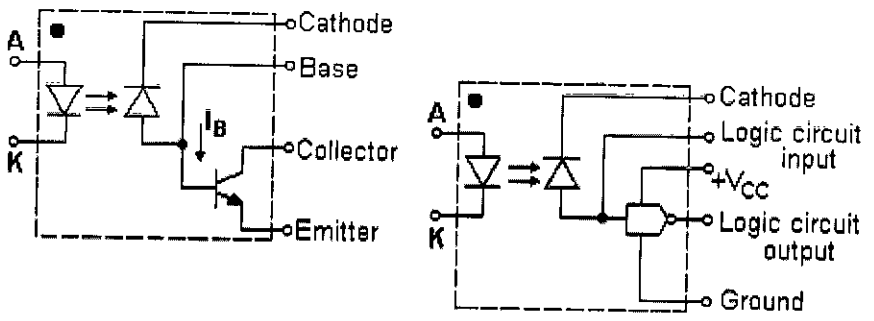


Figure 4.6: Variations of optocoupler output section

4.5 IGBT switching

The switching frequency is set to be 4 kHz. L1- C2, L2- C3 and L3- C4 are the filters for the corresponding phases set by the sources V1, V2 and V3. A single phase model has been analyzed first. The transfer function is mathematically solved that will by pass the high frequency components through the filter L1-C2 for Phase V1.

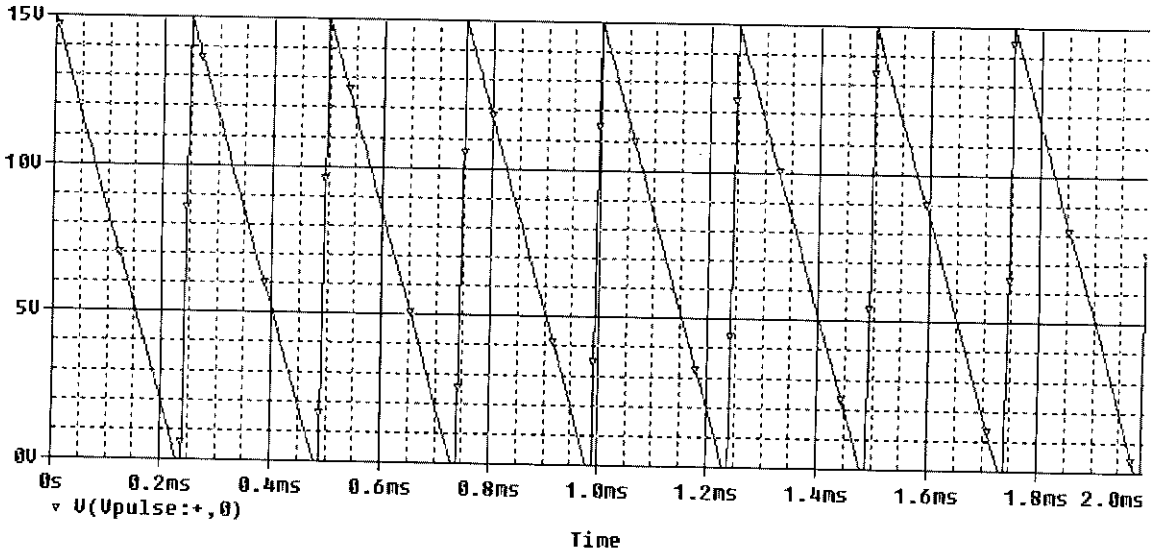


Figure 4.7: Input voltage of opamp

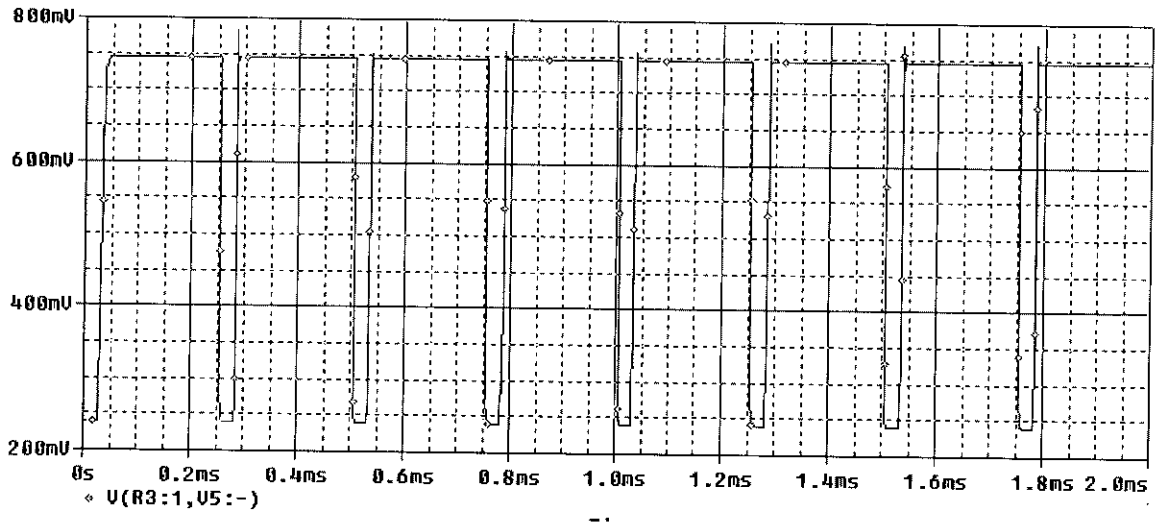


Figure 4.8: Voltage from optocoupler

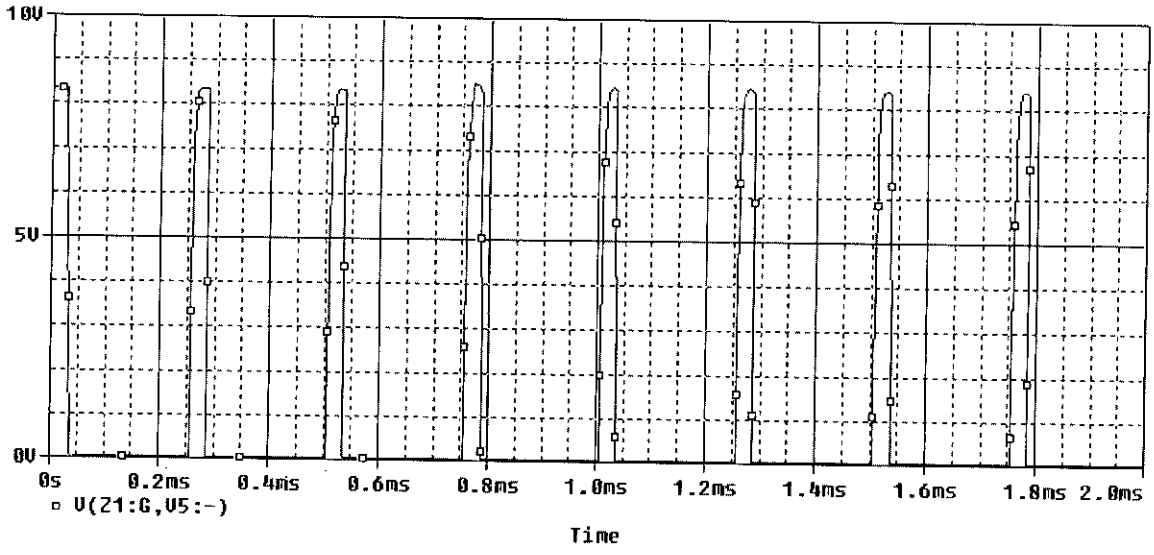


Figure 4.9: Voltage at the base of IGBT

As we can see from Figure 4.8, the voltage at the output of the opto coupler is in the range of 600 to 800 milivolts. To drive the IGBT a BJT amplifier is connected which increases the voltage level at about 8.5 Volts that is shown in Figure 4.9.

4.6 Design model for Transfer Function

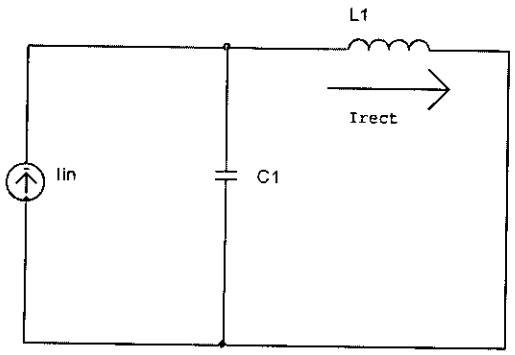


Figure 4.10: Design model for calculating transfer function

To eliminate the high frequency component generated by the active filter or switching we have made a simplified model of the input portion of the rectifier as shown in Figure 4.10.

$$\frac{Iin}{Irect} = \sqrt{\frac{(X_C)^2}{(X_L)^2 + (X_C)^2}} \qquad 4.10$$

Let us put Iin/Irect=1% then with the switching frequency of 4000Hz the product of LC becomes 1.583×10^{-11}. In our model we have set the value of L to be 15.83mH and C to be 0.1mF.

4.7 Design model for Input filter

The input filter is designed so that it resonates with the supply voltage frequency so that no other component is allowed to pass into the input of the rectifier.

$X_L = X_C$ *4.11*

$2\pi f L = 1/2\pi fC$

$LC = 1/4\pi^2 f^2$ *4.12*

As the frequency of the input voltage is 50Hz we have got the LC product to be 1×10^{-5}. In our model we have fixed L=50mH and C=200uF.

4.8 The complete model

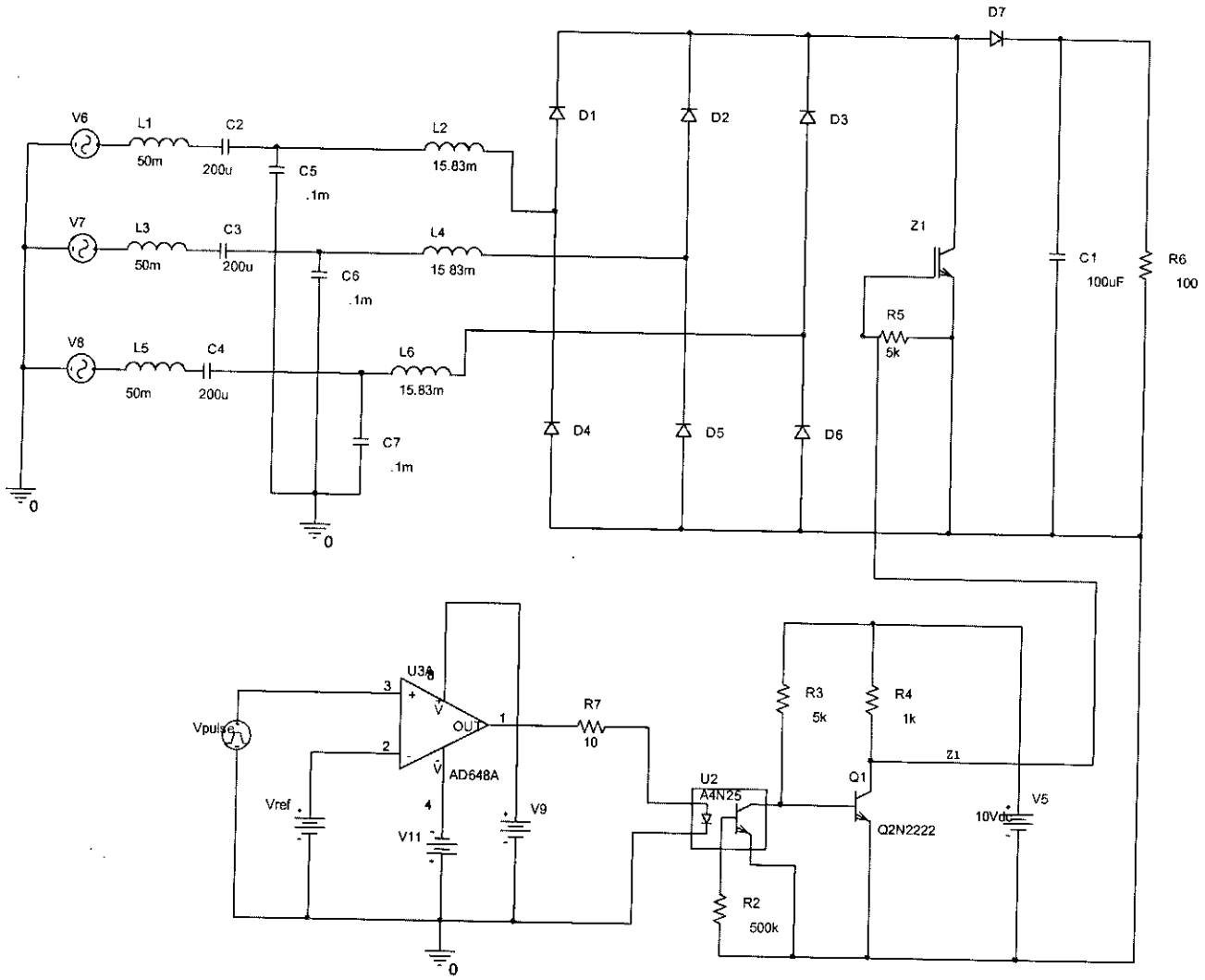


Figure 4.11 Three phase rectifier with switching, EMI and series LC filter

Table 4.1: Harmonics content in single switch boost rectifier with series LC filter

D	Is(mA)	I2	I3	I4	I5	I6	I7	I8	I9	I10	I11	I12	I13	I14	I15	I16	I17	I18	I19
	50	100	150	200	250	300	350	400	450	500	550	600	650	700	750	800	850	900	950
0.96	33000	109	13	31	290	19.75	45	15.5	13.4	12.15	3.8	10.28	10.5	8.8	8.2	7.7	7.75	6.82	6.18
0.88	32500	115	16	30	290	20	43	15	13.5	12	3.8	10.2	10.5	8.7	8.1	7.65	7.75	6.75	6.12
0.85	32000	93	73	30	300	20	44	15	14.1	11.7	0	10	9.8	8.45	7.75	7.5	7.8	6.75	5.9
0.80	31000	80	112	25	283	19	46	14	12.65	10.5	4.8	9	8.15	7.65	7.05	6.7	7.2	5.9	5.45
0.72	30000	180	50	10	280	18	45	11.6	11.4	9.8	6	7.8	5.1	6.6	6.2	5.8	6.4	5.2	5.33
0.64	19700	50	25	0	250	5	49	3.2	2.5	2.5	9	2	5	1.68	1.56	1.58	1.57	1.32	0.85
0.55	16000	74	60	10	242	3	46	1.85	0.55	1	6.4	6	3.4	0.46	0.5	0.39	1.25	0.3	0.078
0.47	13500	80	110	20	230	3	40	1.85	1.6	1.65	6.1	1.4	3.9	1	1.05	0.8	1.6	0.8	1.08
0.39	15800	300	1320	25	185	19.3	15.6	3.8	6.1	3.75	6.25	2.9	1.38	2.1	0.4	1.7	1.36	1.38	6.8
0.32	11700	390	100	29	179	10	27.5	6	3.6	6.56	6.58		6.58	3.9	3.15	3.6	3.1	3	2.65
0.26	10600	370	30	20	165	7.5	27	4	4.9	5.1	4.4	3.35	2.5	2.85	2.81	2.7	2.05	2.24	1.93
0.18	9900	290	20	19	160	7	22	4.45	4.2	4.7	3.2	3.25	2	2.71	2.2	2.48	1.85	2.15	1.65
0.12	9200	128	25	8	141	3.4	18.8	1.8	1.72	1.3	4.8	1	2.45	0.9	0.82	0.75	1.14	0.65	1
0.06	8900	110	27	10	141	4	13.8	2.7	2.5	2	5.5	1.72	2	1.45	1.35	1.225	1.68	1.09	1.05

Table 4.2 Harmonics content in single switch boost rectifier without series LC filter

D	Is(mA)	I2	I3	I4	I5	I6	I7	I8	I9	I10	I11	I12	I13	I14	I15	I16	I17	I18	I19
	50	100	150.0	200	250	300	350	400	450	500	550	600	650	700	750	800	850	900	950
0.80	32330	71	75.6	39.0	2510	45.0	551.0	27.0	66.0	17.7	430.0	14.30	120.0	13.7	30.0	14.2	150.0	8.9	63.5
0.76	31250	47	192.0	58.8	2500	40.8	500.0	30.0	24.6	13.8	421.0	14.40	91.0	9.5	38.0	14.1	141.0	9.3	41.0
0.70	30000	161	71.0	79.0	2272	36.0	520.0	52.0	51.0	12.0	363.0	13.00	208.0	2.0	13.6	20.0	112.0	11.5	99.0
0.63	25600	218	82.0	130.0	1950	34.4	920.0	13.0	24.8	18.2	382.0	6.50	270.0	8.7	11.3	14.2	158.0	11.4	115.0
0.57	20000	156	47.0	66.0	2050	4.2	1020.0	2.0	15.0	11.5	397.0	9.00	270.0	12.5	8.5	8.0	161.0	5.0	125.0
0.51	15000	170	43.0	48.0	2060	35.0	645.0	23.0	18.5	26.0	286.0	3.50	220.0	9.0	17.0	12.5	95.0	13.0	95.0
0.44	12700	180	20.0	50.0	2000	30.0	600.0	24.0	24.0	5.0	283.0	12.5	150.0	12.0	8.0	8.0	85.0	4.0	90.0
0.38	11000	170	35.0	28.0	1850	46.0	550.0	13.0	10.6	22.5	200.0	11.0	190.0	21.0	13.0	8.5	118.0	10.7	96.0
0.31	8000	200	21.0	40.0	1610	18.0	700.0	12.0	6.0	11.5	222.0	6.0	168.0	8.0	5.0	5.6	82.0	5.2	74.0
0.25	9200	220	33.0	150.0	1400	27.0	570.0	35.0	13.4	9.0	200.0	180.0	155.0	11.0	7.0	2.2	70.0	5.2	60.0
0.18	9800	300	30.0	165.0	1550	8.0	570.0	34.0	4.0	14.5	222.0	4.5	190.0	11.2	4.2	6.0	93.0	2.8	72.0
0.12	9000	220	50.0	100.0	1350	8.0	480.0	18.0	11.0	5.0	202.0	1.8	175.0	7.0	1.2	2.6	78.0	1.9	69.0
0.06	8800	210	25.0	100.0	1300	12.5	300.0	8.5	12.5	11.5	217.0	8.2	148.0	8.0	5.0	9.4	67.0	5.5	61.0

Table 4.3: THD and Efficiency with duty cycle in single switch boost rectifier with series LC filter

D	THD	P.F	Vout	Ifund	Pout	Pin	efficiency
0.960	0.96	0.8090	450.00	33000	2025	12013.65	0.17
0.880	0.98	0.8090	412.00	32500	1697.44	11831.63	0.14
0.848	1.03	0.8090	640.00	32000	4096	11649.6	0.35
0.800	1.04	0.8090	800.00	31000	6400	11285.55	0.57
0.720	1.14	0.8900	930.00	30000	8649	12015	0.72
0.640	1.33	1.0000	900.00	19700	8100	8865	0.91
0.548	1.65	1.0000	820.00	16000	6724	7200	0.93
0.472	2.01	1.0000	750.00	13500	5625	6075	0.93
0.316	3.78	0.9500	635.00	11700	4032.25	5001.75	0.81
0.260	3.85	0.8400	590.00	10600	3481	4006.8	0.87
0.180	3.37	0.7289	550.00	9900	3025	3247.25	0.93
0.122	2.10	0.6100	510.00	9200	2601	2350.278	0.99
0.056	2.04	.6100	483.00	8900	2332.89	2273.639	0.99

Table 4.4: THD and Efficiency with duty cycle in single switch boost rectifier without series LC filter

D	THD	P.F	Vout	Ifund(A)	Pout	Pin	efficiency
0.8000	8.10	0.8440	632.00	32.33	3994.24	12278.93	0.3253
0.7580	8.32	0.8607	795.00	31.25	6320.25	12103.59	0.5222
0.6984	7.94	0.8952	928.00	30	8611.84	12085.2	0.7126
0.6340	8.72	0.9500	964.00	25.6	9292.96	10944	0.8491
0.5680	11.78	0.9891	896.00	20	8028.16	8901.9	0.9018
0.5100	14.67	0.9990	821.00	15	6740.41	6743.25	0.9996
0.4392	16.73	0.9939	753.00	12.7	5670.09	5680.139	0.9982
0.3840	17.86	0.9759	695.00	11	4830.25	4830.705	0.9999
0.3120	22.41	0.9510	645.00	8	4160.25	3423.6	0.9758
0.2500	17.06	0.8910	600.00	9.2	3600	3688.74	0.9759
0.1848	17.52	0.8090	555.00	9.8	3080.25	3567.69	0.8634
0.1220	16.47	0.7070	520.00	9	2704	2863.35	0.9443

4.9 Relation of efficiency (η) with duty cycle (D)

To find out the relationship between efficiency (η) with duty cycle (D), let us redraw the equivalent circuit for a single phase.

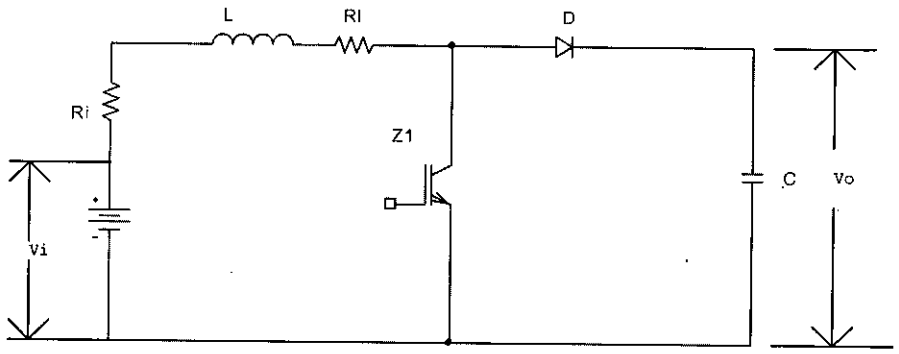


Figure 4.12: single phase equivalent model for 3ϕ rectifier with switching

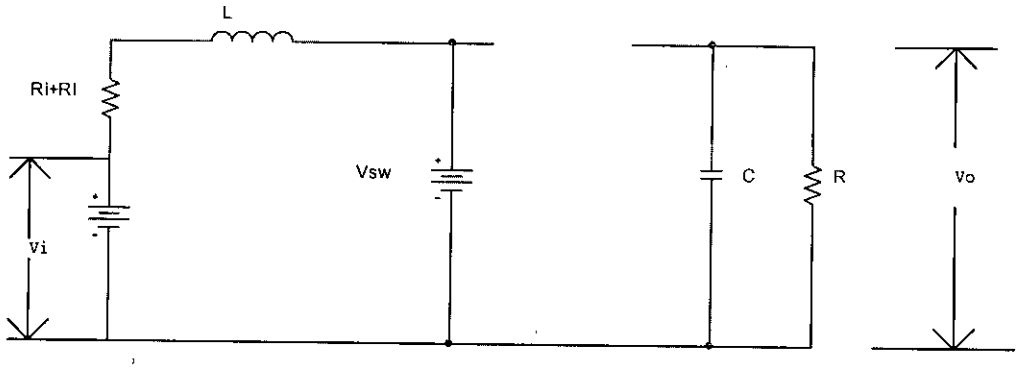


Figure 4.13: single phase equivalent model for 3ϕ Rectifier with switching when switch is ON

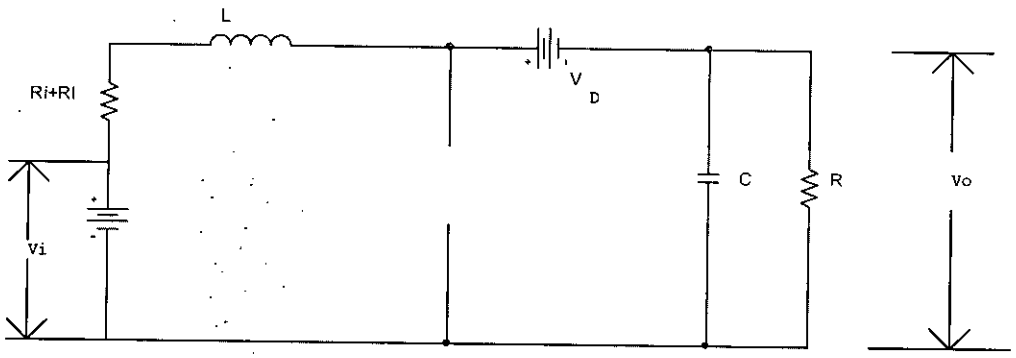


Figure 4.14: single phase equivalent model for 3ϕ Rectifier with switching when switch is OFF

For ideal case the relation between voltage gain and duty cycle is

$$\frac{V_o}{V_i} = \frac{1}{1-D} \qquad 4.13$$

But for ideal and practical case we have to consider the relation between current gain with duty cycle (D), which is

$$\frac{I_o}{I_i} = (1-D) \qquad 4.14$$

Now we can define efficiency as the ratio of output power to input power. So

$$\eta = \frac{I_o{}^2 R}{I_o{}^2 R + losses} \qquad 4.15$$

The loss term includes resistive loss $(R_l+R_i)I_{in}{}^2$, switching loss $V_{sw}DI_{in}$ and diode loss $V_D(1-D)I_{in}$ etc. Here

$$I_{in} = D * i_{in} \qquad 4.16$$

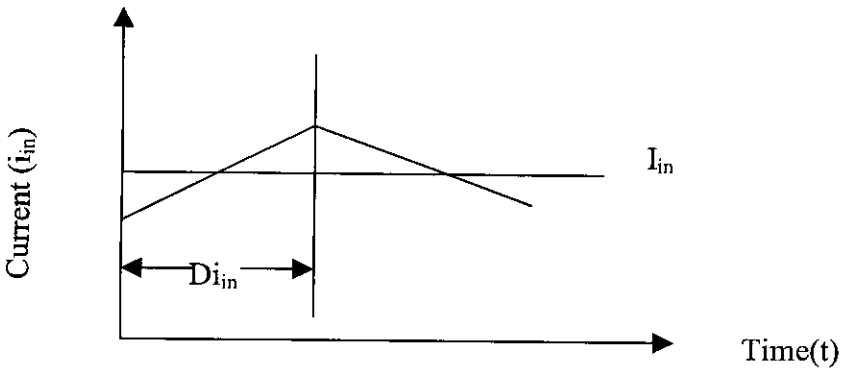


Figure 4.15:The varying input current as the switch turns ON and OFF.

So the equation becomes:

$$= \frac{I_o^2 R}{I_o^2 R + I_{in}^2 (R_l + R_i) + V_{sw} D I_{in} + V_D (1-D) I_{in}} \qquad 4.17$$

$$= \frac{1}{1 + (I_{in}/I_o)^2 (R_l + R_i) + V_{sw} D I_{in} / I_o^2 R + V_D (1-D) Iin / I_o^2 R} \qquad 4.18$$

$$= \frac{1}{1 + 1/(1-D)^2 (R_l + R_i) + V_{sw} D / R(1-D) I_o + V_D / I_o R} \qquad 4.19$$

So for a particular load this equation can be written as

$$\eta = \frac{1}{1 + A/(1-D)^2 + B/(1-D) + C} \quad , \qquad 4.20$$

Where A, B and C are constant for a fixed load.

So we can see that η does not have any linear relationship with duty cycle D rather it will perform very well in a particular range duty cycle.

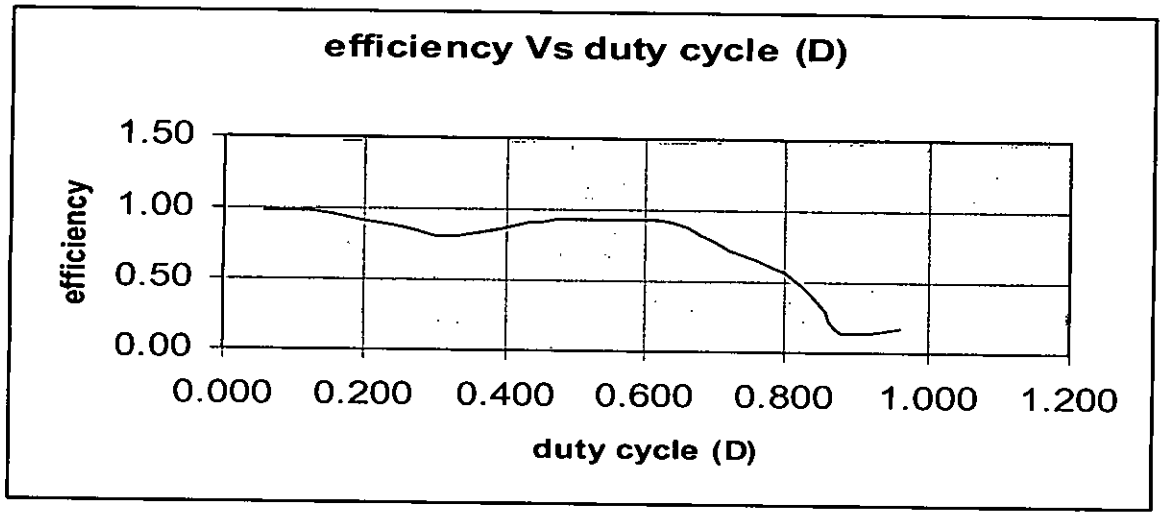


Figure 4.16: The efficiency Vs duty cycle curve with the proposed model

The curve for efficiency (η) Vs. duty cycle (D) shows that for duty cycle in the range of 0.4 to 0.6 gives the optimum efficiency range for the module. After 0.6 duty cycle (D) the efficiency drops down rapidly.

4.10 Relation of output voltage (V_O) with duty cycle (D)

To find out the relation between output voltage (V_O) with duty cycle (D) we have to analyze the equation for efficiency first. We know,

$$\eta = \frac{outputpower}{inputpower} \qquad 4.21$$

$$= \frac{V_O I_O}{V_{IN} I_{IN}} \qquad 4.22$$

$$= \frac{V_O}{V_{IN}} * \frac{I_O}{I_{IN}} \qquad 4.23$$

$$= \frac{V_O}{V_{IN}} * (1-D) \qquad 4.24$$

So,

$$\frac{V_O}{V_{IN}} = \frac{\eta}{(1-D)} \qquad 4.25$$

Putting the value of efficiency we get

$$\frac{V_O}{V_{IN}} = \frac{1}{1+\frac{(R_l+R_i)}{(1-D)^2}+\frac{V_{sw}D}{R(1-D)I_O}+\frac{V_D}{I_O R}} * \frac{1}{(1-D)} \qquad 4.26$$

$$= \frac{1}{(1+\frac{V_D}{I_O R})(1-D)+\frac{R_l+R_i}{(1-D)}+\frac{V_{sw}D}{RI_O}} \qquad 4.27$$

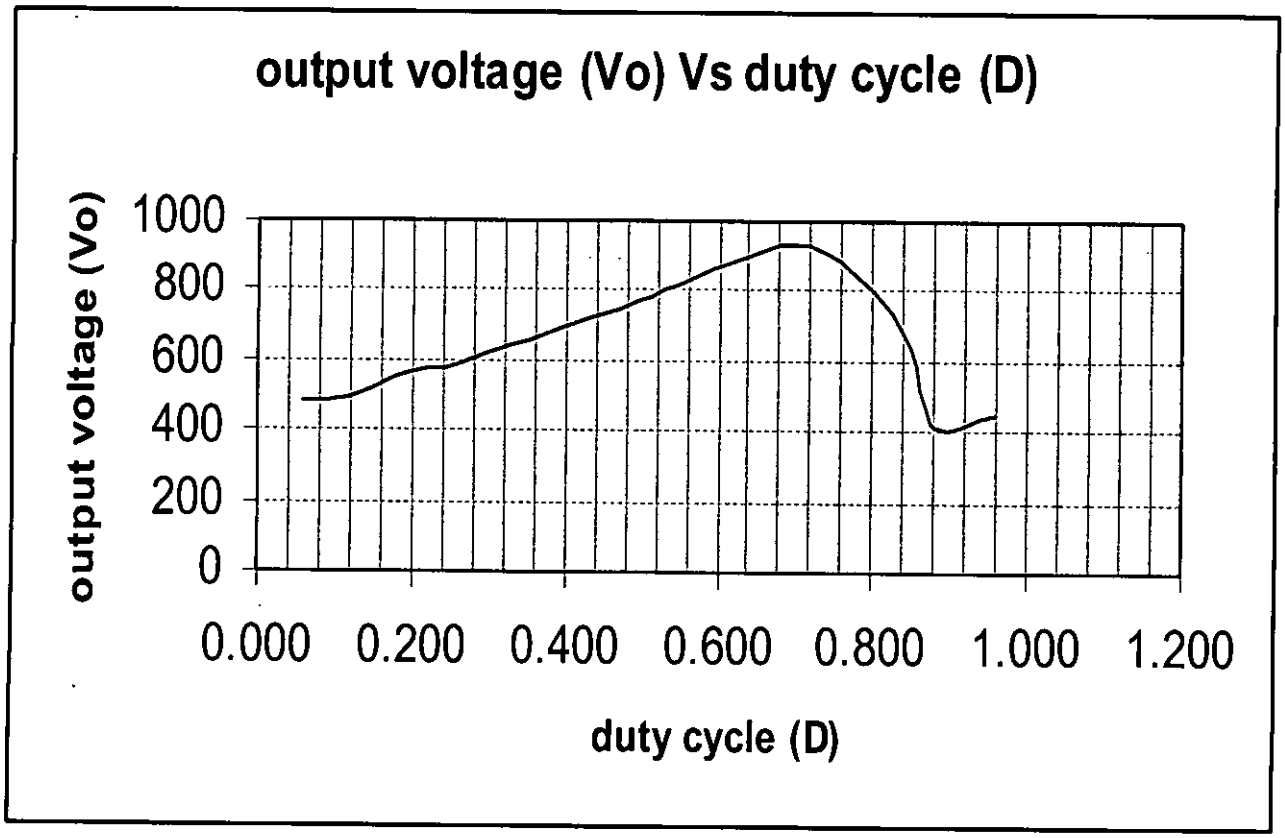


Figure 4.17: The output voltage (V_o) Vs. duty cycle (D) curve with the proposed model

From Figure 4.17 we can see that the relationship between output voltage (V_o) Vs. duty cycle (D) is not linear. Rather for a certain range, the output voltage increases as the duty cycle increases. But after the duty cycle of 0.7 the output decreases as the duty cycle increases because of the losses incorporates as we can see in the equation.

4.11 Comparison of THD values

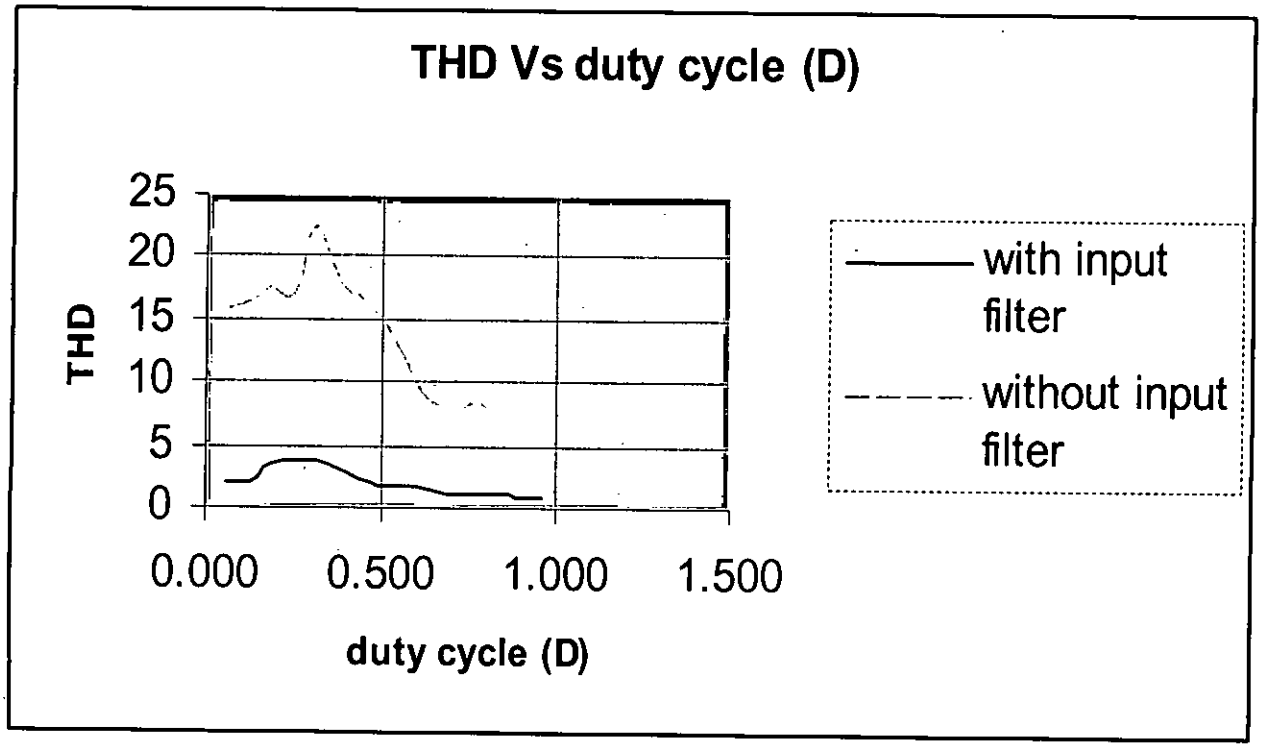


Figure 4.18: the comparison of THD values for with and without input LC filter

From Figure 4.18 we can see that the THD value has been reduced a lot with the introduction of series LC filter at the input side. Without input LC series filter at the input, the THD value could be as high as 22% and the lowest possible value is around 8% only at 0.7 duty cycle. But with the introduction of series LC filter at the input the overall THD value has been reduced to be less than 4% and with duty cycle more than 0.5 the value is as low as 1.7%.

4.12 Relation of Efficiency (η) with output voltage (V_o)

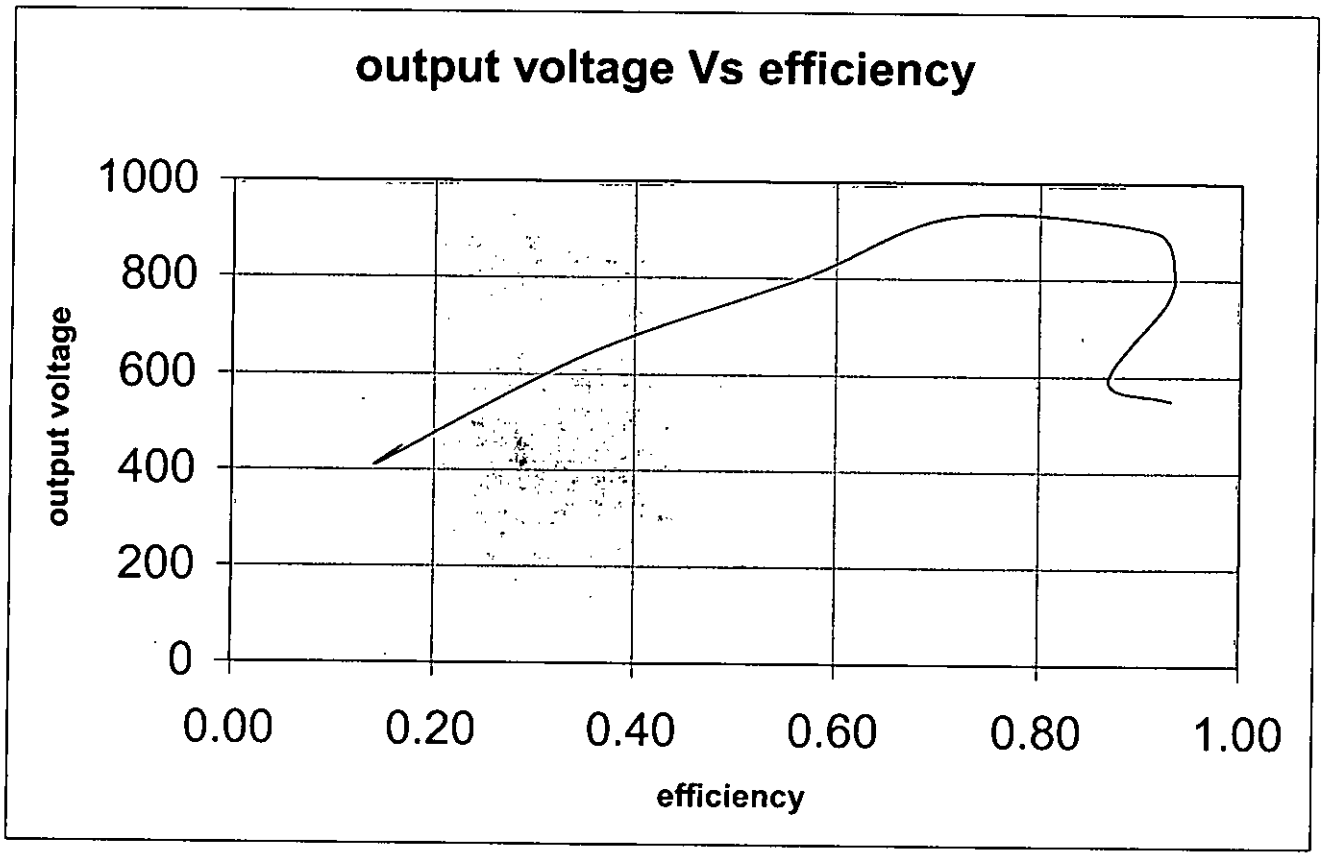


Figure 4.19: output voltage (V_o) Vs. efficiency for complete range of duty cycle (D)

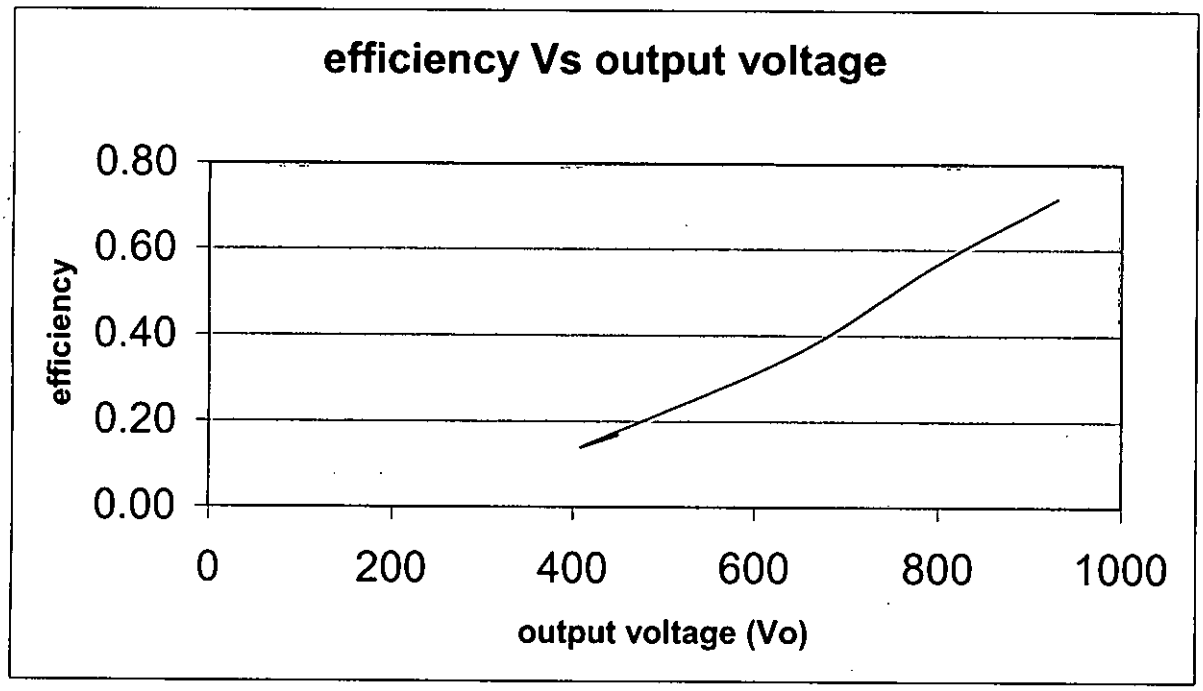


Figure 4.20: Efficiency Vs. output voltage (V_o) for duty cycle (D) more than 0.5

In Figure 4.19 the complete curve for output voltage Vs. efficiency has been given. Then in Figure 4.20 a portion of the curve is redrawn. From Figure 4.20 we can see that efficiency (η) is almost linearly dependent on output voltage (V_o) .As the voltage increases the efficiency also increases up to certain range of duty cycle. For more than 0.5 duty cycle the efficiency is almost linearly dependent on the output voltage. But output voltage Vs. duty cycle do not have linear relationship for the complete range of duty cycle.

In the next two articles the simulation results of input currents of different rectifier circuits are given. In 4.13 the input current of a three phase rectifier with PWM module and EMI filter is given. Then in 4.14 the input current for the proposed rectifier with additional series resonating LC filter is given to compare the difference in input currents.

4.13 Input current with only EMI filter

In this section the input currents for three-phase rectifier with only EMI filter for different duty cycles are shown from Figure 4.21 to Figure 4.30.

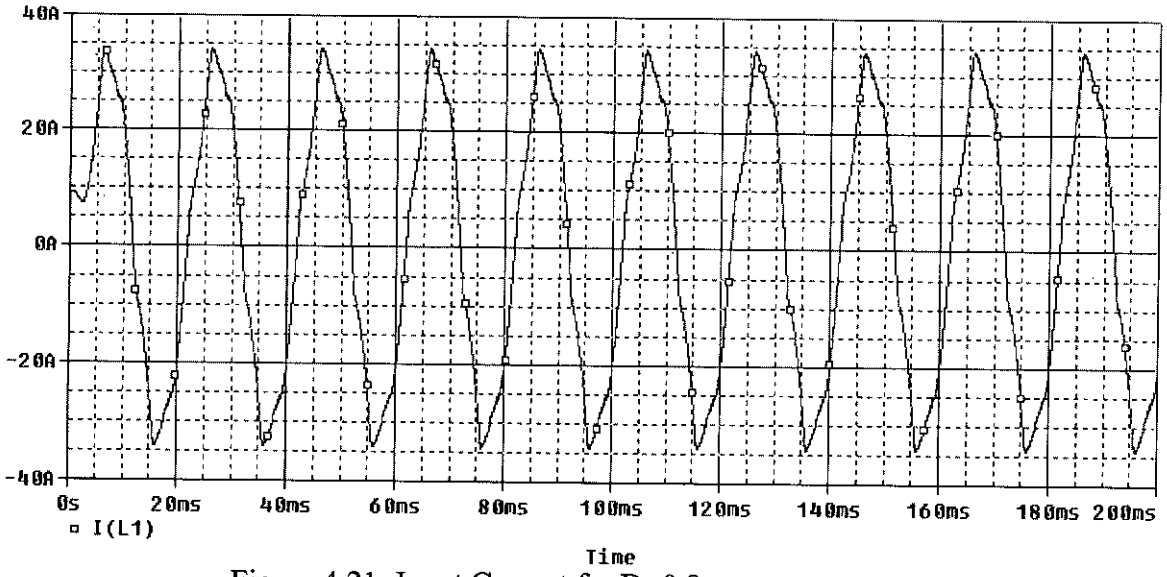


Figure 4.21: Input Current for D=0.8

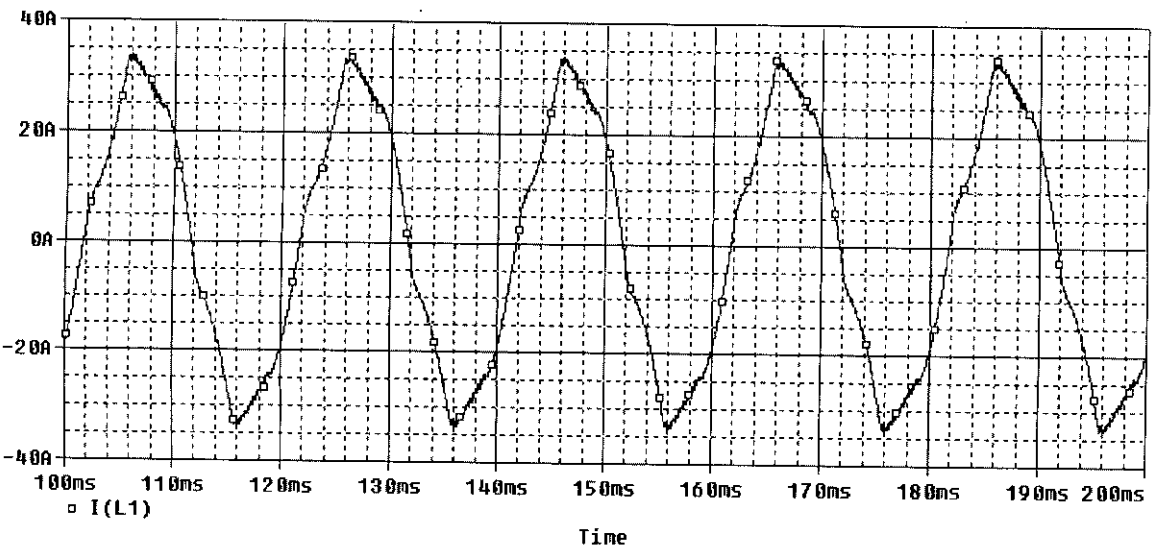


Figure 4.22: Input Current for D=0.76

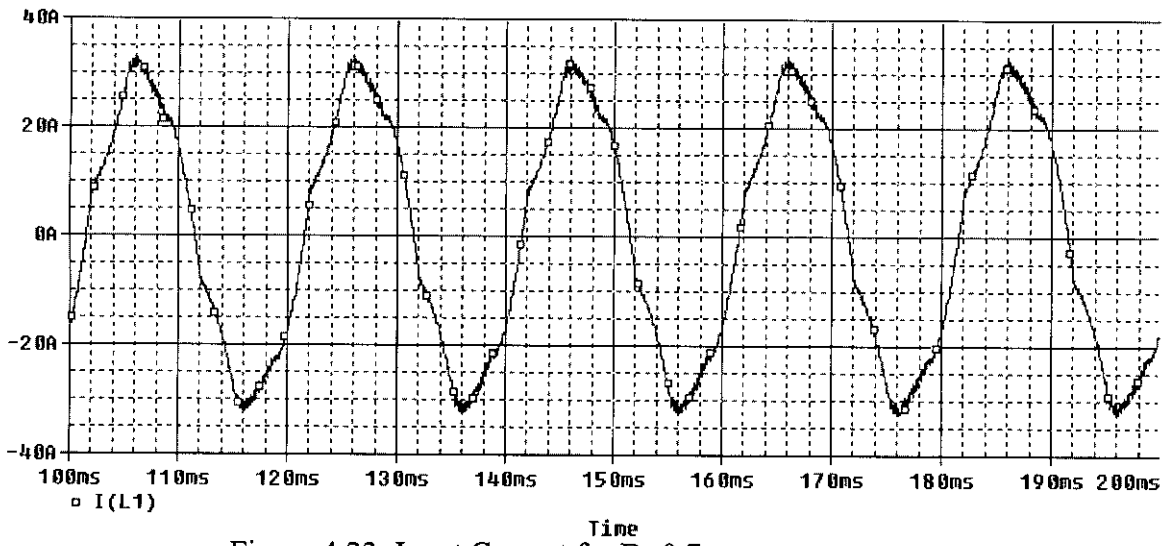


Figure 4.23: Input Current for D=0.7

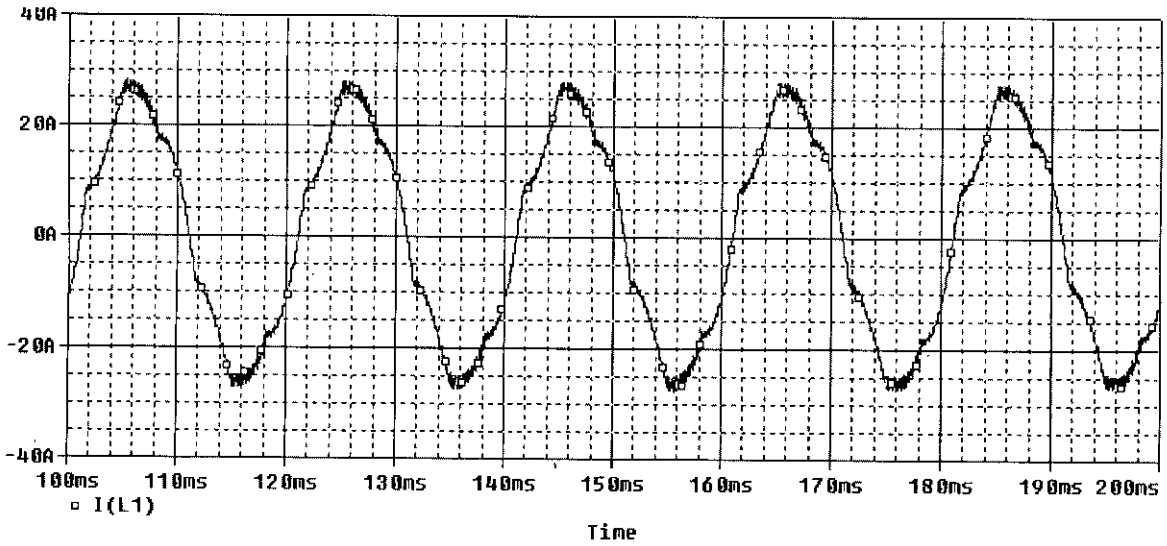


Figure 4.24: Input Current for D=0.632

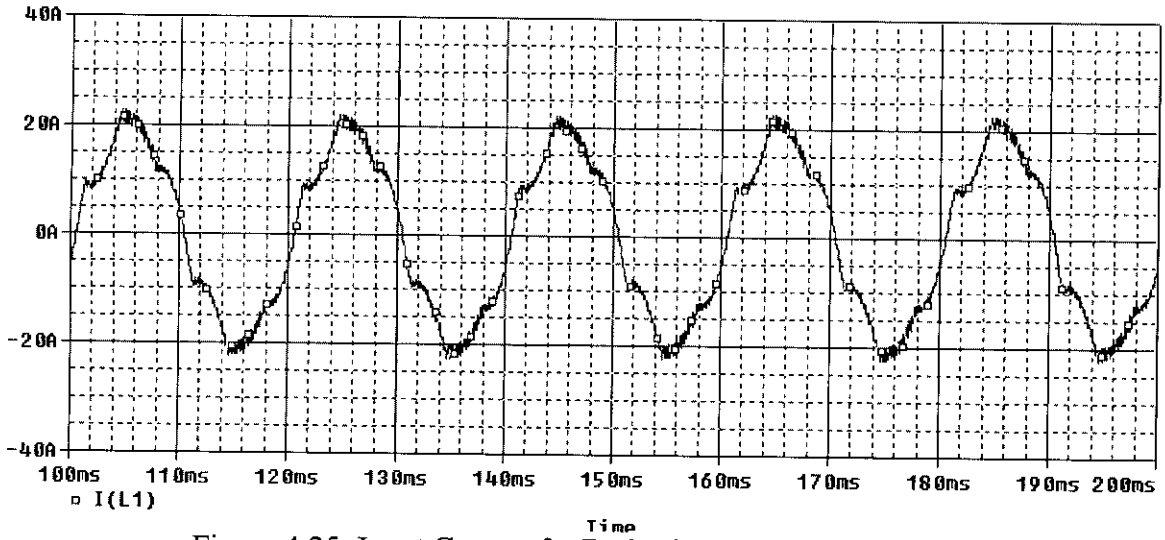


Figure 4.25: Input Current for D=0.56

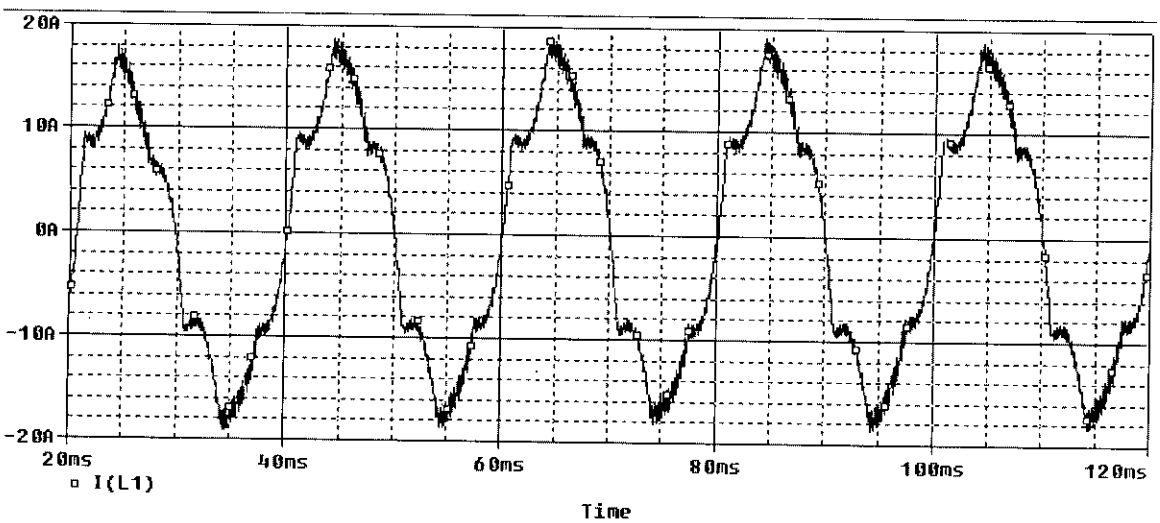


Figure 4.26: Input Current for D=0.5

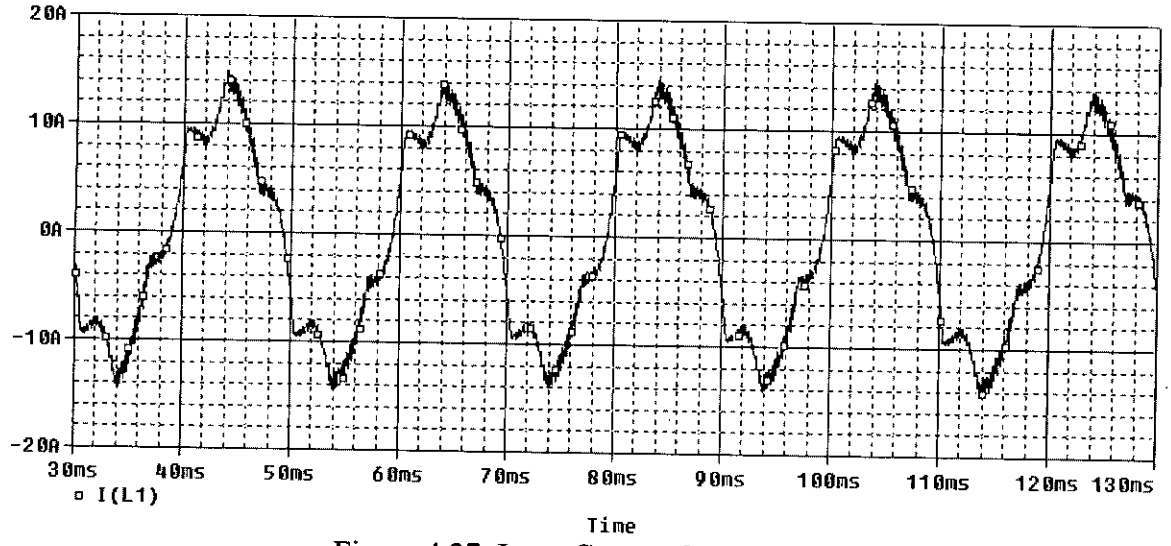


Figure 4.27: Input Current for D=0.36

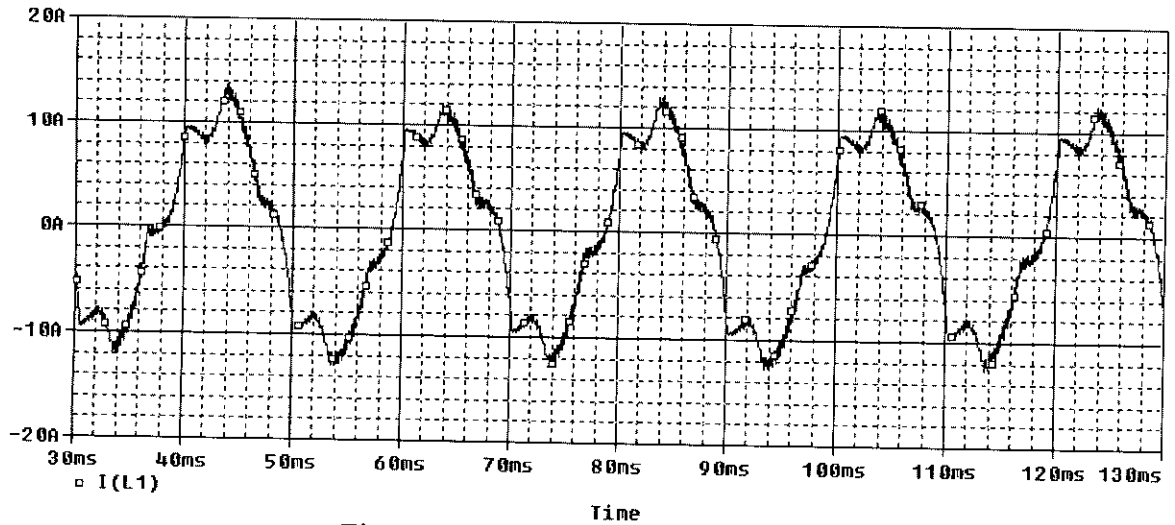


Figure 4.28: Input Current for D=0.31

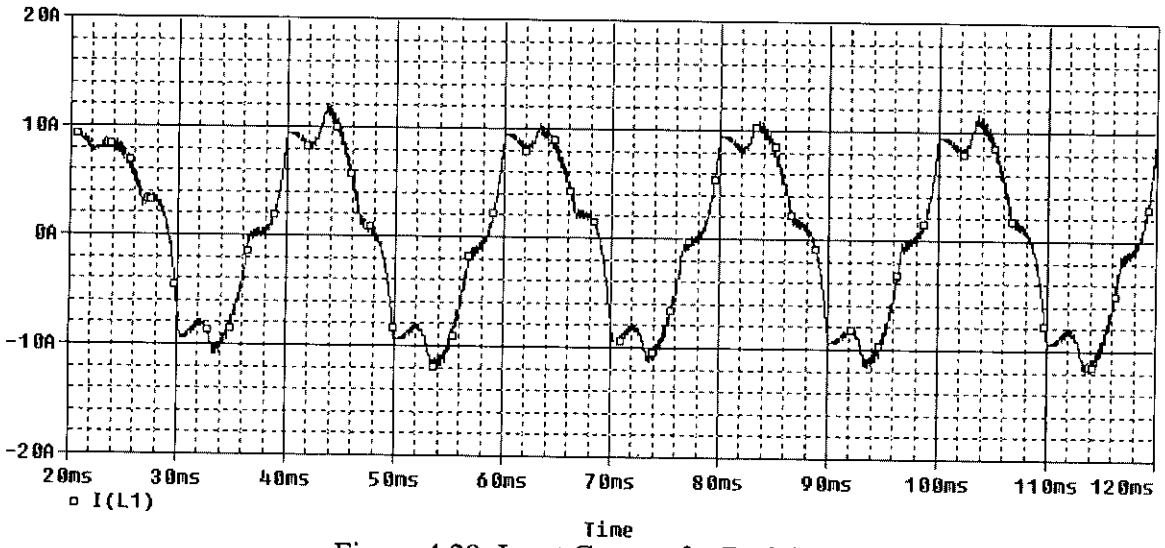


Figure 4.29: Input Current for D=0.24

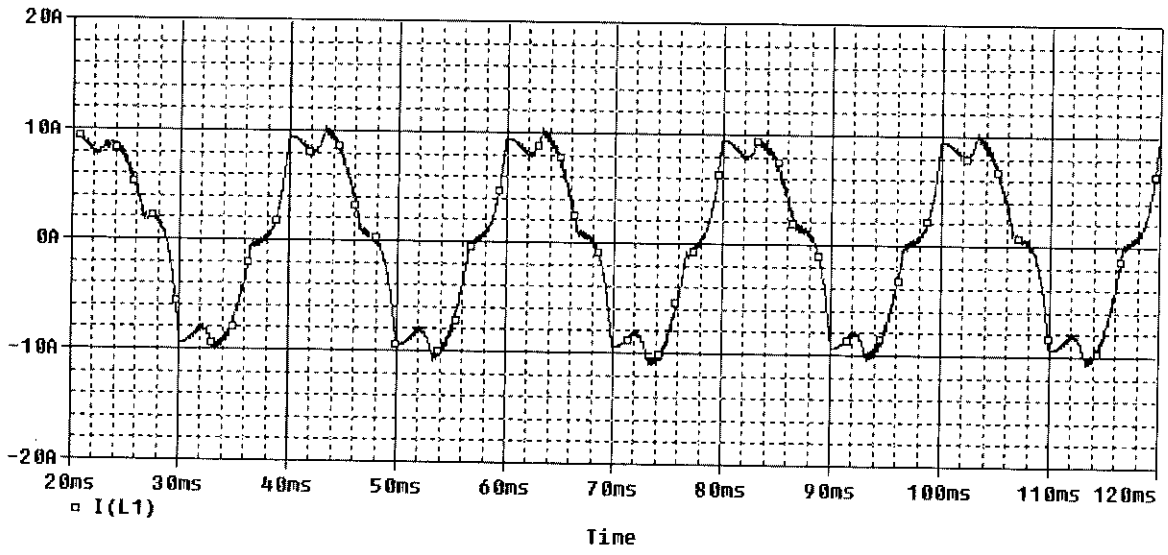


Figure 4.30: Input Current for D=0.188

So observing the Figures we see that even with the added EMI filter to the rectifier the input current is not sinusoidal. Rather it contains low frequency components.

4.14 Input Current with Resonating and EMI input Filter

In this section the simulated results for input currents with added EMI filter and series resonating LC filter are shown from Figure 4.31 to Figure 4.42 for different duty cycles.

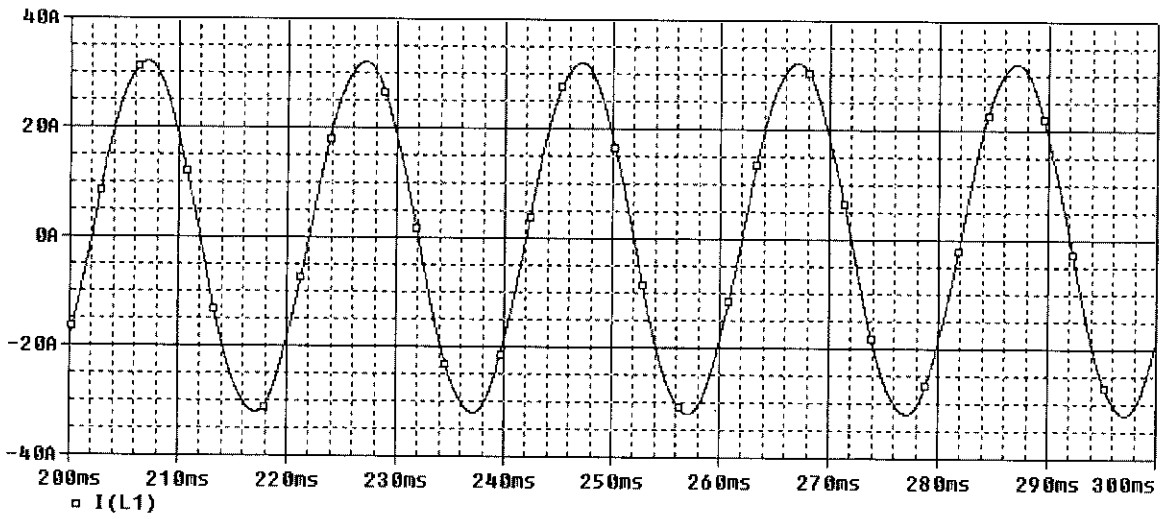


Figure 4.31: Input current for D=0.84

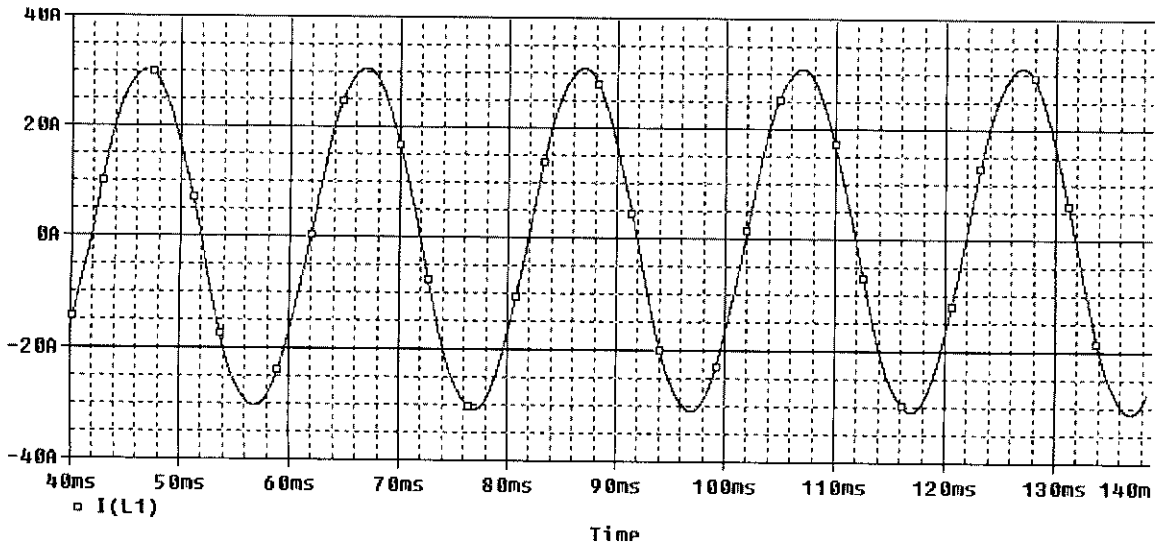


Figure 4.32: Input current for D=0.8

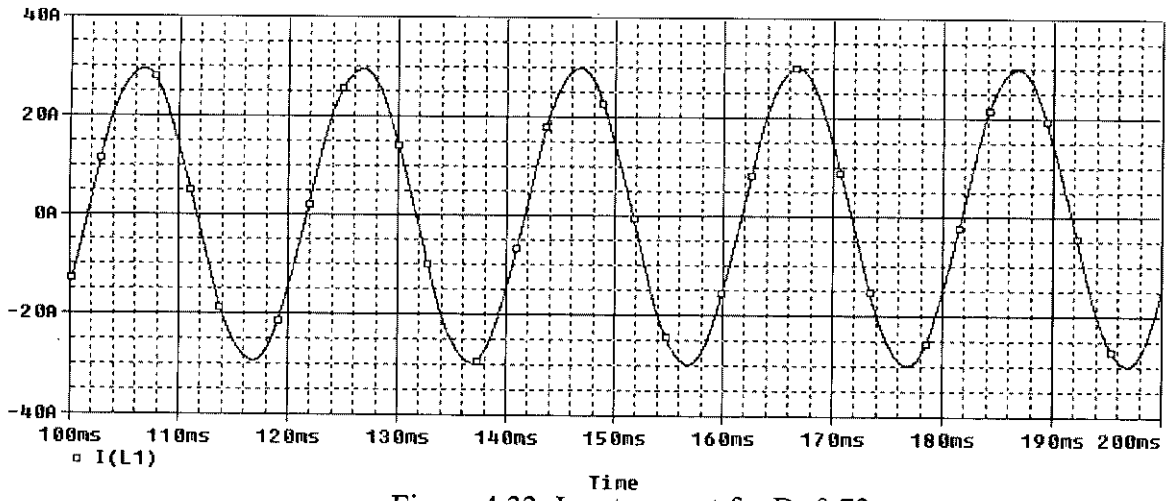


Figure 4.33: Input current for D=0.72

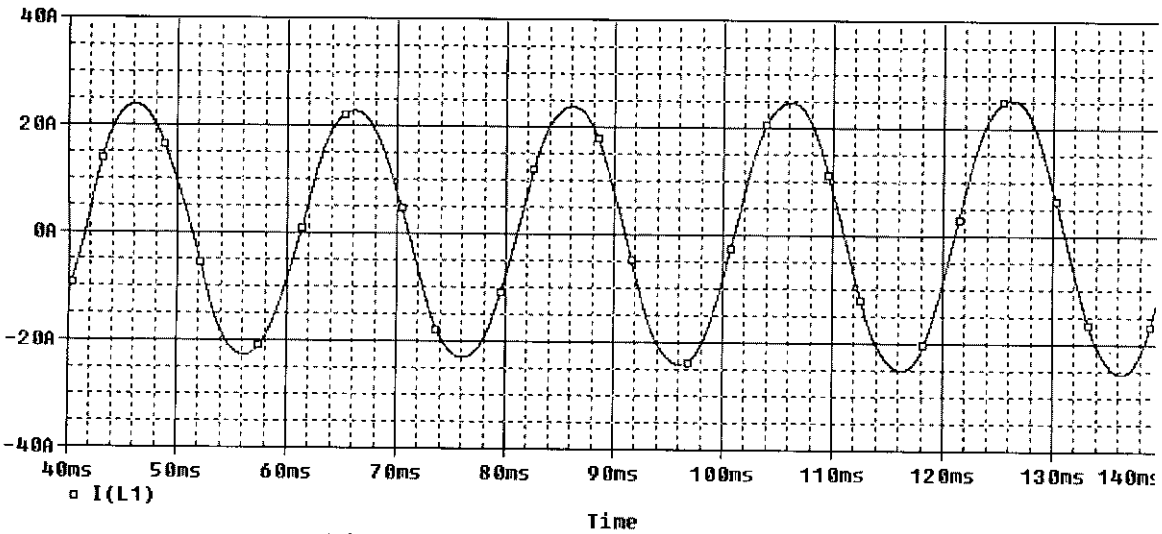


Figure 4.34: Input current for D=0.672

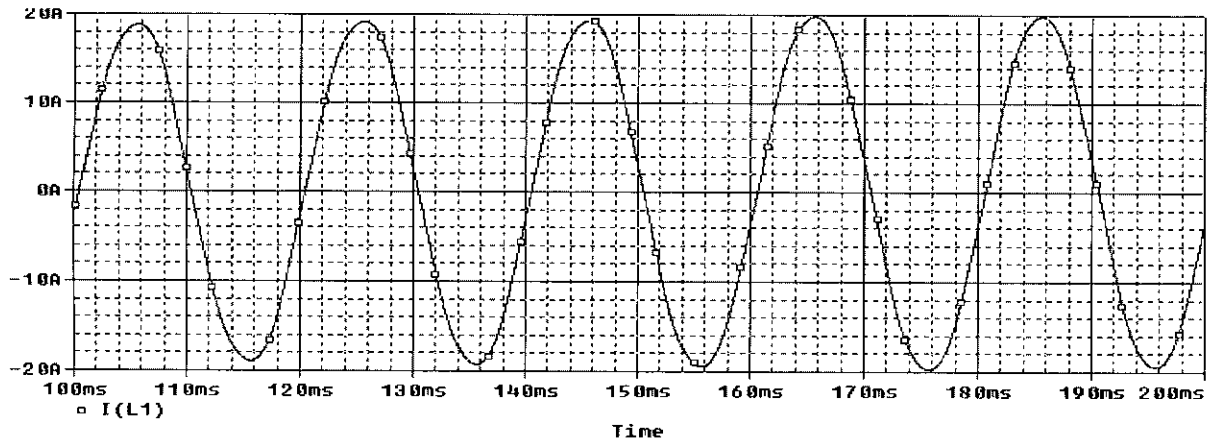


Figure 4.35: Input current for D=0.6

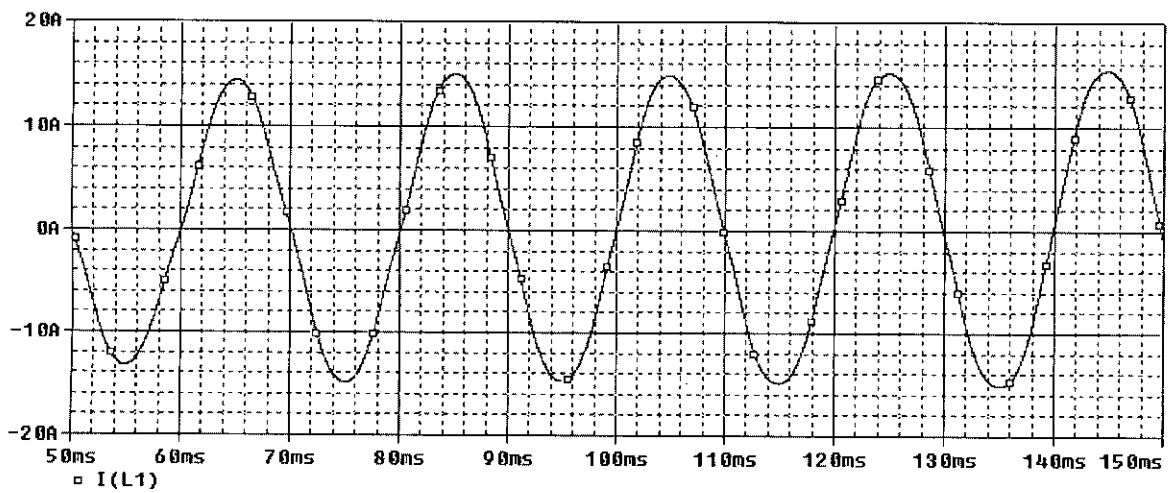


Figure 4.36: Input current for D=0.56

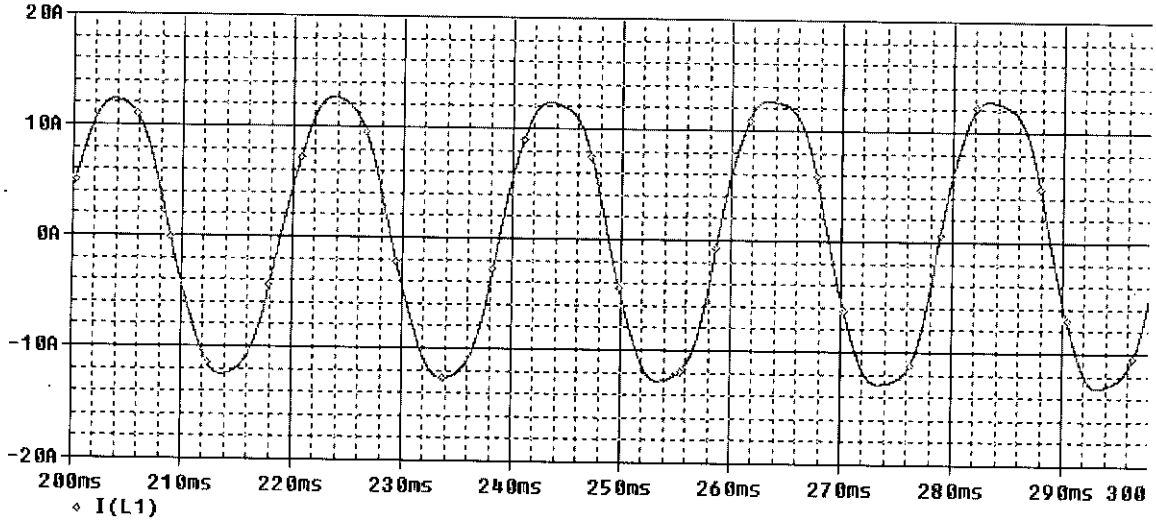


Figure 4.37: Input current for D=0.52

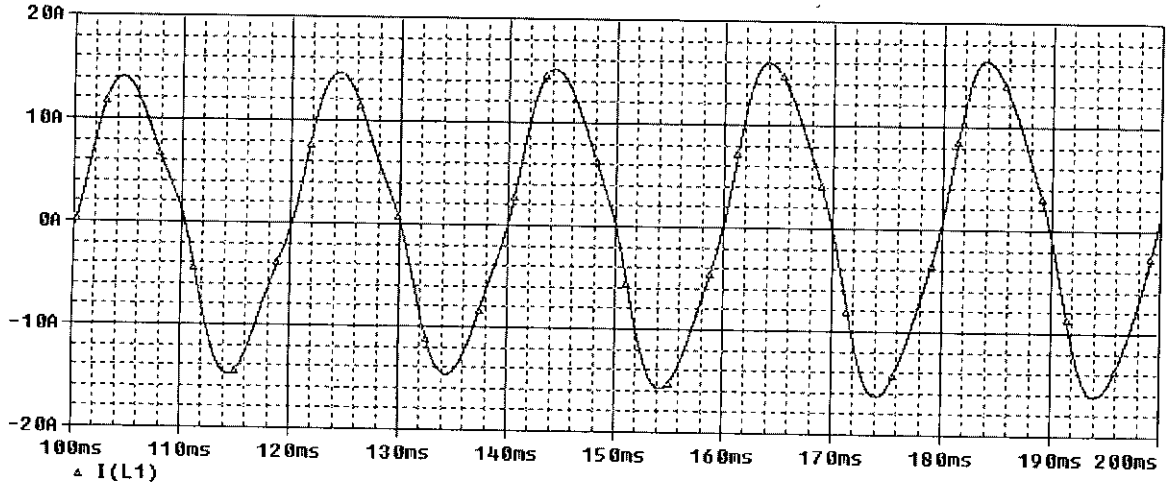


Figure 4.38: Input current for D=0.4

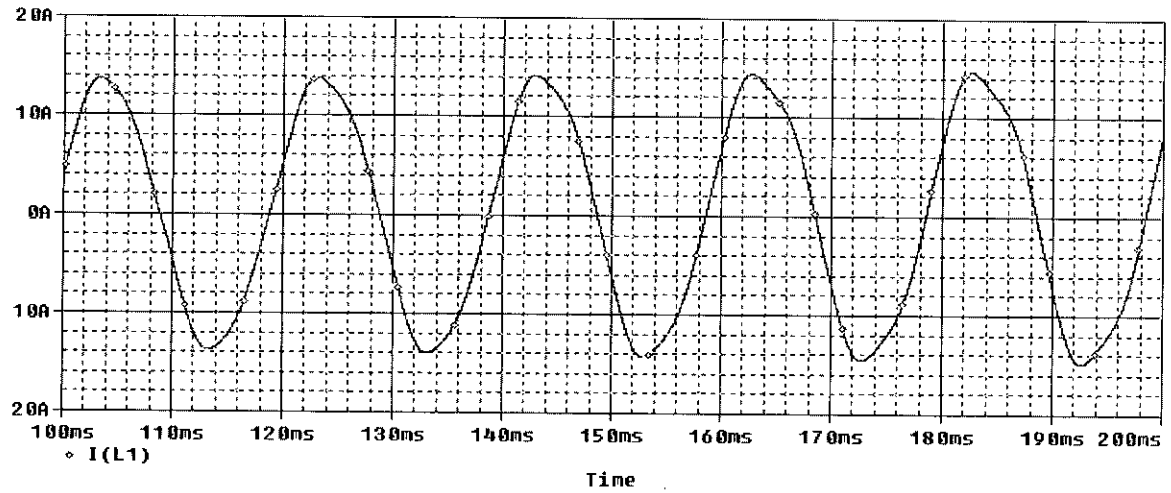


Figure 4.39: Input current for D=0.312

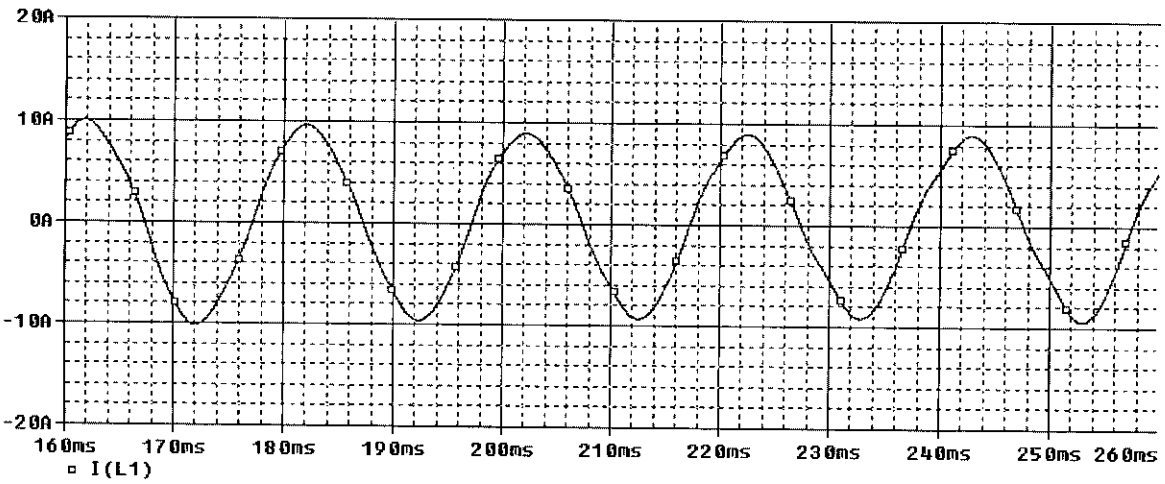


Figure 4.40: Input current for D=0.2

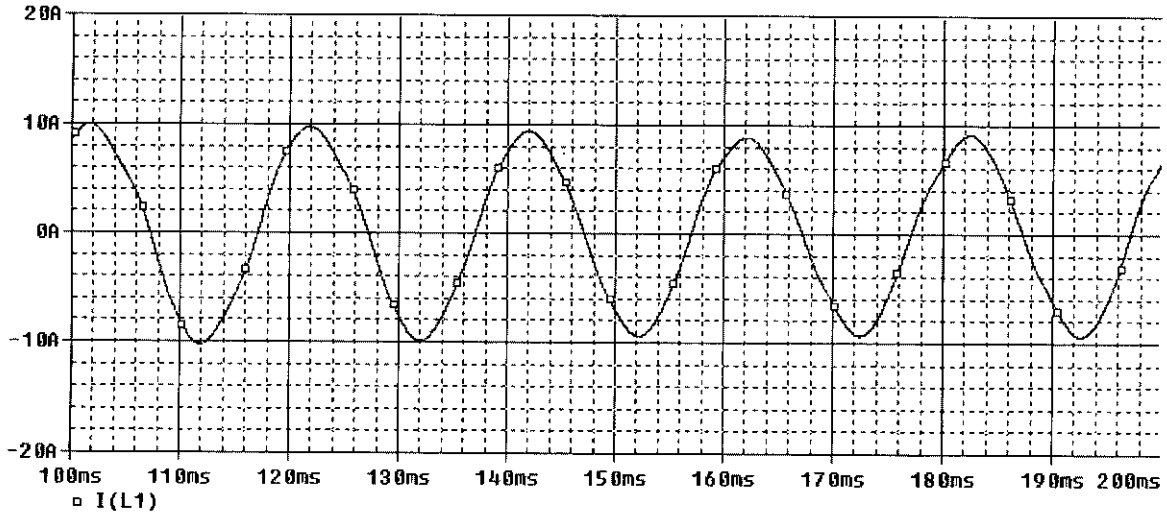


Figure 4.41: Input current for D=0.12

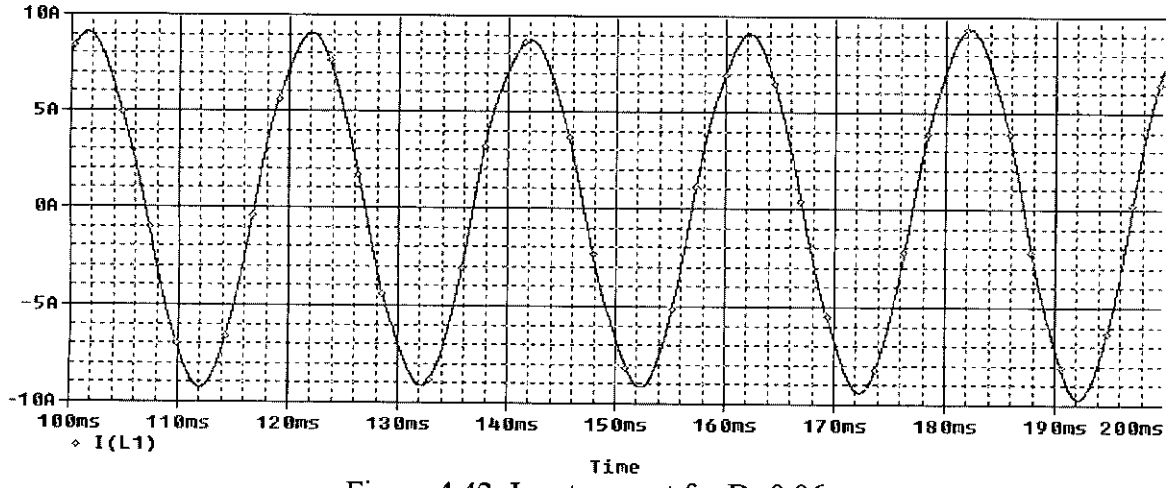


Figure 4.42: Input current for D=0.06

Here the input currents are more sinusoidal than that of the previous section that does not include the series resonating LC filter. The low order harmonic components can be drive out with the added resonating filter.

4.15 Wave shapes with load variation

Input and output wave shapes with 10ohm load connected to the output are also observed for different duty cycles. In the following Figures input current, output voltage and output currents for different duty cycle are shown.

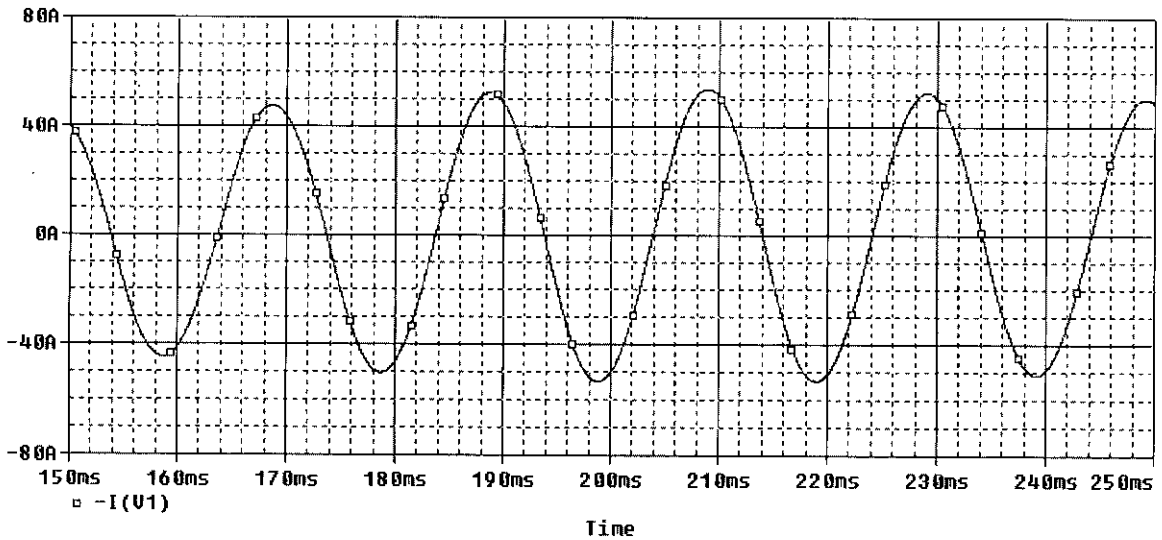


Figure 4.43: Input current with d=0.64

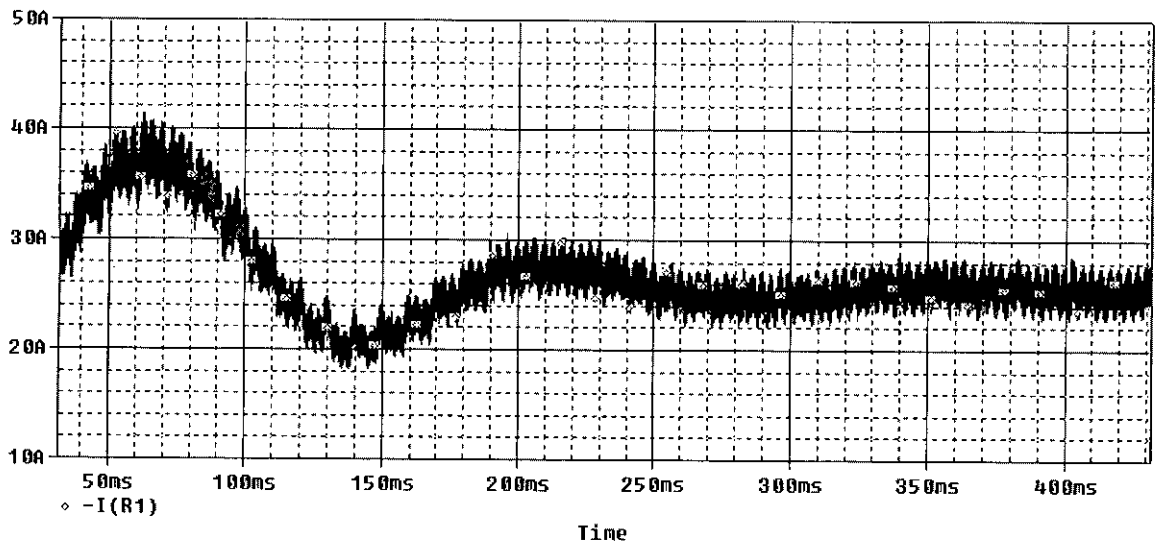


Figure 4.44: Output current with d=0.64

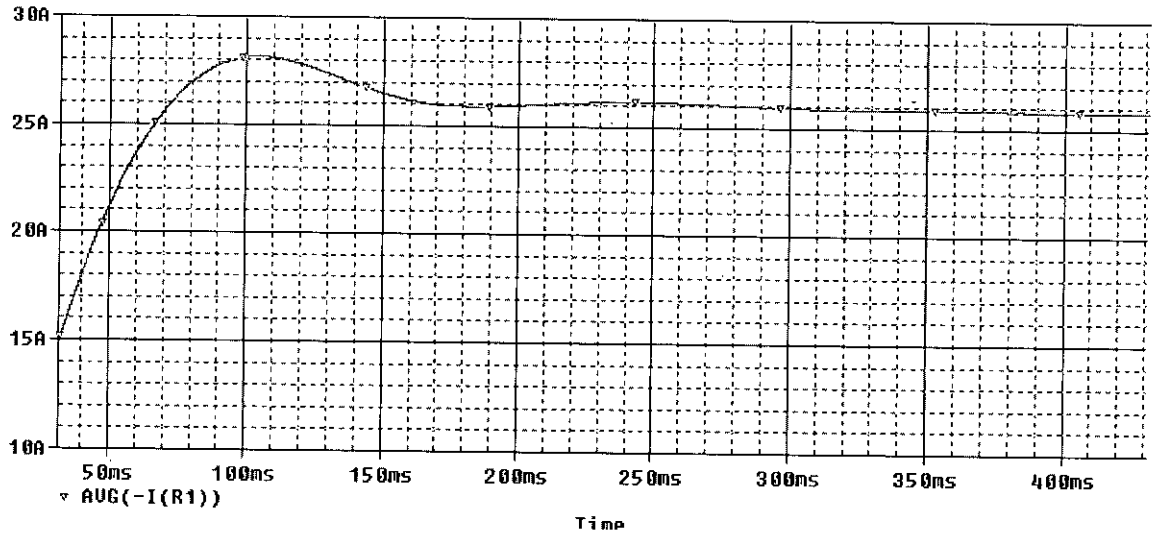


Figure 4.45: Average output current with d=0.64

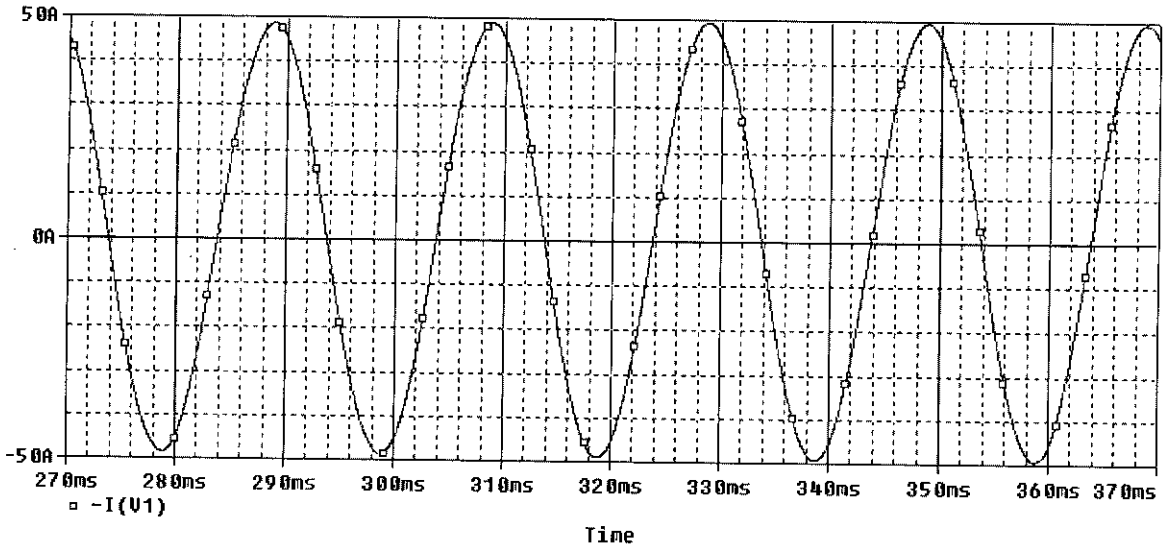


Figure 4.46: input current with d=0.548

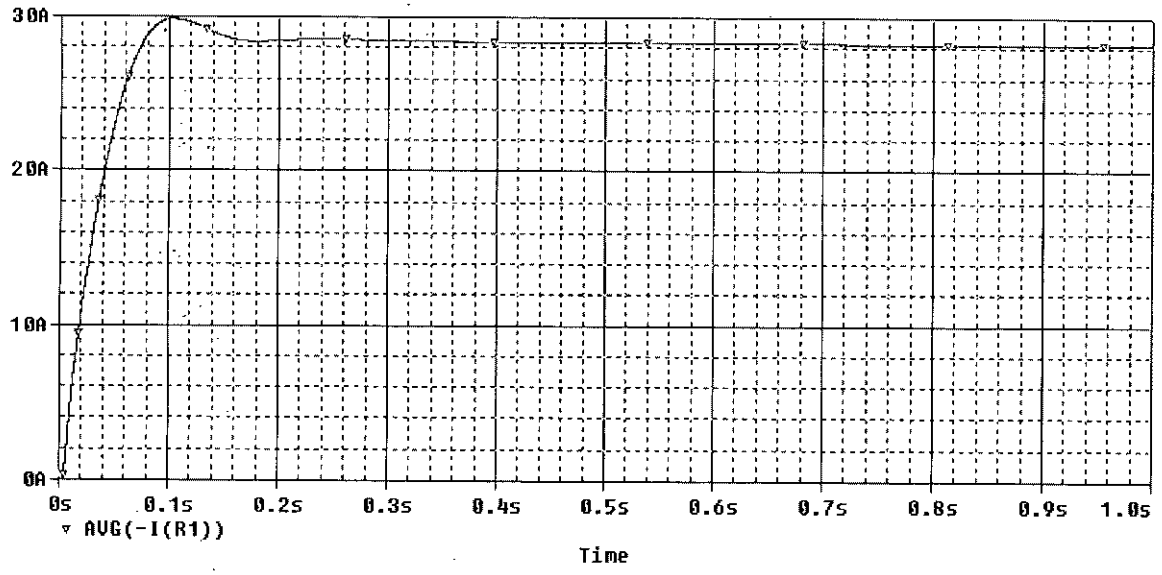


Figure 4.47: Average output current with d=0.548

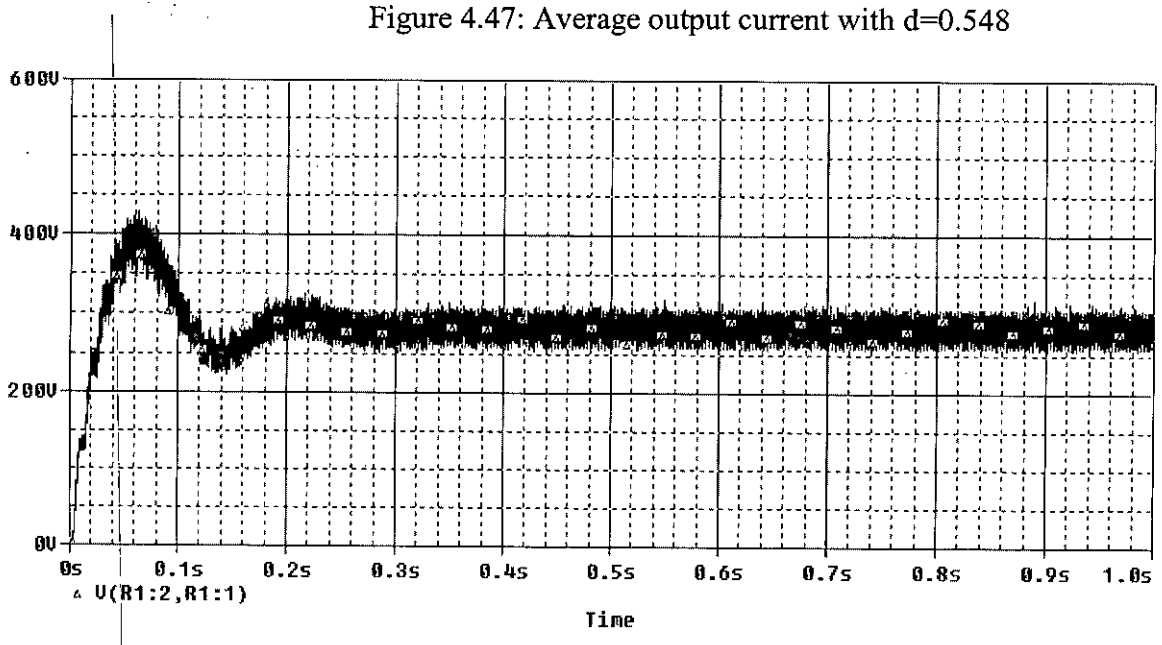


Figure 4.48: Output voltage with d=0.548

AS the load resistance is reduced the output voltage is also reduced. Because of the large output current drawn at the output the series resonating filter may come out of resonance and output reduces. But the input current wave shapes are still nearly sinusoidal.

Now Input and output wave shapes with 500ohm load connected to the output are also observed for different duty cycles. In the following figures input current, output voltage and output currents for different duty cycle is shown.

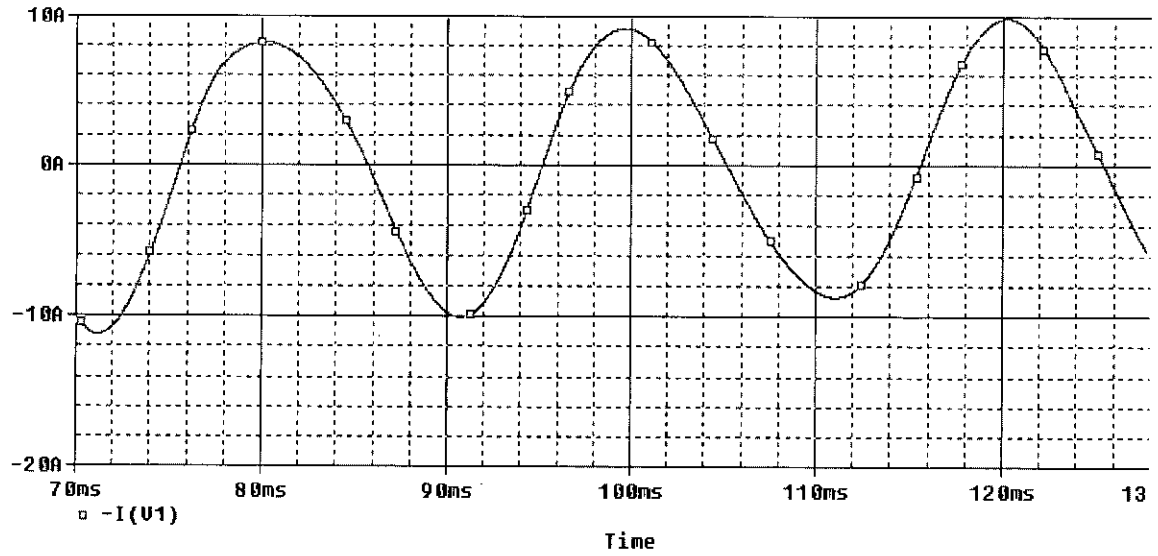


Figure 4.49: Input current with d=0.548

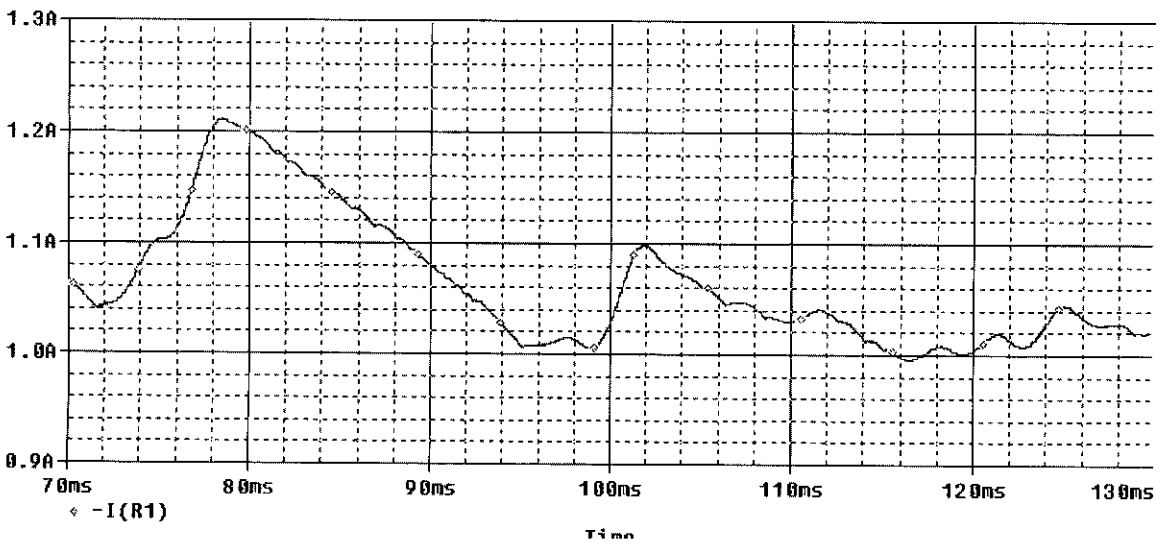


Figure 4.50: Output current with d=0.548

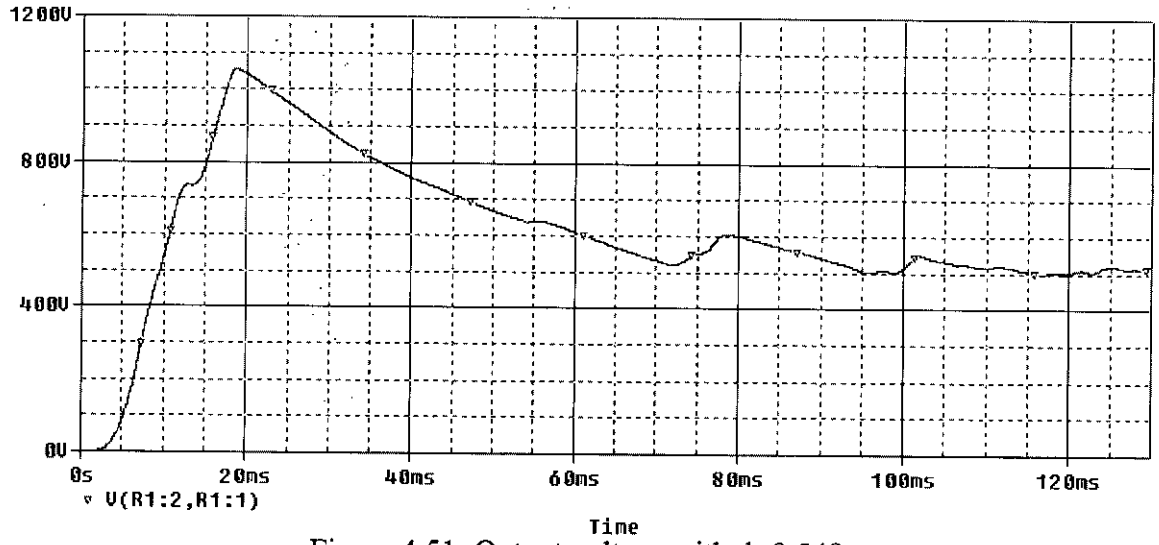


Figure 4.51: Output voltage with d=0.548

The output current is given below for RL load connected in the output.

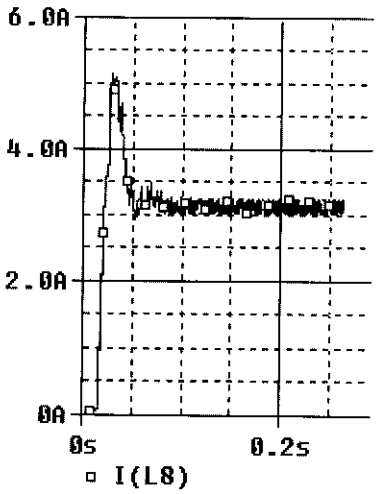


Figure 4.52: Output current with d=0.548

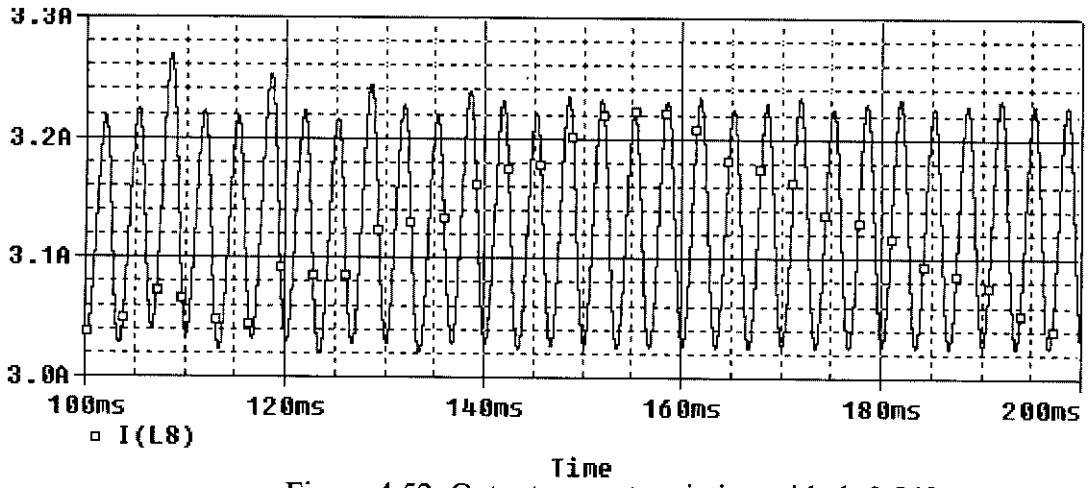


Figure 4.52: Output current variation with d=0.548

Inductive load connected at the output is actually reduces the ripple at the output current. In Figure 4.52 this output current variation is shown.

Chapter 5

Conclusions and Suggestions

5.1 Introduction

In a three phase rectifier circuit the input currents are non sinusoidal in nature and contains harmonics. The power infrastructure has to carry these currents and should be oversized to accommodate the flow of harmonics. Harmonics produce heat loss, lower efficiency, input ac mains voltage distortion, lower power conversion reliability, excitation of undesirable system resonances and increased VA rating of the equipments. That is why harmonic reduction is one of the main concerns. An easy technique to smooth out the input current is to use boost rectifier where the added inductor shape the input current from not being pulsating. But this leads to lessen the value of the output voltage. That is why active switching had been introduced. Different authors had already studied rectifiers with active switching and EMI filter. Even with the incorporation of EMI filter the THD value does not go below acceptable range. So in our work we have studied the effect of another series LC resonating filter to get an acceptable THD limit of the three phase rectifier input current.

5.2 Discussion of Results and achievements

First of all we had studied the normal rectifier circuit with and without capacitor. Without capacitor the THD value of a normal three-phase rectifier that we have designed is 22%. But when the capacitor was included at the output to filter the voltage ripple the THD value rise to 81%. Another drawback of this circuit is that it does not have voltage control option.

Then we studied the boost rectifier circuit and observed the improvement over normal rectifier. For boost action inductors are placed in the input side. The inductive behavior improves the input current shape from pulsating to be uniformly varying in nature. As we increase the value of the inductor the wave shape improves significantly but the output voltage decreases and in this circuit voltage regulation is also not possible.

Next boost rectifier with active switching has been studied. The switching circuit has PWM module for regulation of the output voltage. In this circuit the input current is DCM in nature and contains high frequency component. Though the average input current is sinusoidal in nature, this type of current would radiate interference in the nearby electronics equipments will hamper their normal operation.

Electro Magnetic Interference (EMI) filter is added in a boost rectifier with PWM module. Behavior of this module has been studied. It was found that the THD value has reduced to 17%. The input current wave shape is not purely sinusoidal. It contains low frequency components.

To suppress the low frequency component from the input current another series resonating filter has been introduced, which resonate at the supply frequency. With this added resonating filter the THD value has reduced to less than 4%. The wave shape is nearly pure sinusoid and voltage could be regulated from 400V to 1KV. But if efficiency is also the concern then the choice of a certain range of duty cycle to operate the module is necessary.

Our best operating point is at the range of duty cycle 0.47 to 0.64. At this range the power factor is unity and output voltage can be regulated from 750V to 900V and efficiency is 91% to 93% and the THD value is less than 2%.

5.3 Suggestion for Future work

In this work a resonant filter is used at the input side to filter low frequency components. The resonant filter resonates at line frequency (50Hz). As a result the filtering components are very large in values. Any other method like series parallel combination of LC filter to reduce the component values may be studied in future to see whether the component values can be lowered.

Only simulation is performed in this study. The work was not implemented practically. So another suggestion for future work of this thesis would be to implement the practical circuit. The practical circuit would require implementing two parts.

(1) Power circuit design
(2) Boost switching stage

Power circuit had been already designed in normal rectifiers. So the concern is to design the switching stage, which has four more stages to implement;

(a) Control circuit,
(b) Base drive,
(c) Base isolation and
(d) Voltage Controller.

The design of control circuit, base drive and base isolation stages are already given in this study. But the voltage control circuit has to be implemented as required by loads. Future work may include designing voltage control circuit.

Chapter 6

Appendix-A

6.1 A simple method of steady-state analysis of PWM converters

The steady-state analysis of PWM converters is the basis of the analysis and design. Here, a simple method [26] for steady-state analysis is introduced based on a DC/DC boost converter as shown in Fig. A-1. Generally, this method can be extended to other type converters, including DC/AC and AC/DC converters.

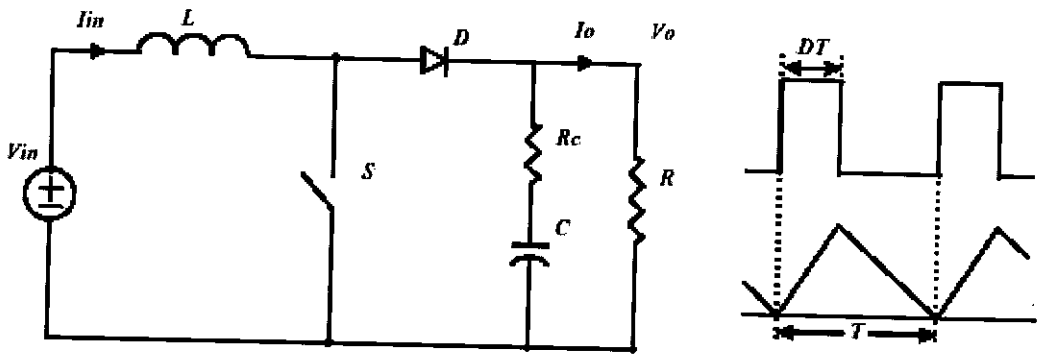


Fig. A-1. A DC/DC boost converter and its current waveform

At first, some nomenclatures are defined: the *critical power* P_c is load power when the converter operates in conduction boundary between CCM and DCM; the *delivered power* P is power delivered by the converter from source to load; and d is the duty cycle in both CCM and DCM. Therefore, the critical power can be derived according to its inductor current waveform. At the boundary operation, the average inductor current is given by

$$I_{in} = \frac{1}{2} i_p = \frac{1}{2} \frac{V_{in}}{L} DT \qquad \text{(A-1)}$$

where i_p is peak inductor current, D duty cycle in CCM, f switching frequency and $T=1/f$. Since it is still in CCM, the voltages and currents in input and output are related by

$$V_{in} = (1-D)V_o \quad \text{and} \quad Iin = \frac{1}{1-D} I_o \tag{A-2}$$

Substituting Eq. (A-2) into (A-1), the load current can be found. Thus, the critical power is obtained

$$P_c = V_o I_o = \frac{Vo^2}{2Lf} D(1-D)^2 \tag{A-3}$$

Comparing the delivered power $P= V_o^2/R$ to the critical power in Eq. (A-3), the steady-state can be obtained. On the one hand, if P is larger than P_c, the converter operates in CCM. The duty cycle is load independent and given by

$$d = D \tag{A-4}$$

On the other hand, if P is less than P_c, the converter operates in DCM, and the duty cycle is load-dependent. For constant output voltage, the voltage gain in DCM (referred to Table A-1) should be regulated to be equal to that in CCM, i.e.:

$$\frac{V_o}{V_{in}} = \frac{1+\sqrt{1+2d^2R/Lf}}{2} = \frac{1}{1-D} \tag{A-5}$$

Figuring out d and simplifying the expression by using P and P_c, the duty cycle can be represented as

$$d = \sqrt{\frac{P}{P_c}} D. \qquad \text{(A-6)}$$

So, from heavy load to no load, the steady-state is given by Eq. (A-4) or (A-6). Summarizing the above discuss yields a method for steady-state analysis of a DC/DC boost converter:

For a designed PWM converter, if the delivered power P is larger than P_c, then the converter operates in CCM and the duty cycle is given by Eq. (A-4). Otherwise, the converter operates in DCM and the duty cycle is given by Eq. (A-6).

It can be proved that the above method still holds for buck-type and flyback-type converters. The difference is that the critical power in Eq. (A-3) should be replaced by those listed in Table A-1.

In AC/DC PFC converters or DC/AC inverters, the above method still can be used, even though the duty cycle D, critical power, and delivered power are functions of the instant of the sinusoidal input voltages. This is because, at a fixed ωt, these variables can be considered as constant and the method can be applied just as on a DC/DC converter. The most important thing is to find out proper representations of these variables.

Table A-1. Critical power and voltage gains for general type PWM converters

	Boost-Type	Buck-Type	Flyback-Type
Critical Power	$\frac{V_o^2}{2Lf} D(1-D)^2$	$\frac{V_o^2}{2Lf}(1-D)$	$\frac{V_o^2}{2Lf}(1-D)^2$
Voltage Gain (CCM)	$\frac{1}{1-D}$	D	$\frac{D}{1-D}$
Voltage Gain (DCM)	$\frac{1+\sqrt{1+2d^2R/Lf}}{2}$	$\frac{2}{1+\sqrt{1+8Lf/d^2R}}$	$\sqrt{\frac{R}{2Lf}}\, d$

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