“IGBT BASED CURRENT CONTROLLED INVERTER”

Major Project Report

Submitted in Partial Fulfillment of the Requirements for the degree of

MASTER OF TECHNOLOGY

IN

ELECTRICAL ENGINEERING

(Power Apparatus & Systems)

By

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Warm regards,
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ABSTRACT

IGBT based current controlled inverter is used for High Power applications. It is based on Voltage source inverter (VSI) and the output is in terms of current. According to reference command output current will be changed. This project presents a study of 24 Pulse Rectifier, Buck Converter, H-bridge Inverter etc. Particular Power topology is the H-bridge inverter.

Due to advances of power electronics and inverter topology, current controlled inverter is usually preferred for quick response and accurate control. Current controller forces the load current to follow the current command in some apparatus.

This major project presents a simulation of 24-Pulse Rectifier also presents the simulation of Proposed Power topology H-bridge inverter. By using multi pulse we improve the line current distortion. Thus we improved the T.H.D. This major project also presents the designing of heat sink & Power loss calculation & Driver card testing for H-bridge Inverter etc. The experimental results are obtained and presented in this thesis.

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NOMENCLATURE

V_s : Supply voltage

I_s : Supply current

V_{dc} : Input DC voltage

V_o : Output voltage

I_o : Load current

P : Total power loss

R : Load resistance

L : Switching Inductor

C : Filter capacitor

f : Switching frequency

T : Time period

D : Duty Cycle

ΔI : Current ripple

V_L : Voltage across inductor

V_{ripple} : Voltage ripple

I_L : Inductor current

I_d : Diode current

P_{cond} : Conduction loss

$P_{s/w}$: Switching loss

P_{total} : Total power loss

E_{on} : Turn on energy loss

E_{off} : Turn off energy loss

T_j : Junction temperature

$R_{th(j-c)}$: Junction to case thermal resistance

$R_{th(c-f)}$: Case to fin thermal resistance

$R_{th(f-a)}$: Fin to ambient thermal resistance

CONTENTS

CHAPTER - 1

INTRODUCTION

1.1 General

IGBT based current controlled inverter is used for High Power applications. Due to advances of power electronics and inverter topology, current controlled inverter is usually preferred for quick response and accurate control. Current controller forces the load current to follow the current command in some apparatus.

It requires a constant 400V DC high power output with a 3-phase 415 V AC input to supply power to a varying load.

This major project presents the simulation of power topologies, designing of heat sink & Power loss calculation, manufacturing, testing etc. Particular Power topology used is H-bridge Inverter etc.

In this chapter, scope of work, basic block diagram, literature survey and organization of thesis are discussed. IGBT module consists of IGBT chips and FWD chips. Power loss can be classified as either Conduction losses or switching losses.

Heat Sink Selection depends on many factors:

- Performance
- Dimensional Constraints
- Cost etc.

1.2 Scope of Work

In this Project to design and implement an Ac to dc converter with a regulated output voltage. The objectives are to design a converter with the following requirements:

- High power output
- 400 V DC output with varying 415 V AC input
- Remain within the allocated budget.

The decided input voltage will vary between 415V ± 10 % V and the permissible output voltage will be ± 5 percent of the stated regulated voltage of 400V DC. The ratings of all the devices and all the components of the circuit will be designed

1.3 Basic Block Diagram

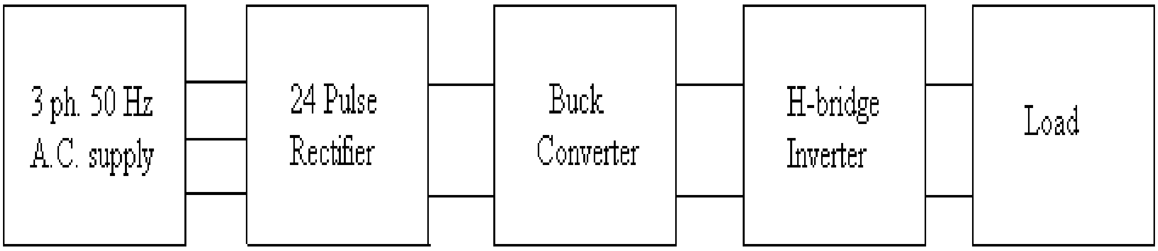


Fig. 1.1 Basic block diagram

A Rectifier transformer shall transfer the incoming 3 phase, 415 V AC power. The transformer secondary is connected to a DC converter module, the output of which should be connected to a properly designed filter bank.

A DC converter shall convert the incoming AC power to controlled DC power for charging, which shall form the dc source for a four quadrant transistorized 'DC to AC/DC' current regulated inverter.

1.4 Literature Survey

Literature survey plays a very important role in the project Literature survey consists of current controlled inverter related papers that includes different power topologies, control schemes, simulation and experiments. Papers were taken from IEEE conference proceedings, journal proceedings and other standard publications.

The paper titled, "Proposal of an Isolated Current Controlled Inverter" [5] A combination of a two transistor forward converter (input) and a conventional full-bridge inverter (output) with current control loop is proposed, resulting in an isolated inverter. The forward converter operates in high frequency switching through a fixed frequency hysteresis controller, making the load current modulation.

The paper titled "Boost-Buck Push-Pull Converter for Very Wide Input Range Single Stage Power Conversion" [1] presents that Push pull converter that has both buck and boost regulator characteristics. At low input voltage the push-pull converter acts as a boost, and at high input voltages it acts as a buck converter. In Boost mode the duty-cycle of both switches is increased to greater than 50%.

In this paper titled, “A Novel Winding-Coupled Buck Converter for High-Frequency, High-Step-Down DC–DC Conversion” [3] analyzes the fundamental limitations of the buck converter for high-frequency, high-step-down dc–dc conversion. Further modification with additional coupled windings in the buck converter yields a novel topology, which significantly improves the efficiency without compromising the transient response.

In this paper titled, “Load current estimation for control algorithms in buck converter” [4] a switching load current estimator for a buck converter is introduced. By combining two estimation principles in a single estimator, both a fast response ad a high accuracy of the load current estimation can be obtained. The controller of the buck converter then uses this estimated load current, enabling it to respond fast on load changes and regulate the output voltage with high accuracy.

Here, for Power loss calculation & Heat Sink design Fuji IGBT catalogue used & Mitsubishi catalogue referred.

1.5 Organization of the thesis

The thesis has been organized in ten chapters. The topics to be covered in each chapter are briefly explained below.

Chapter-I

Introduction to the topic of work, Scope of the work, literature survey and Basic block diagram of the project and thesis construct are presented in Chapter-I of the thesis.

Chapter-II

In this chapter brief description of 24 pulse rectifier, study of buck converter and Power topology operation are presented.

Chapter-III

In this chapter buck converter design, Power loss calculation and Heat sink design for full scale version for IGBT based chopper has been presented.

Chapter-IV

In this chapter Simulation of the 24 pulse rectifier, simulation of buck converter and simulation of power circuit for full scaled version and down scale version has been presented and simulation results discussed for both full scale version and down scale version.

Chapter-V

In this chapter Power loss calculation and heat sink design done for scaled down version for different IGBT and then selected IGBT for scaled down version.

Chapter-VI

In this chapter Assembly of 24 pulse rectifier has been presented and prototype hardware model done for scaled down version & experimental results discussed.

Chapter-VII

In this chapter Testing of chopper controller has been presented. Test circuit has been discussed.

Chapter-VIII

In this chapter prototype assembly for H-bridge has been done and driver card testing for H-bridge inverter discussed and experimental results shown.

Chapter-IX

In this chapter Experimental result for prototype hardware has been discussed for scaled down version.

Chapter-X

In this chapter major Conclusions of this work and some possibilities for future work research work are discussed. A list of referred papers is given under References. APPENDIX shows the Experimental set up and overview of chopper controller IC TL594.

CHAPTER – 2

STUDY OF 24-PULSE RECTIFIER & BUCK CONVERTER & POWER TOPOLOGY

2.1 General

This chapter presents 24 pulse Rectifier operation, Study of Buck converter and power topology. Particular power topology used is H-bridge Inverter.

2.2 24 Pulse Rectifier Operation

The rectifiers can be configured as 12-, 18- and 24-pulse rectifiers, powered by a phase shifting transformer with a number of secondary windings. Here, 24 Pulse Rectifier is discussed is shown in fig.2.1, where ia, ib, ic,id are the primary currents referred from the secondary side of the transformer. Each of these currents has a THD of 24%. The primary line current *iA* is virtually a sinusoid with only 1.49% total harmonic distortion. Each secondary winding feeds a six-pulse diode rectifier. Here, phase-shifting transformer is used to power four sets of six-pulse diode rectifiers.

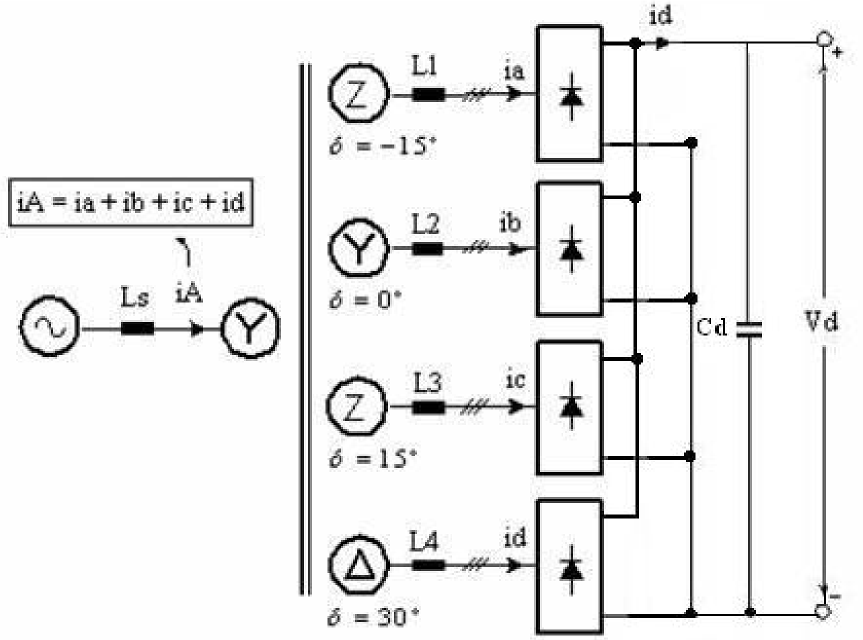


Fig.2.1 24 Pulse diode Rectifier

The main features of the multipulse (24-Pulse) diode rectifiers are:

- To reduce the line current harmonic distortion.
- does not require any LC filters or power factor compensators
- To improve the input power factor

The line-to-line voltage of each secondary winding is usually one fourth that of the primary winding. This is achieved by the phase shifting transformer, through which some of the low-order harmonic currents generated by the six-pulse rectifiers are canceled.

In general, the higher the number of rectifier pulses, the lower the line current distortion.

Table 2.1 Theoretical amplitudes of various harmonics

Harmonic	6-pulse	12-pulse	18-pulse	24-pulse
5	.200			
7	.143			
11	.091	.091		
13	.077	.077		
17	.059		.059	
19	.053		.053	
23	.043	.043		.043
25	.040	.040		.040

Conclusion:

From the above table, it is apparent that higher pulse numbers directly translates in to elimination of lower order harmonics.

2.3 Buck Converter Power Circuit

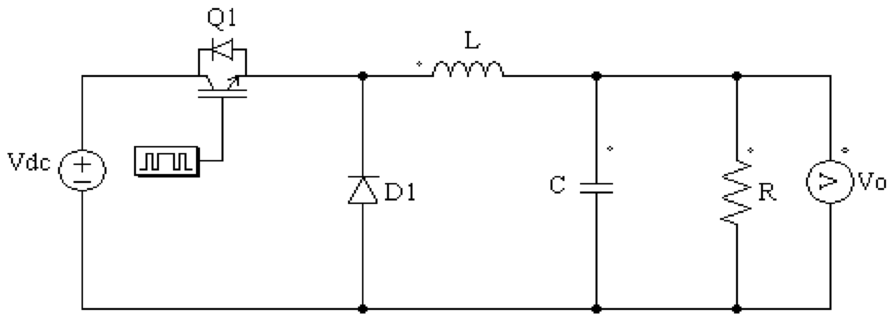


Fig. 2.2 Buck Converter Circuit

The Power circuit of Buck Converter is shown in fig. 2.2. In a buck converter the output voltage is less than the input voltage. As can be seen from the circuit there is one switch is in the series with the Dc input Vdc. The LC filter is added in series.

The features of Buck Converter are:

1) It has low internal losses
2) High Frequency
3) It requires smoothing input filter

2.4 Power Topology

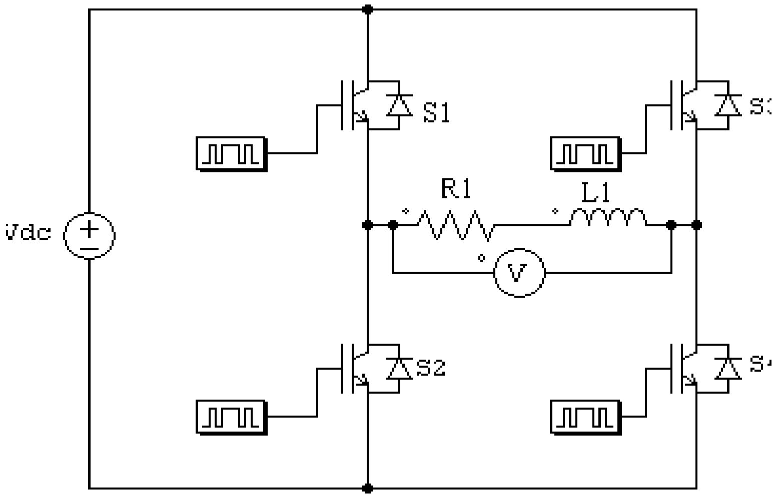


Fig. 2.3 H-bridge Inverter Power Circuit

2.4.1 Circuit Operations

With 4 switches there are 6 combinations to be examined.

- **S1-S2 ON or S3-S4 ON**

  Both create short circuits across the DC source and are invalid.

- **S1-S4 ON**

  Applies Vdc to the load. With iL positive current passes through S1-S4, for iL negative the current is through D1-D4.

- **S2-S3 ON**

  Applies - Vdc across the load. With iL positive the current flows through D2-D3 and returns energy to the DC source. With iL negative the current flows through S2-S3 and draws energy from the supply.

- **S1-S3 ON**

  Applies 0 volts across the load. For iL positive the path is S1-S3 for iL negative the path is D1-S3.

- **S2-S4 ON**

  Applies 0 volts across the load. For iL positive the path is through D2-S4. For iL negative the path is S2-D4.

CHAPTER – 3

DEVICE SELECTION AND HEAT SINK DESIGN FOR IGBT BASED CHOPPER

3.1 General

This Chapter presents Buck Converter Design, Comparison of Buck Converter Design Parameters for different frequencies at constant input and output voltage and Power Dissipation Loss Calculation for device selection & Heat Sink Design for IGBT Based Chopper.

3.2 Buck Converter Design

Input Data:

$V_s = 560$ V

$V_o = 400$ V

f = 5 KHz and

$I_o = 650$ A

$$\begin{aligned}\delta &= V_o/V_{dc} \\ &= 400/560 \\ &= 0.714 \end{aligned} \tag{3.1}$$

$$\begin{aligned}T &= 1/f \\ &= 1/5000 \\ &= 0.2 \text{ msec.} \end{aligned} \tag{3.2}$$

$$\begin{aligned}T_{on} &= \delta T \\ &= 0.714*(0.2*10^{-3}) \\ &= 0.143 \text{ msec} \end{aligned} \tag{3.3}$$

Toff = T- Ton

= 0.2-0.143

= 0.057 msec (3.4)

Inductor Calculation

Current Ripple: ΔI = 25 % of Load Current

= 0.25 * 650

= 162.5 A (3.5)

V_L = Vs – Vo

= 560 – 400

= 160 V (3.6)

L = (V_L * Ton) / ΔI

= (160 * 0.143 * 10^-3) / (162.5)

= 140.8 μH (3.7)

Capacitor Calculation

Vripple = 5 % of output Voltage

= 0.05 * 400

= 20 V (3.8)

C = (Io* Ton)/(Vripple)

= (650 * 0.143)/(20)

= 4647.5 μF

Diode Current:

I_D = (Io * Toff)/ T

= (650 * 0.057)/(0.2)

= 185.25 A (3.9)

I_T = (Io * Ton) / T

= (650 * 0.143) / (0.2)

= 465 A (3.10)

I_L = Io + ΔI

= 650 + 162.5

= 812.5 A (3.11)

Ipeak = Io + (ΔI/2)

= 731.25 A (3.12)

Table 3.1: Comparison of Buck Converter Design Parameters

V_O (V)	V_{dc} (V)	f (KHz)	I_O (A)	L (µH)	C (µF)	I_D (A)	I_T (A)	I_L (A)	I_{peak} (A)
400	560	5	100	915	715	28.5	71.5	125	112.5
400	560	10	100	457	357	28.6	71.4	125	112.5
400	560	5	650	140.8	4647.5	185.25	465	812.5	731.25
400	560	10	650	70.3	2320.5	185.9	464.1	812.5	731.25

Conclusion:

For constant Input and output voltage if we increase the value of load current then peak current will be increased. If we increase the frequency for same load current then values of inductor & capacitor decreased & Inductor current will be same.

3.3 Power Dissipation Loss Calculation

3.3.1 Type of Power Loss

An IGBT module consists of IGBT chips and FWD chips. Power loss can be classified as either Conduction losses or switching losses.

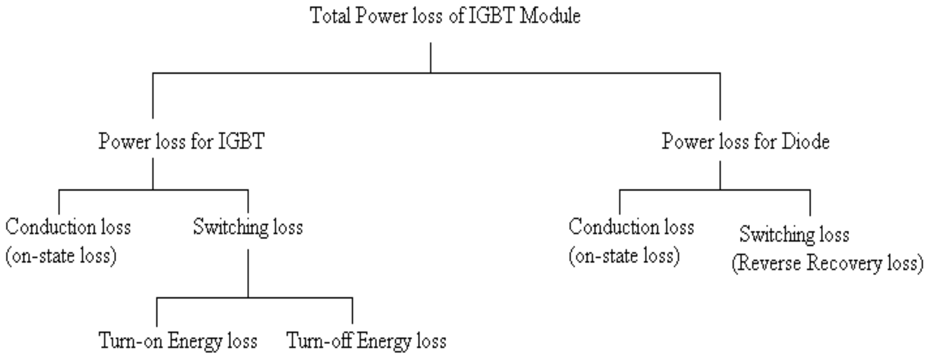


Fig. 3.1 Estimation of Total Power Loss

3.3.2 Definition of Conduction Loss & Switching Loss

- **Conduction Loss:**

Conduction losses are the losses that occur while the IGBT is on and conducting current.

- **Switching Loss**

Switching loss is the power dissipated during the turn-on and turn-off switching transitions.

3.3.3 Power loss calculation for IGBT

IGBT Power dissipation loss = Conduction loss + Switching loss

= On-state loss + Turn-on loss + Turn-off loss

Vdc = 560 V

Vo = 400 V

Ic = 325 A

f = 5 KHz

From datasheet of BSM300GB120DLC,

When I_c = 325 A then $V_{CE(sat)}$ = 2.5 V, E_{on} = 38 mJ, E_{off} = 36 mJ, E_{rec} = 20 mJ and V_F = 1.8 V

- Conduction-loss for IGBT

$$\begin{aligned} Pcond &= [V_{CE(sat)} * Ic * \delta] \\ &= (2.5 * 325 * 0.714) \\ &= 580.12\ W \qquad (3.13) \end{aligned}$$

- Switching loss for IGBT

$$\begin{aligned} Ps/w &= [f * (E_{on} + E_{off})] \\ &= [5000 * ((38 + 36) * 10\text{^}{-3})] \\ &= 370\ W \qquad (3.14) \end{aligned}$$

So, IGBT Power dissipation loss = Conduction loss + Switching loss

$$\begin{aligned} &= 580.12 + 370 \\ &= 950.12\ W \qquad (3.15) \end{aligned}$$

3.3.4 Power loss calculation for Free-wheeling Diode

- Conduction-loss for Free-wheeling diode

$$\begin{aligned} Pcond &= (VF * IF * Toff)/ T \\ &= (1.8 * 325 * 0.057) /(0.2) \\ &= 166.72\ W \qquad (3.16) \end{aligned}$$

- Switching loss for Free-wheeling diode

Ps/w = (Erec * f)

= (20 * 10^-3 * 5 * 10^3)

= 100 W (3.17)

So, Free-wheeling diode Power loss dissipation = Conduction loss + Switching loss

= 166.72 + 100

= 266.72 (3.18)

3.3.5 Total Power loss of IGBT Module

Total Power loss of IGBT Module = Power loss for IGBT + Power loss for Forward diode

= 950.12 + 266.72

= 1216.84 W (3.19)

Table 3.2 Comparison of Power Loss Calculation for IGBT & Diode

Ic (A)	**f KHz**	**$V_{CE(sat)}$ V**	**Eon (mJ)**	**Eoff (mJ)**	**Erec (mJ)**	**V_F (V)**	**Power Loss for IGBT (W)**		**Ptotal for IGBT (W)**	**Power Loss for Diode (W)**		**Ptotal for Diode (W)**	**Ptotal of IGBT Module**
							Pcond	**Ps/w**		**Pcond**	**Ps/w**		
40	5	1.2	8	9	5	1.1	34.27	85	119.27	12.58	50	62.58	181.85
40	10	1.2	8	9	5	1.1	34.27	170	204.27	12.58	25	37.54	241.83
50	5	1.25	8	11	6	1.2	44.62	95	139.62	17.16	30	47.10	186.72
50	10	1.25	8	11	6	1.2	44.62	190	234.62	17.16	60	77.16	311.78
325	5	2.5	38	36	20	1.8	580.12	370	950.12	166.72	100	266.72	1216.84
325	10	2.5	38	36	20	1.8	580.12	740	1320.1	166.72	200	367.32	1687.42

Conclusion:

From above table we conclude that if we increase the frequency for same load current then there is no change in Conduction loss of IGBT and Forward diode and there are switching loss changes.

If we increase the load current for different frequencies then total loss of IGBT module will be changed.

3.4 Heat Sink Design for IGBT Based Chopper

Input Data

$V_{dc} = 560$ V

$V_o = 400$ V

I = 100 A

f = 5 KHz

- From Table 3.2 Ptotal of IGBT Module = 186.72 W
- From the BSM300GB120DLC IGBT datasheet

  Rth(j-c) = 0.050 K / W

  Rth(c-f) = 0.010 K / W

Assume Tj(max) = 115 ºC

Ta(max) = 50 ºC

- The junction temperature can be calculated using the following thermal equation:

  Tj = Ptotal (IGBT Module) * {Rth(j-c) + Rth(c-f) + Rth(f-a)} + Ta

Therefore,

- Thermal Resistance between heat sink and ambient temperature

  Rth(f-a) = {[(115 – 50) / 186.72] –[(0.050+ 0.010)]}

  = 0.288 K / W (3.20)

- Junction temperature for IGBT

  Tj(T) = [WT * Rth(j-c)] + [(WT + WD) * Rth(c-f)] + [(WT + WD) * Rth(f-a)] + Ta

  = [139.62 * 0.05] + [(139.62 + 47.1) * (0.01)] + [(139.62 + 47.1) * 0.288] + 50

  = 112.62 ºC (3.21)

- Junction temperature for Forward diode

Tj(D) = [WD * Rth(j-c)] + [(WT + WD) * Rth(c-f)] + [(WT + WD) * Rth(f-a)] + Ta

= [47.10 * 0.125] + [(139.62 + 47.1) * (0.01)] + [(139.62 + 47.1) * 0.288] + 50

= 111.5 ºC (3.22)

Conclusion:

From the above heat sink design for IGBT based chopper we can conclude that to select a heat sink that can keep the junction Tj below the Tj(max).

CHAPTER-4

SIMULATION AND RESULTS OF 24 PULSE RECTIFIER & BUCK CONVERTER & H-BRIDGE INVERTER

4.1 General

This chapter presents simulation of 24 Pulse Rectifier & Buck Converter. The simulation has been performed with the help of PSIM software. PSIM is a simulation package specifically designed for power electronics and motor control. With fast simulation, friendly user interface and waveform processing, PSIM provides a powerful simulation environment for power converter analysis, control loop design, and motor drive system studies.

4.2 Simulation of 24 Pulse Rectifier

In this simulation of 24-Pulse rectifier has been proposed. The simulation was carried out to verify the output voltage & line current of 24 Pulse Rectifier. Also T.H.D. will be reduced because of multi pulse rectifier.

4.2.1 Full scale version

For the various conditions and taking the calculated values of different component has been used for the simulation of full scale version. The schematic of simulation for 24 pulse Rectifier is shown in fig. 4.1

The mains line-line voltages are taken as 415V rms with a supply frequency of 50Hz. The inductance of 24-Pulse rectifier is taken as 0.5mH and capacitance is 4.7uF. The output of 24-Pulse rectifier is given to buck converter as a input. This input voltage is 560 V. By Buck converter design calculation, the inductance of buck converter is 140.8mH and capacitance is 4700uF. The switching frequency is chosen for buck converter design is 5 kHz

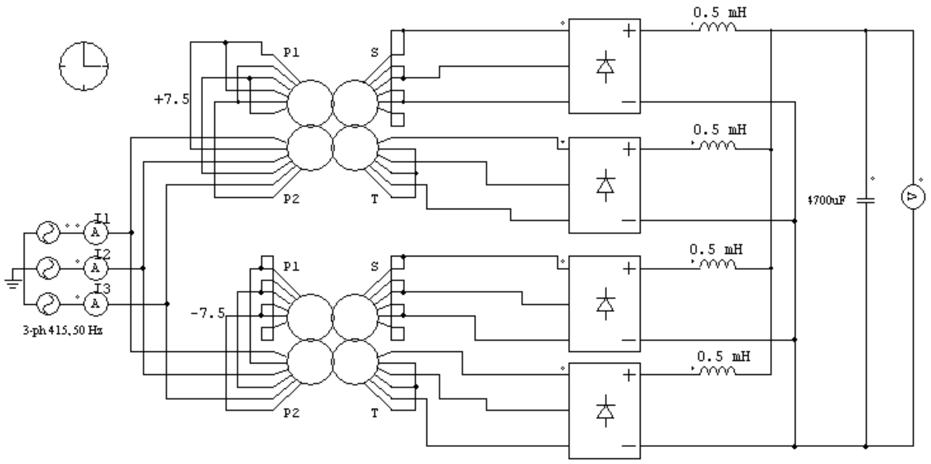


Fig. 4.1 Simulation Schematic of 24 Pulse Rectifier

4.2.2 Simulation Result

Fig. 4.2 Line current of 24-Pulse Rectifier & Fig. 4.3 Output wave-form of 24-Pulse Rectifier, Vo = 560 V.

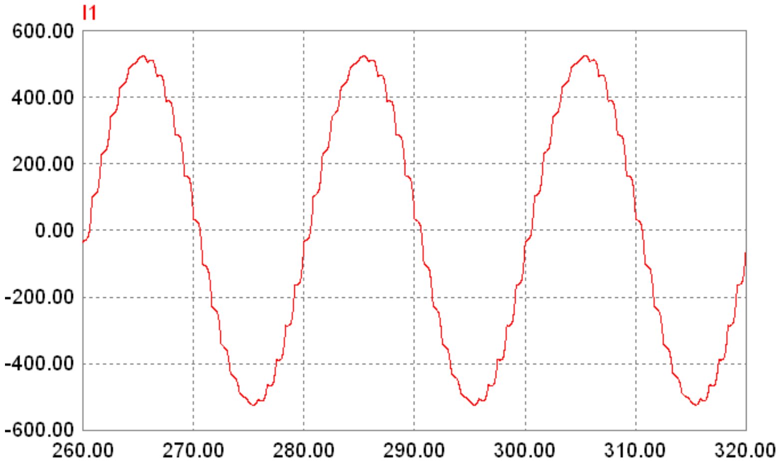


Fig. 4.2 Line current of 24-Pulse Rectifier

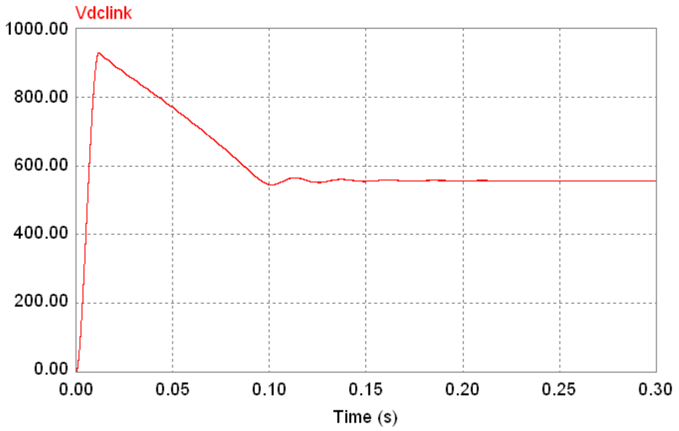


Fig. 4.3 Output wave-form of 24-Pulse Rectifier, Vo = 560 V

4.3 Simulation of 24 Pulse Rectifier & Buck Converter

Fig. 4.4 shows the schematic of 24 pulse rectifier & Buck converter in a closed loop.

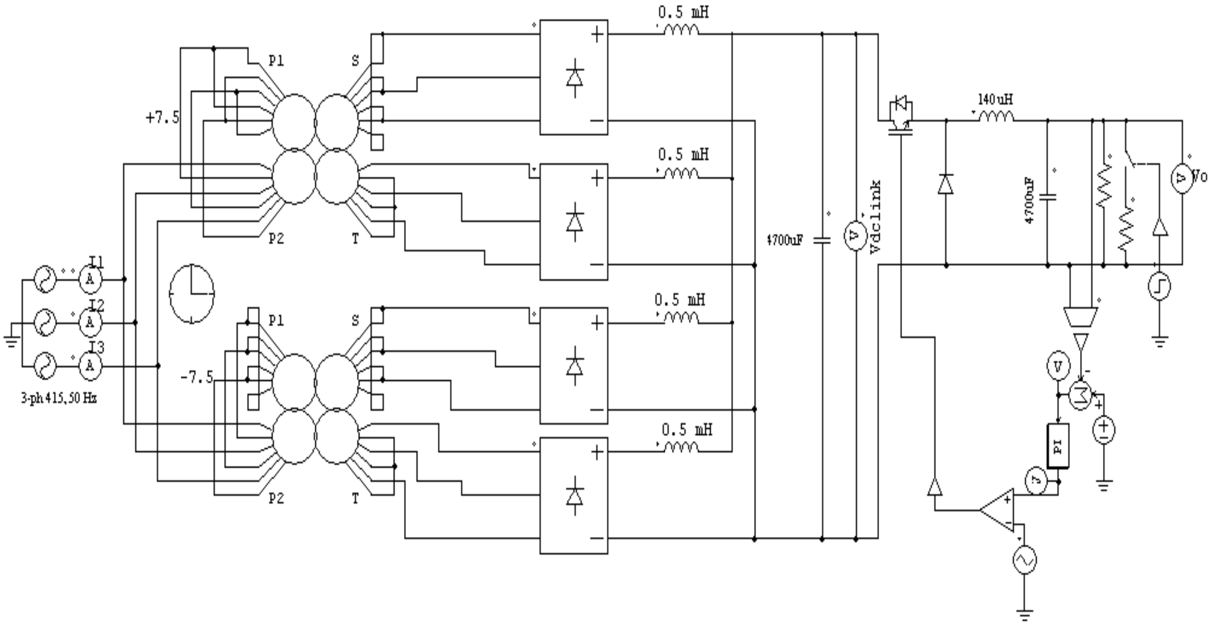


Fig.4.4 Output of 24 pulse rectifier as a input of buck converter

4.3.1 Simulation Results

Fig. 4.5 Output wave-form of 24-Pulse Rectifier as a Input voltage of Buck converter. Here, we get 560 V. Fig. 4.6 shows the Output voltage of Buck-Converter when load changed after 1 sec. Here, we get constant dc voltage as our requirements. Fig. 4.7 shows the Load current of Inverter. Here, we get 650 A approximately as shown in fig. Fig. 4.8 Wave-forms of Buck converter switching regulator.

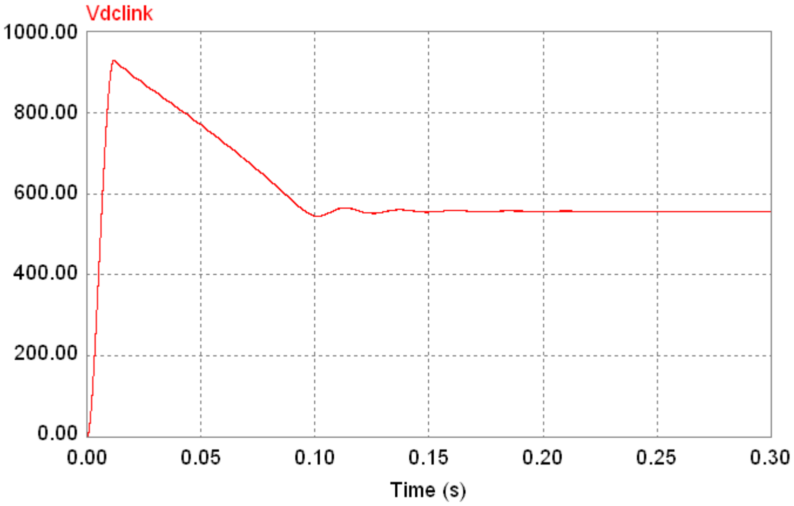


Fig. 4.5 Output wave-form of 24-Pulse Rectifier as a Input voltage of Buck converter

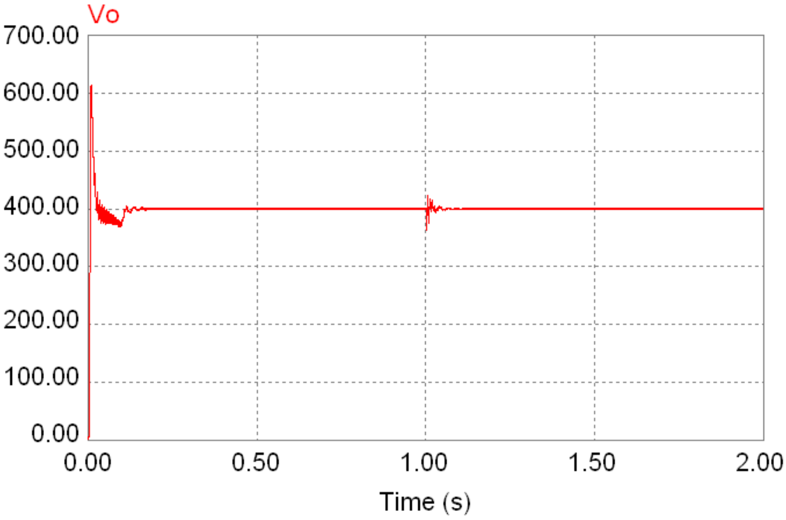


Fig. 4.6 shows the Output voltage of Buck-Converter when load changed after 1 sec

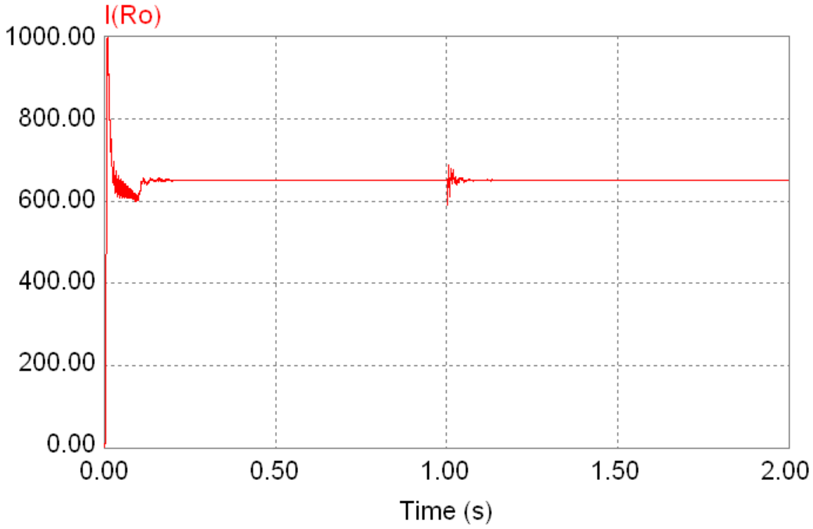


Fig. 4.7 shows the Load current of Inverter

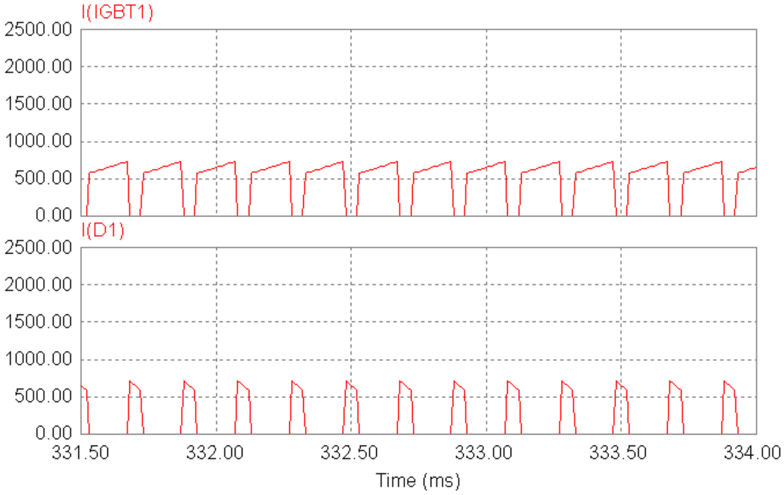


Fig. 4.8 Wave-forms of Buck converter switching regulator.

4.4 Simulation of Power Circuit

Fig. 4.9 shows the Simulation Schematic of whole power circuit for 650 A load current.

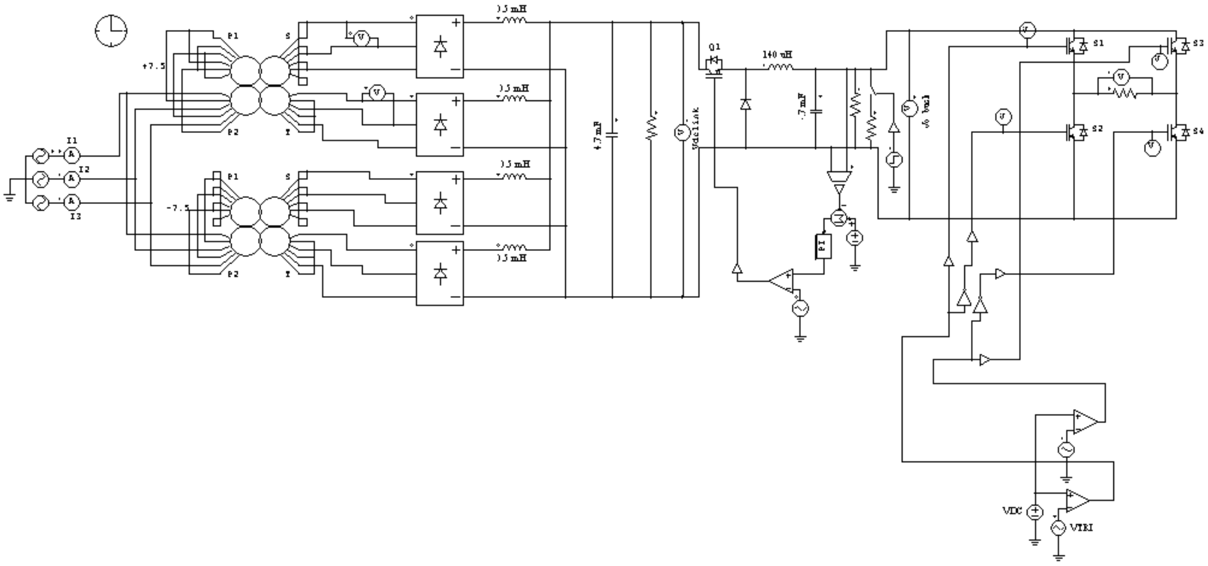


Fig. 4.9 Simulation Schematic of Power circuit

4.4.1 Simulation Results

Fig. 4.10 shows the Line current I_1, I_2 & I_3 wave-form with phase shifted by 15°. Here, we get 24 Pulse Operation.

Fig. 4.11 shows the output voltage of 24 pulse Rectifier. Here, we get 560 V approximately.

Fig. 4.12 shows the output voltage wave-form at buck converter side when load changed after 1 sec. Here, we are getting output voltage 400 V Dc with varying load as our requirements. Fig. 4.13 shows the switching wave-form of H-bridge Inverter.

Fig. 4.14 shows the output voltage of H-bridge Inverter. Here, we are getting 400 V A.C. as our requirements.

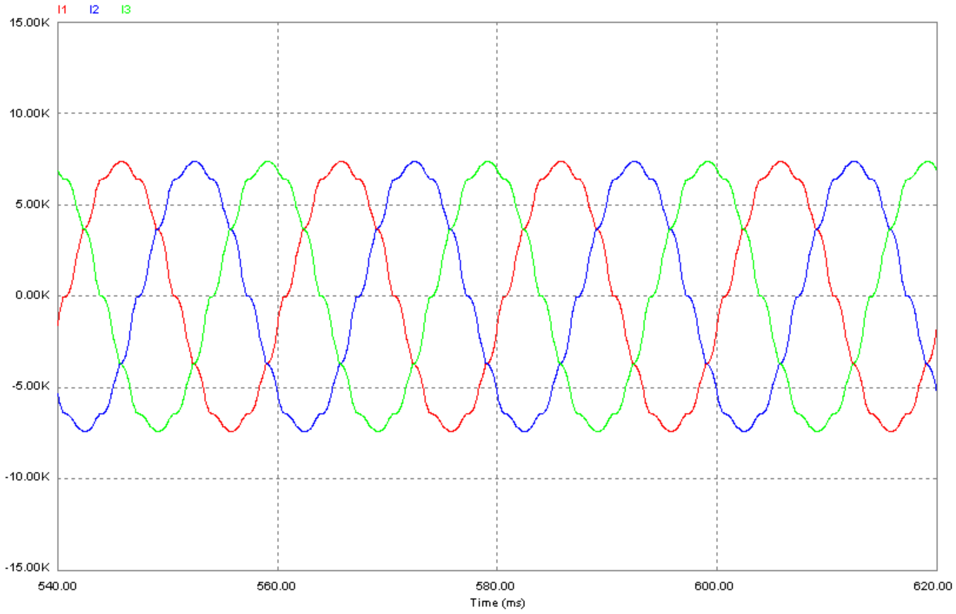


Fig. 4.10 Line current wave-forms with phase shifted by 15˚

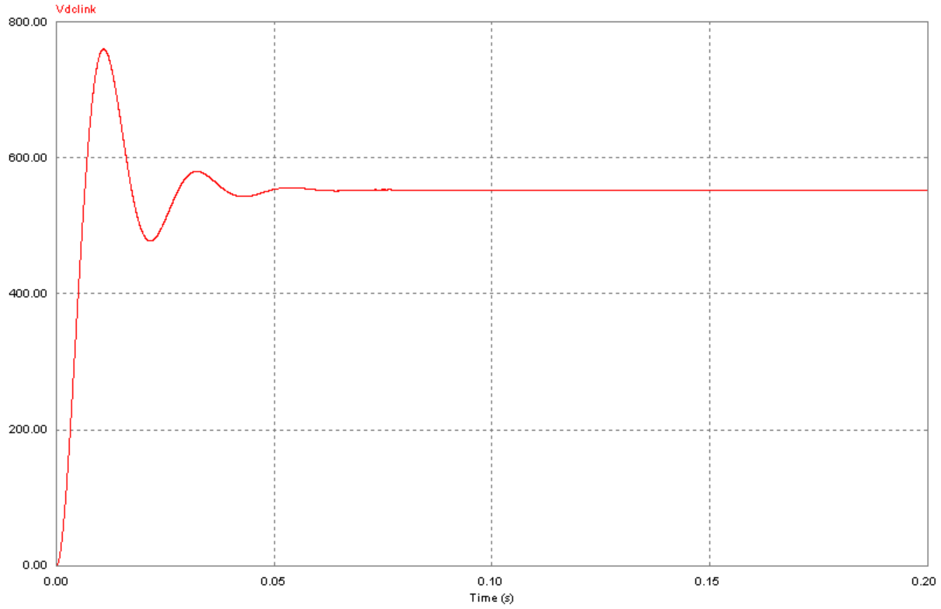


Fig. 4.11 Output voltage wave-form of 24 pulse rectifier

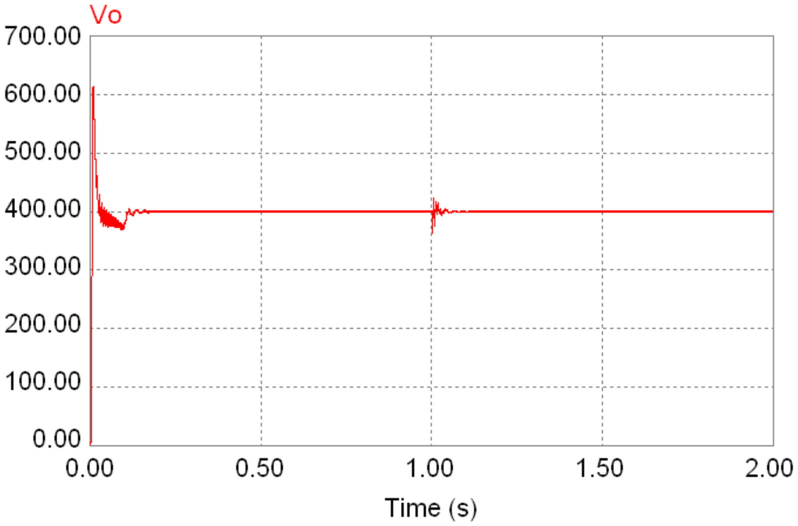


Fig. 4.12 Output voltage wave-form with varying load changed after 1 sec, Vo = 400 V

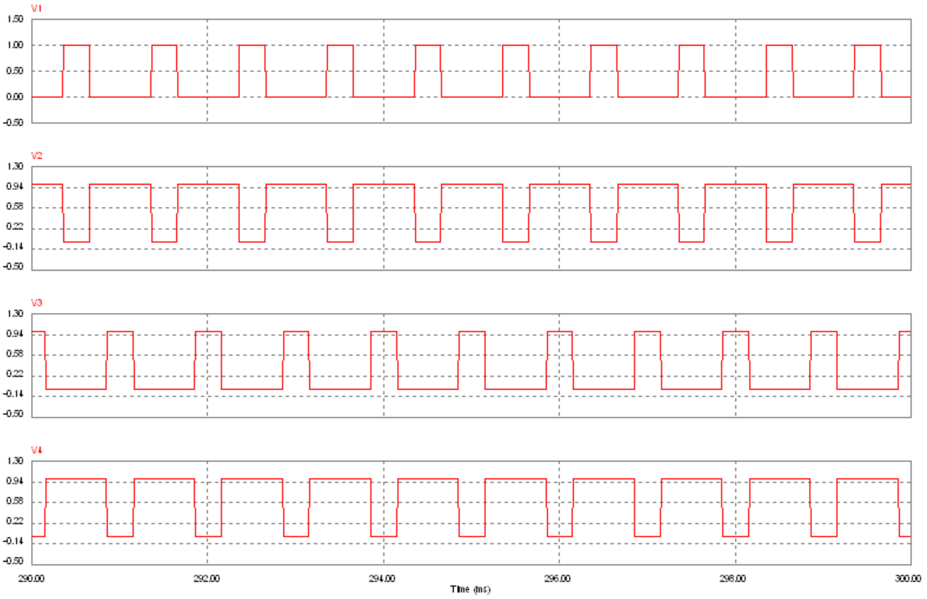


Fig. 4.13 Switching Wave-form of H- Bridge Inverter

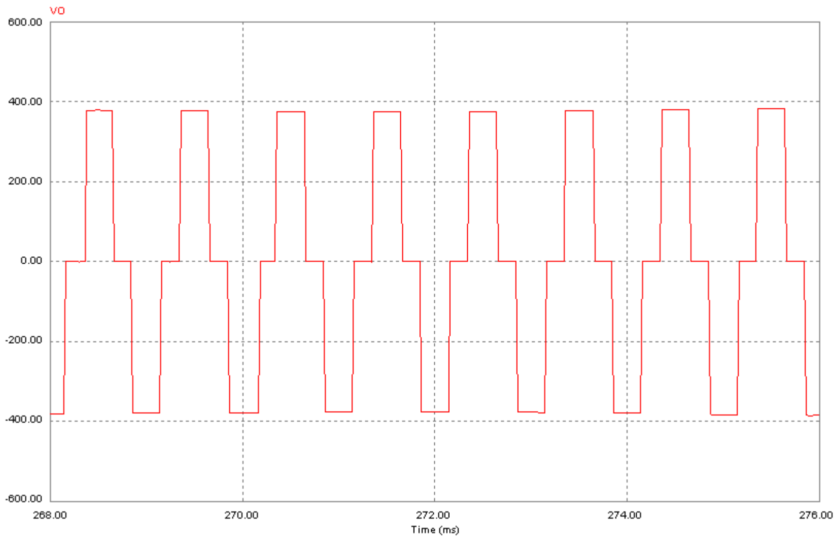


Fig. 4.14 output voltage of H-bridge Inverter, V_O = 400 V A.C.

4.5 Scaled down Model

Down scale version has been simulated after the full scale version. The component rating has been decided after this simulation. Fig. 4.15 shows the schematic for scaled down model of 24 Pulse Rectifier. The mains line-line voltages are taken as 415V rms with a supply frequency of 50Hz. The inductance of 24-Pulse rectifier is taken as 4.40 mH and capacitance is 4.7mF. Fig. 4.15 shows the schematic for scaled down model of 24 Pulse Rectifier.

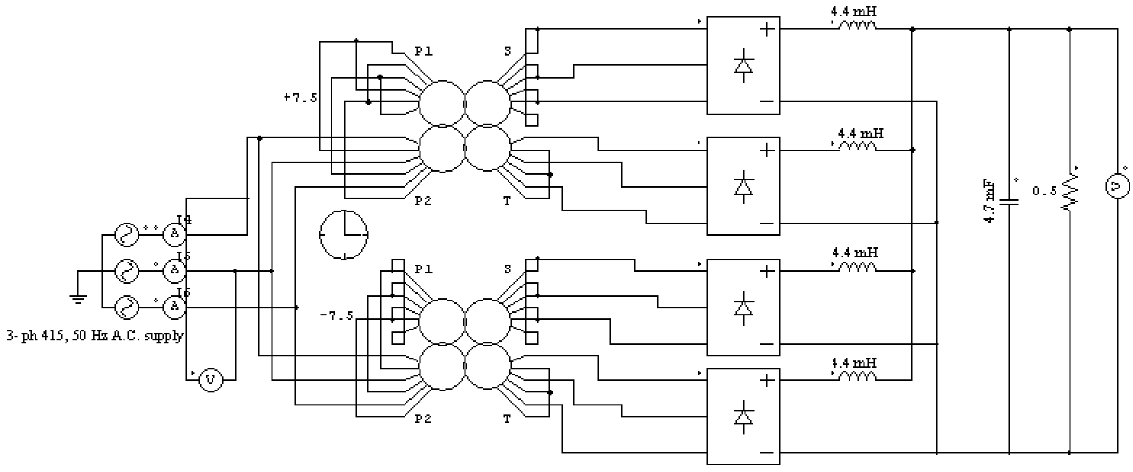


Fig. 4.15 Simulation schematic for scaled down model of 24 pulse rectifier.

4.5.1 Simulation Result

Fig. 4.16 shows the line current wave-form of 24 pulse rectifier. Fig. 4.17 shows the output voltage wave-form of 24 pulse rectifier. Here, we get approximately 25 V.

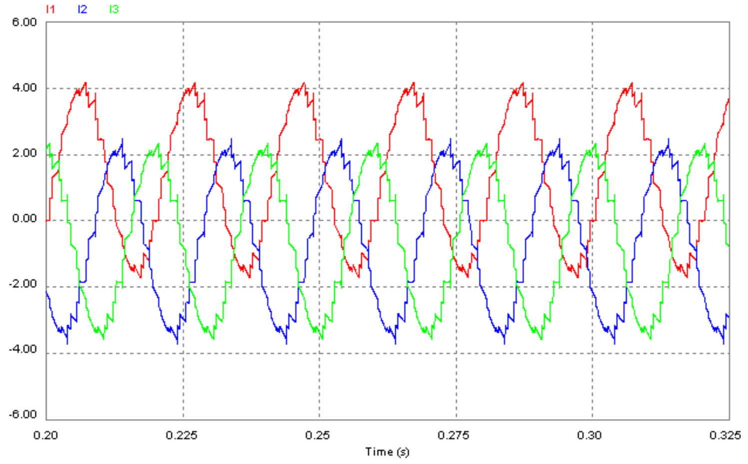


Fig. 4.16 Line current wave-form for scaled down model

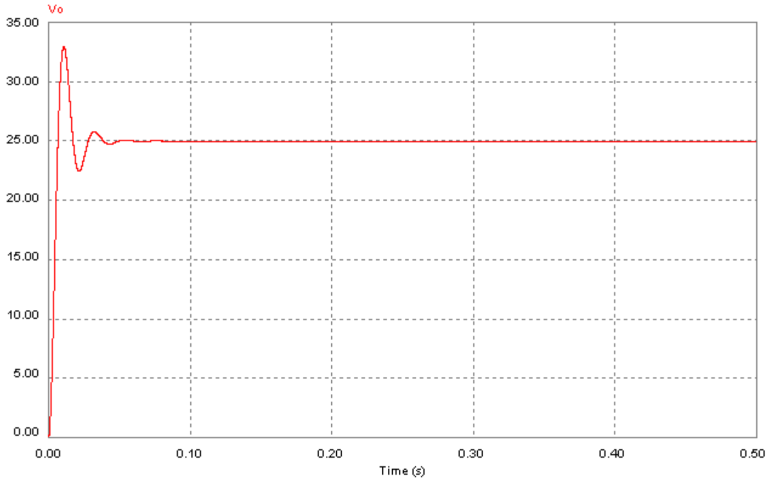


Fig. 4.17 Output voltage wave-form of 24 Pulse Rectifier, Vo = 25 V

4.6 Simulation of Power circuit for scaled down model

Fig. 4.18 shows the simulation schematic for scaled down model of power circuit. Here, output of 24 pulse rectifier is directly given to the H-bridge inverter. Here, no need to give the output of 24 pulse rectifier to buck chopper because load current is only 50 A of power circuit.

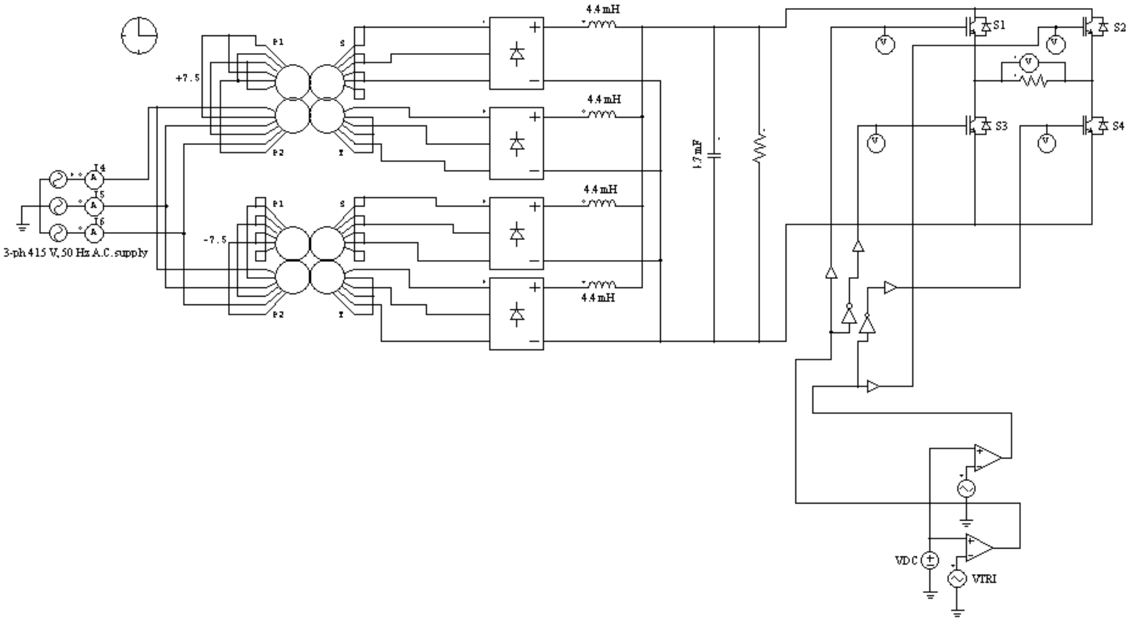


Fig. 4.18 Simulation schematic for scaled down model

4.6.1 Simulation Result

Fig. 4.19 shows the output voltage wave-form across load of H-bridge Inverter. Here, we get 25 V. Fig. 4.20 shows the switching wave-form of H-bridge inverter.

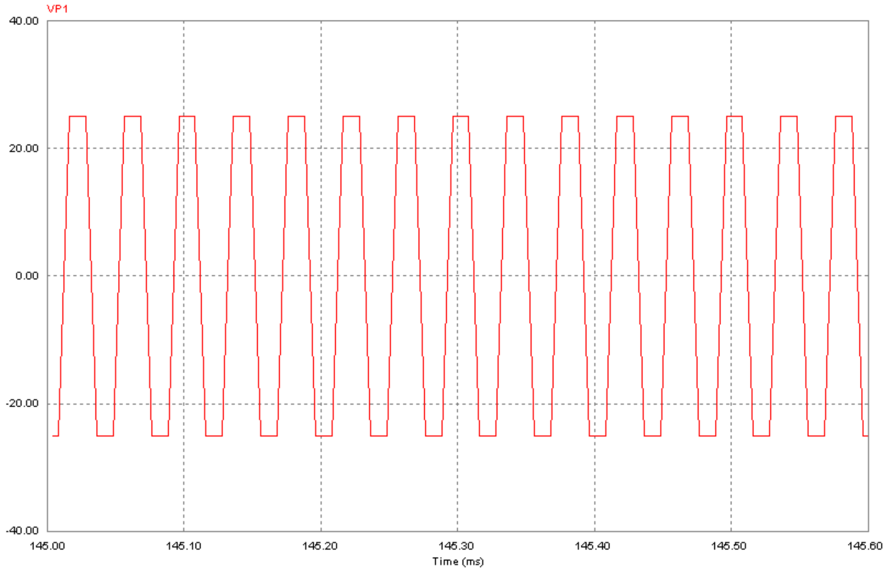


Fig. 4.19 Voltage Wave-form across load at H- bridge inverter side

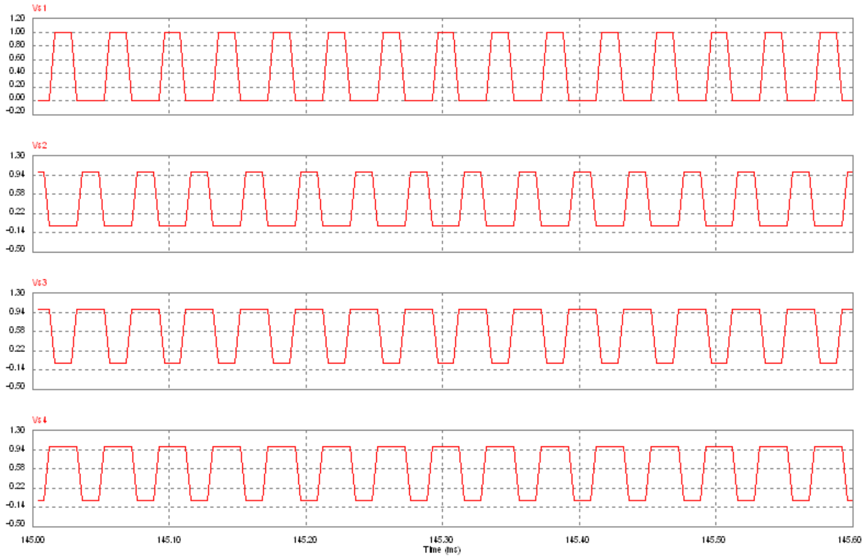


Fig. 4.20 switching wave-form of H-bridge inverter

CHAPTER 5

POWER LOSS CALCULATION & HEAT SINK DESIGN FOR THE SCALED DOWN MODEL

5.1 General

This chapter presents power loss calculation & Heat Sink design for scaled down model for different IGBT. Here, the calculation is done for 10 A current.

5.2 Power loss calculation for IGBT module

Whole IGBT Module consist of IGBT & Free-wheeling diode. First calculation is done for IGBT & then for Free-wheeling diode.

5.2.1 Power loss calculation for IGBT

IGBT Power dissipation loss = Conduction loss + Switching loss

= On-state loss + Turn-on loss + Turn-off loss

Ic= 10 A
f = 25 kHz

- From IGBT data sheet 1KW40T120

When Ic = 10 A then $V_{CE(sat)} = 1.9$ V, $E_{on} = 822$ µJ, $E_{off} = 910$ µJ and $V_F = 1.3$ V

- Conduction-loss for IGBT

$$Pcond = [V_{CE(sat)} * Ic * \delta]$$
$$= (1.9*10*0.5)$$
$$= 9.5 \text{ W} \quad (5.1)$$

- Switching loss for IGBT

$$Ps/w = [f * (E_{on} + E_{off})]$$
$$= \{25000*[(822+910)*10\text{^}-6]\}$$
$$= 43.3 \text{ W} \quad (5.2)$$

IGBT Power dissipation loss = Conduction loss + Switching loss

$$= 9.5 + 43.3$$
$$= 52.8 \text{ W} \quad (5.3)$$

5.2.2 Power loss calculation for Free-wheeling Diode

- Conduction-loss for Free-wheeling diode

$$P_{cond} = (V_F * I_F * T_{off})/ T$$
$$= (1.3*10*0.5)$$
$$= 6.5 \text{ W} \quad (5.4)$$

5.2.3 Total Power loss of IGBT Module

Total Power loss of IGBT Module = Power loss for IGBT + Power loss for Free-wheeling diode

$$= 52.8 + 6.5$$
$$= 59.3 \text{ W} \quad (5.5)$$

Table 5.1 Comparison of Power Loss Calculation for IGBT

IGBT NAME	Ic (A)	f kHz	δ	$V_{CE(sat)}$ V	Eon (mJ)	Eoff (mJ)	Power Loss for IGBT (W)		Ptotal for IGBT (W)
							Pcond	Ps/w	
1KW40T120	10	25	0.5	2.25	2.3	2.3	11.25	115	126.25
STGW30NC120HD	10	25	0.5	0.7	1.2	3.0	3.5	105	108.5
IRG4PH50KPbF	10	25	0.5	4.0	1.3	1.2	20	62.5	82.5
IRG4PH40KDPbF	10	25	0.5	4.0	1.2	1.1	20	57.5	77.5
IRGP30B120KD-E	10	25	0.5	1.9	0.8	0.9	9.5	43.3	52.8
IRG4PH50K	10	25	0.5	3.26	1.3	1.2	16.3	62.5	78.8
HGTG18N120BND	10	25	0.5	3	2	1.75	15	93.7	108.75
HGTG11N120CND	10	25	0.5	2.8	1.8	2.1	14	97.5	111.5

5.3 Heat Sink Design for Scaled down Model

I = 10 A

f = 25 kHz

Assume Tj(max) = 115 ºC

Ta(max) = 50 ºC

- From Table 6.1 Ptotal of IGBT Module = 52.8 W
- From the IRGP30B120KD-E IGBT data sheet

Rth(j-c) = 0.42 ºC / W

Rth(c-f) = 0.24 ºC/ W

- The junction temperature can be calculated using the following thermal equation:

Tj = Ptotal (IGBT Module) * {Rth(j-c) + Rth(c-f) + Rth(f-a)} + Ta (5.6)

Therefore,

- Thermal Resistance between heat sink and ambient temperature

Rth(f-a) = {[(115 – 50) /52.8] –[(0.42+ 0.24)]}

= 0.48 K / W (5.7)

- Junction temperature for IGBT

Tj(T) = [WT * Rth(j-c)] + [(WT + WD) * Rth(c-f)] + [(WT + WD) * Rth(f-a)] + Ta

= [52.8 * 0.42] + [(52.8 + 6.5) * (0.24)] + [(52.8 + 6.5) * 0.48] + 50

= 114.8 ºC (5.8)

- Junction temperature for Forward diode

Tj(D) = [WD * Rth(j-c)] + [(WT + WD) * Rth(c-f)] + [(WT + WD) * Rth(f-a)] + Ta

= [6.5 * 0.83] + [(52.8 + 6.5) * (0.24)] + [(52.8 + 6.5) * 0.288] + 50

=90.5 ºC (5.9)

Conclusion:

From the above heat sink design for IGBT based chopper we can conclude that to select a heat sink that can keep the junction Tj below the Tj(max).

Table 5.2 Comparison of Heat sink design

IGBT NAME	**Ic (A)**	**f kHz**	**δ**	**Rth(j-c) (K/W)**	**Rth(c-f) (K/W)**	**Rth(f-a) (K/W)**
1KW40T120	10	25	0.5	0.45	0.3	-0.27
STGW30NC120HD	10	25	0.5	0.57	0.3	-0.32
IRG4PH50KPbF	10	25	0.5	0.64	0.24	-0.15
IRG4PH40KDPbF	10	25	0.5	0.77	0.24	-0.23
IRGP30B120KD-E	10	25	0.5	0.42	0.24	0.48
IRG4PH50K	10	25	0.5	0.64	0.24	-0.11
HGTG18N120BND	10	25	0.5	0.32	0.3	-0.06
HGTG11N120CND	10	25	0.5	0.42	0.3	-0.18

From the above table we can conclude that IRGP30B120KD-E IGBT is suitable for scaled down model because it has positive thermal resistance.

CHAPTER-6

ASSEMBLY OF 24 PULSE RECTIFIER FOR SCALED DOWN MODEL

6.1 Ratings for Scaled Down Model

- Power Rating = 1.2 KW
- **Transformer Rating:**

I/P Voltage: 415 V
O/P: 18.5 V Delta / 18.5 V Star
f = 50 Hz
Cooling: AN
VA Type: ISO
Class – F

- **Inductor Rating :**

L = 4.40 mH
I = 20 A
Type: DC
f = 1 kHz
Ph. 1

- **Capacitor Rating:**

C = 4700 uF, 450 V

- **Current Transformer Rating** :

Turns Ratio : 1000:1
AC1025

6.2 Calculations

Here, Power loss assumed to be 1.2 KW

Transformer I/P voltage = 415 V

O/P of transformer = 18.5 V

So, we getting voltage Vo = 18.5 * 1.35

= 25 V (6.1)

- For the Calculation of Current:

By Power Equation,

P = VI

So,

I = P/ V

= (1.2 * 1000) / 25

= 50 A (6.2)

- For the Calculation of Resistor:

P = (V*V) / R

So,

R = (V*V) / P

= (25*25) / (1200)

= 0.5 Ω (6.3)

6.3 Hardware Results

Fig. 6.1 shows the Line current wave-form of 24 Pulse Rectifier. Here we get 24 pulse operation.

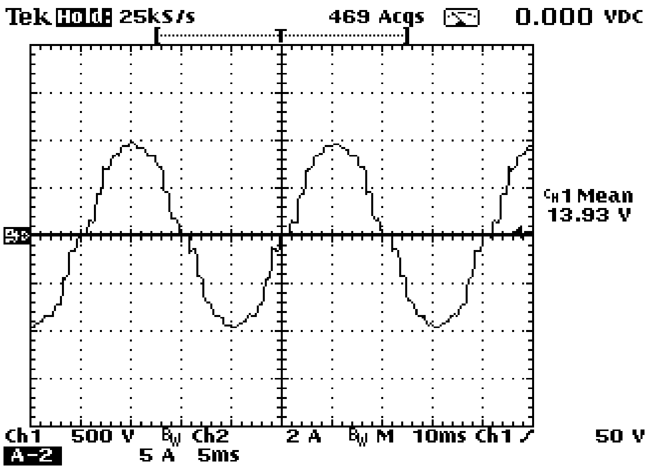


Fig. 6.1 Line current wave-form of 24 Pulse Rectifier

Fig. 6.2 shows the output voltage wave-form of 24 pulse diode rectifier. Here, we get 25 V because output of transformer is 18.5 V. So, we get Vo = 18.5 * 1.35 = 25 V as our requirements.

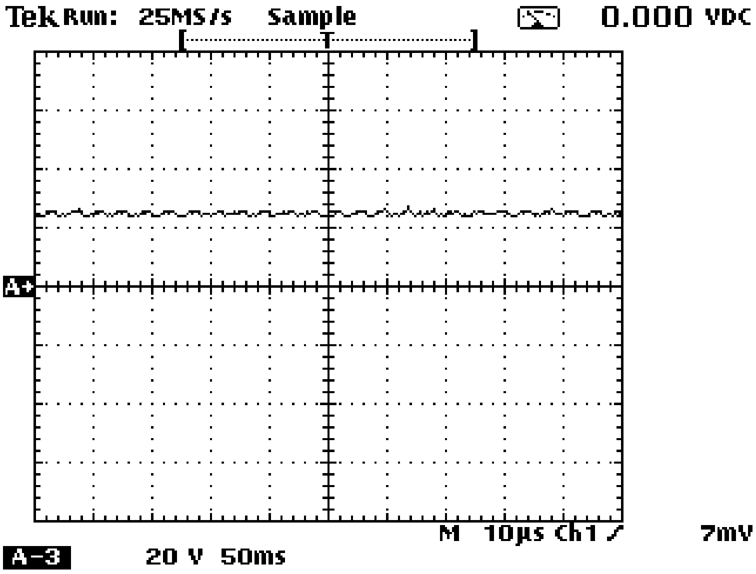
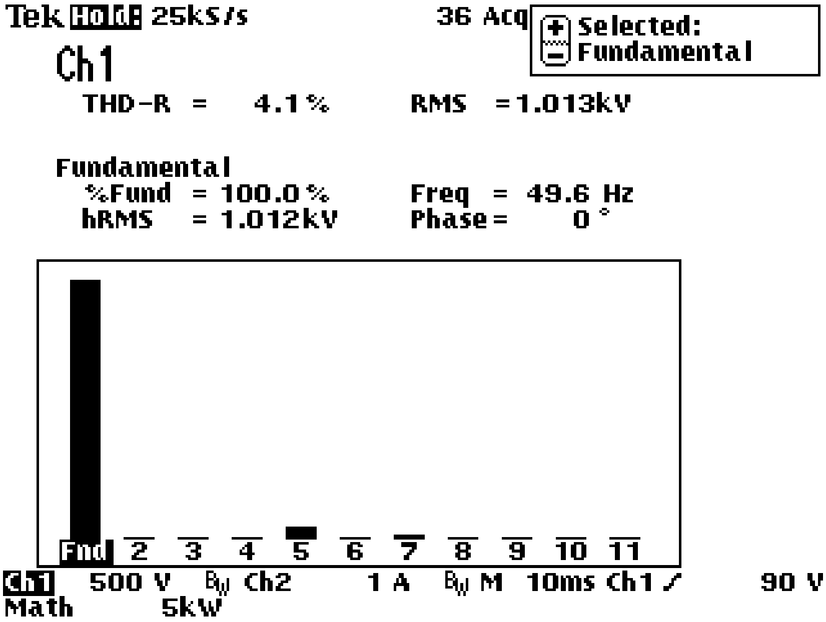


Fig. 6.2 Dc-link output Voltage V = 25 V, Waveform of 24 Pulse Rectifier

Fig 6.3 shows the THD measurement wave-form for 24 pulse rectifier. For multi pulse THD is reduced. Here, we get 4.1 % THD for 24 pulse rectifier.



Tek Hold 25kS/s 36 Acq

Selected: Fundamental

Ch1

THD-R = 4.1 % RMS = 1.013kV

Fundamental

%Fund = 100.0 % Freq = 49.6 Hz

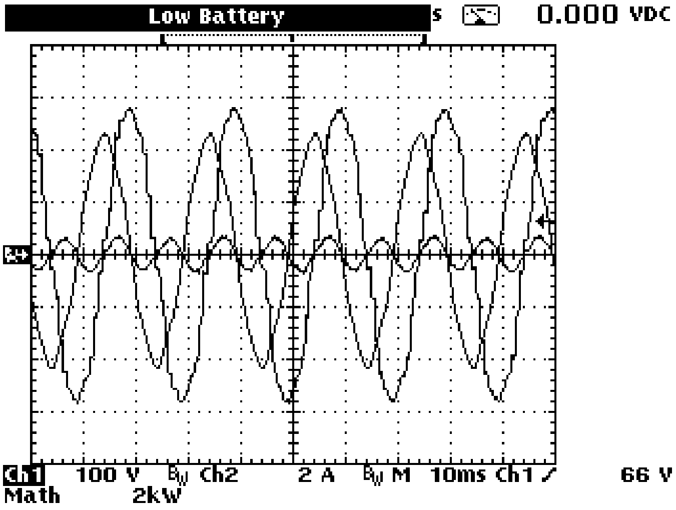
hRMS = 1.012kV Phase = 0 °

Ch1 500 V Bw Ch2 1 A Bw M 10ms Ch1 ⁄ 90 V

Math 5kW

Fig. 6.3 THD Measurement 4.1%

Fig. 6.4 shows the Line current I1, I2 and wave-form with phase shifting angle 15˚.



Low Battery 0.000 VDC

Ch1 100 V Bw Ch2 2 A Bw M 10ms Ch1 ⁄ 66 V

Math 2kW

Fig. 6.4 Line Current Waveform, phase shifted by 15˚

CHAPTER -7

TESTING OF CHOPPER CONTROLLER CIRCUIT

7.1 Test Circuit

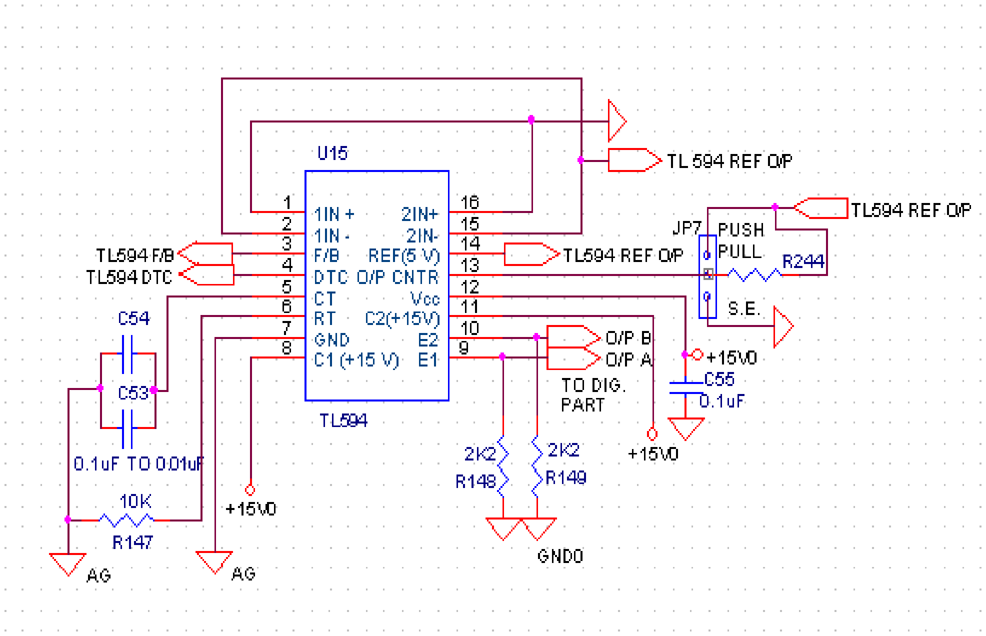


Fig. 7.1 Test Circuit

7.2 Test Circuit Result

Fig. 7.2 shows the Test circuit wave-form, here we get PWM wave-form with 180° phase shift.

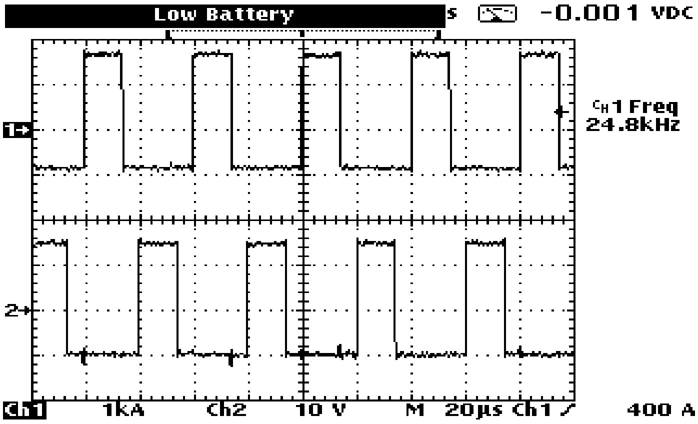


Fig. 7.2 Test circuit output voltage wave-form

7.2.1 Waveforms for Different Duty Cycle

Fig. 7.3 shows the PWM wave-form for 50% duty cycle & 10 kHz frequency. Fig. 7.4 shows the PWM wave-form for 10 kHz frequency & 90 % duty cycle. Fig. 7.5 shows the PWM wave-form for 25 kHz frequency & 50 % duty cycle. Fig. 7.6 shows the PWM wave-form for 25 kHz frequency & 90 % duty cycle.

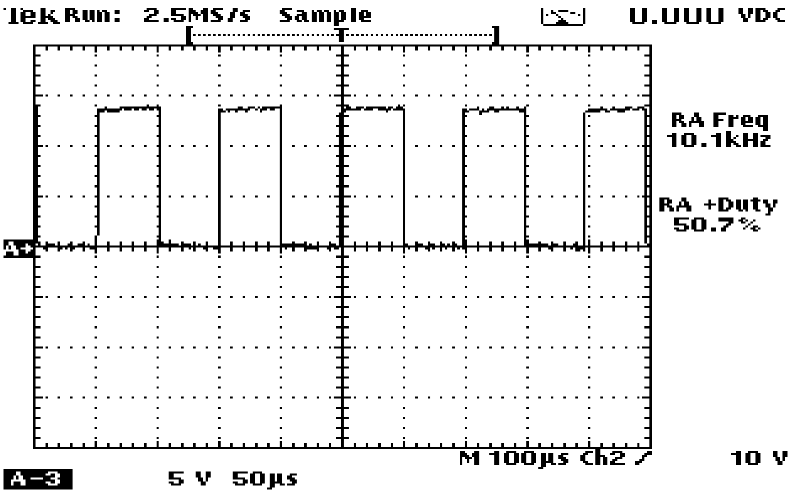


Fig. 7.3 PWM wave-form, f= 10 kHz, Duty cycle = 50 %

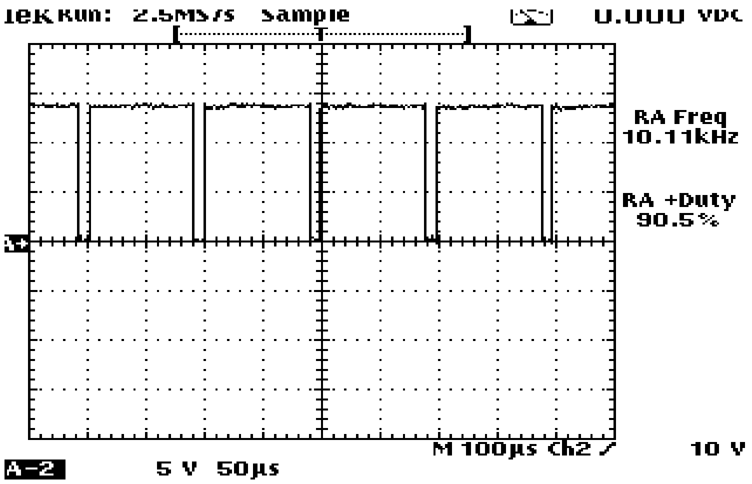


Fig.7.4 PWM wave-form when f =10 kHz, Duty cycle = 90.5%

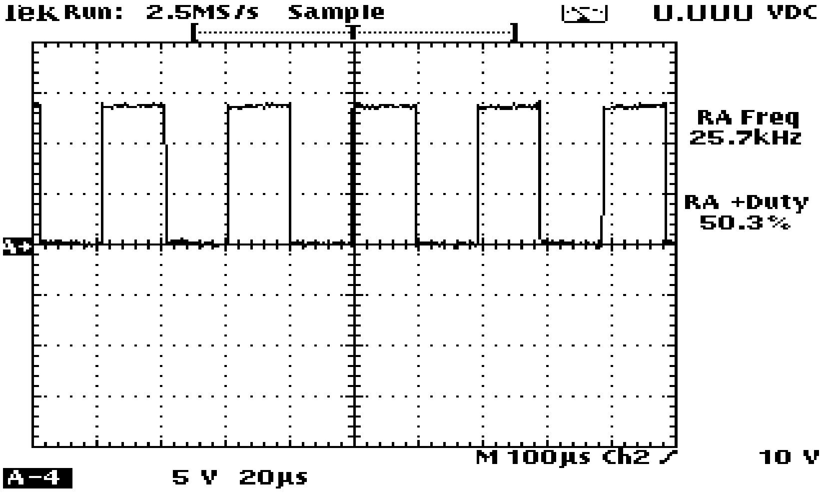


Fig. 7.5 PWM wave-form when f =25 kHz, Duty cycle = 50%

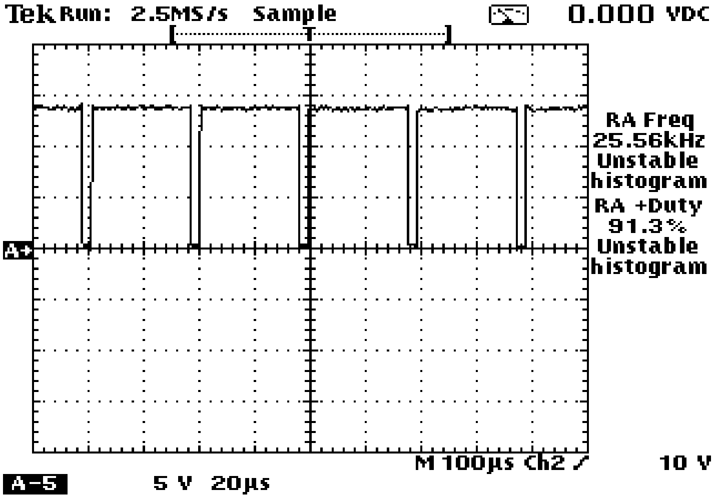


Fig. 7.6 PWM wave-form when f =25 kHz, Duty cycle = 90%

CHAPTER - 8

PROTOTYPE ASSEMBLY FOR H-BRIDGE INVERTER

8.1 Testing of Driver card for H-bridge Inverter:

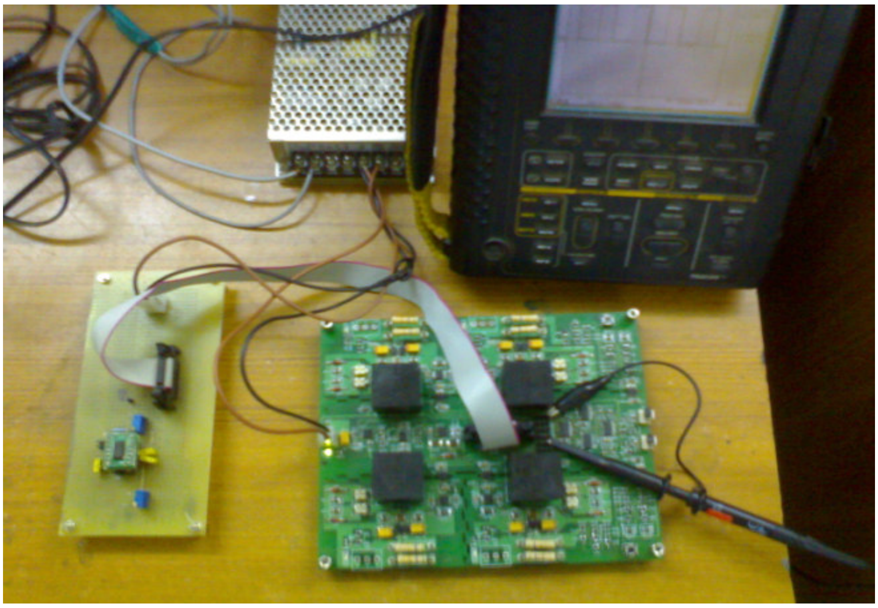


Fig. 8.1 Testing of Driver card for H-bridge Inverter

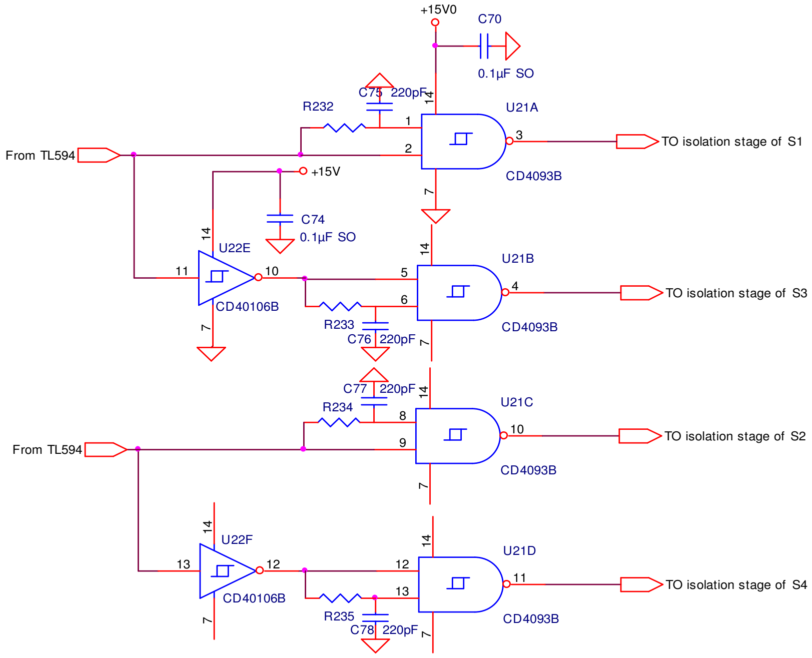


Fig. 8.2 Driver card PWM generator circuit

Four channel isolated driver card is used to perform the H-bridge inverter. Driver card provide four PWM pulses to the IGBT. TL594 provides two PWM signals which are 180 phase shifted as shown in fig 2. These two pulses are given to the driver card. Here, we provide dead band = 220pF*910E = 220nSec, as shown in fig. 2. From this stage we get four switching PWM pulses S1, S2, S3 and S4. Here S1 and S2 pulses are 180 phase shifted as shown in fig 8.4. S2 and S4 pulses are inverted as shown in fig 8.6, same for S1 & S3.

Here, for isolation ISO721 IC used. It provides isolation up to 4000V. Used in conjunction with isolated power supplies, these devices prevent noise currents on a data bus and interfacing with or damaging sensitive circuitry. Across the isolation barrier, a differential comparator receives the logic transition information, then sets or resets a flip flop and the output circuit accordingly.

8.2 Driver card test set up

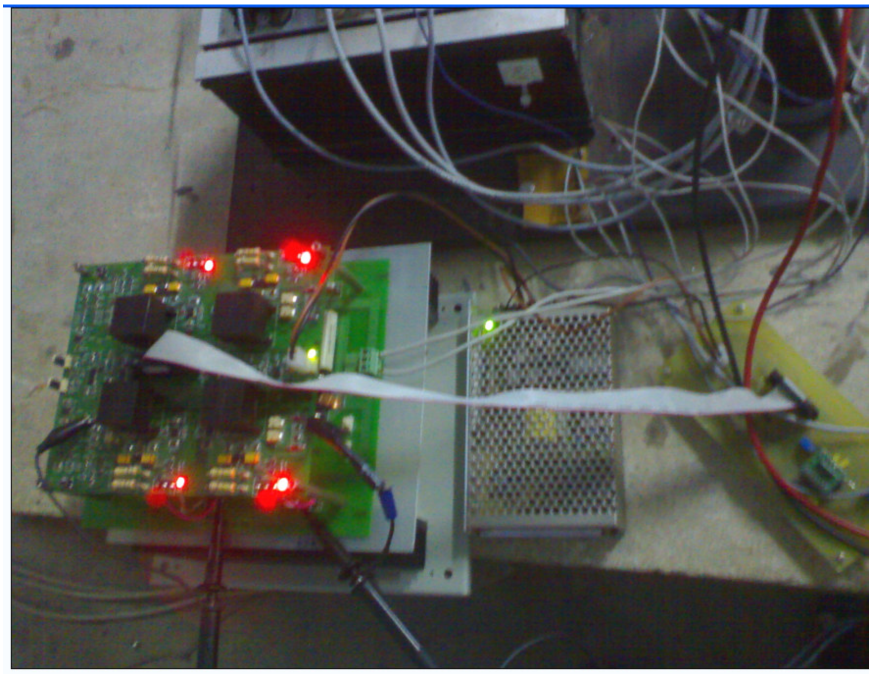


Fig. 8.3 driver card test set up

8.2.1 Driver card testing result

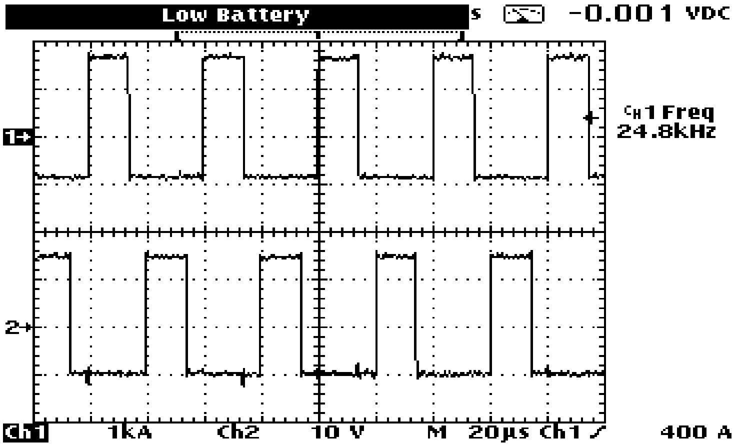


Fig. 8.4 Switching waveform when switch s1 & s2 on

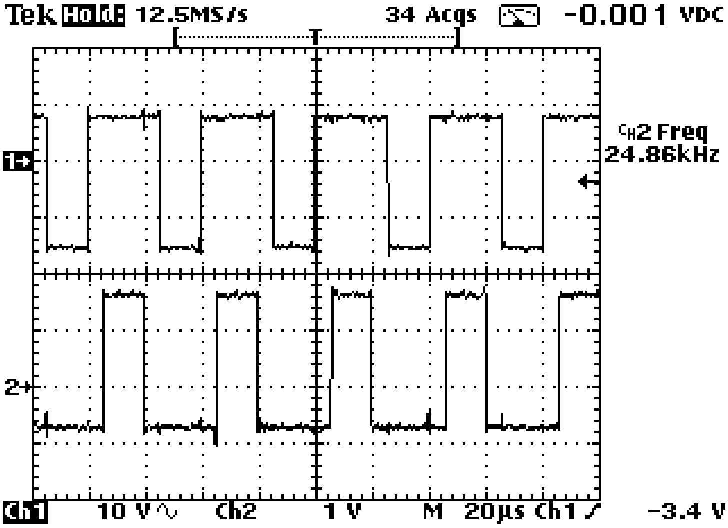


Fig. 8.5 Switching waveform when switch s4 & s1 on

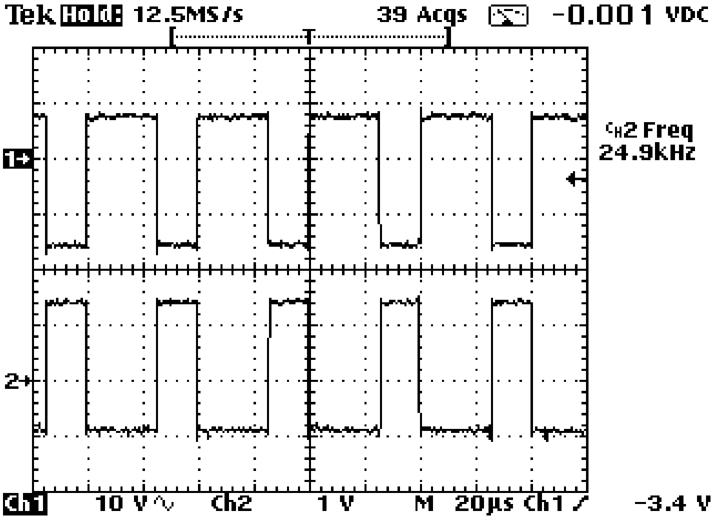


Fig. 8.6 Switching waveform when s4 & s2 on

CHAPTER - 9

EXPERIMENTAL RESULTS

9.1 Prototype Hardware Model for Scaled down Model

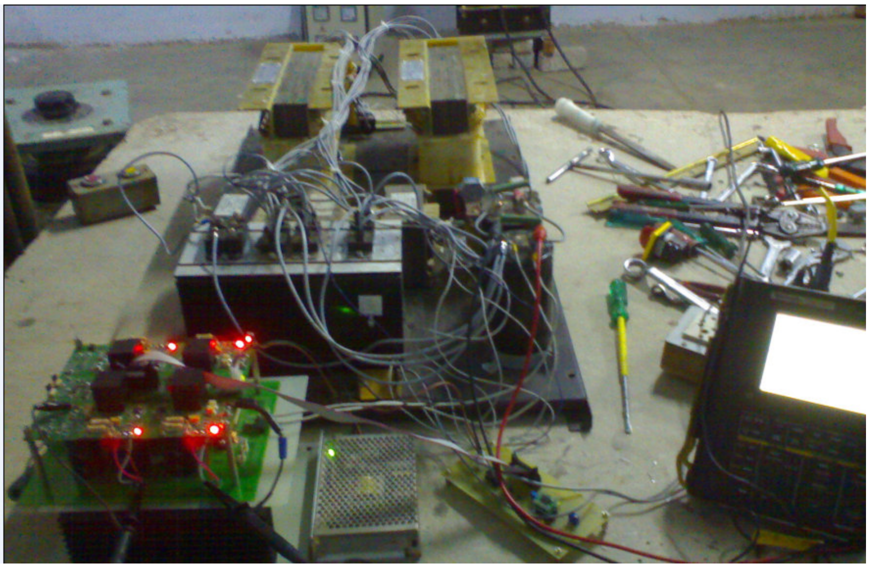


Fig. 9.1 Prototype Model for scaled down version

9.2 Hardware Result for Scaled down Model

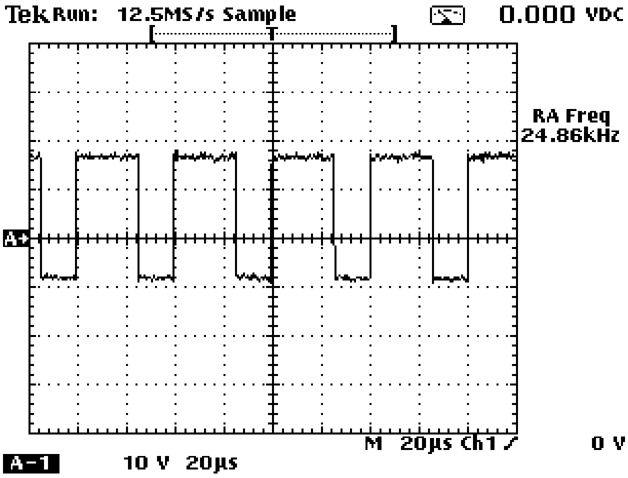


Fig.9.2 switching wave-form when S1 is on

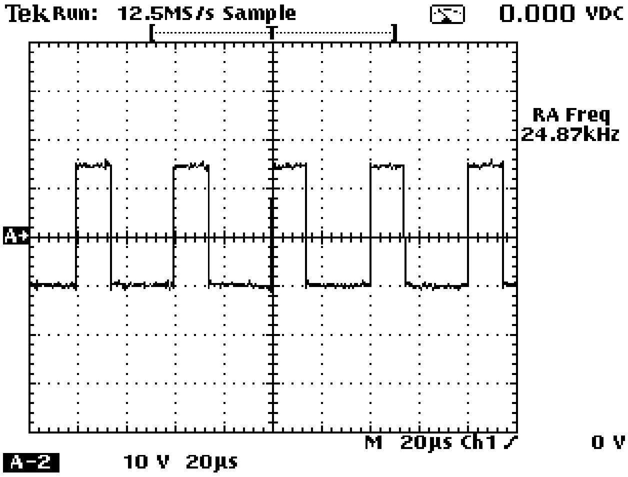


Fig. 9.3 Switching wave-form when S2 is on

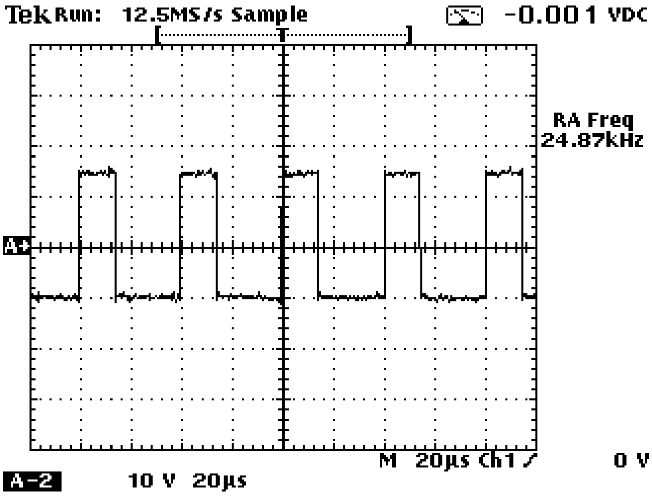


Fig.9.4 switching wave-form when S3 is on

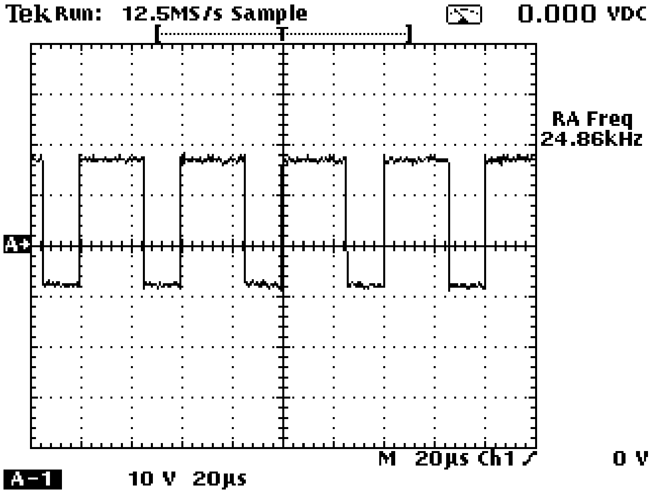


Fig.9.5 Switching wave-form when S4 is on

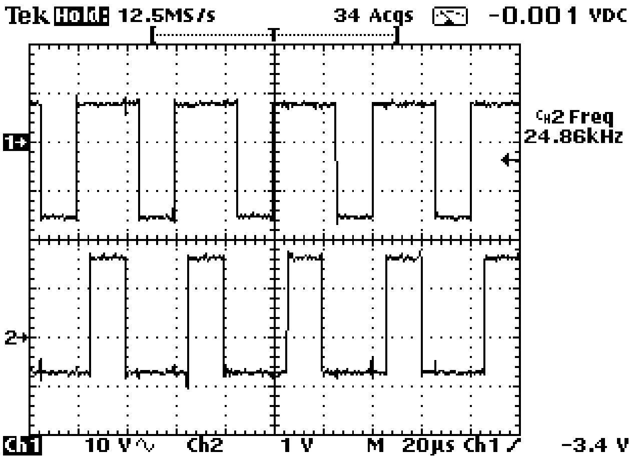


Fig.9.6 Switching waveform when switch s4 & s1 on

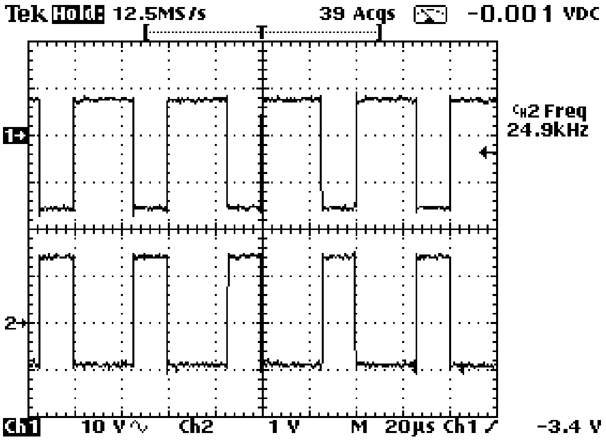


Fig. 9.7 Switching wave-form when s4 &s2 on

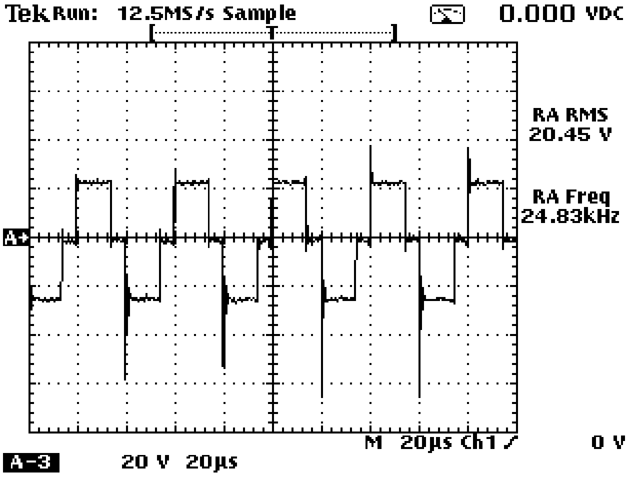


Fig. 9.8 Voltage wave-form across load R = 50 Ω

CHAPTER - 10

CONCLUSION & FUTURE WORK

10.1 Conclusion

24-Pulse transformer is proposed to achieve the 24-Pulse operation and it is used to reduce the line current harmonic distortion. Thus we improve the T.H.D. with the multi pulse operation. After Completion of 24 Pulse operation we get THD approximately 4.1 %.

From the testing of Chopper Controller Circuit, we get two PWM waveform of 25 kHz and both are 180 phase shifted.

From the testing of whole set up we conclude that we get 20 V RMS at 25 kHz frequency at the output stage of H-bridge inverter across the load. Load resistance is of 50 Ω.

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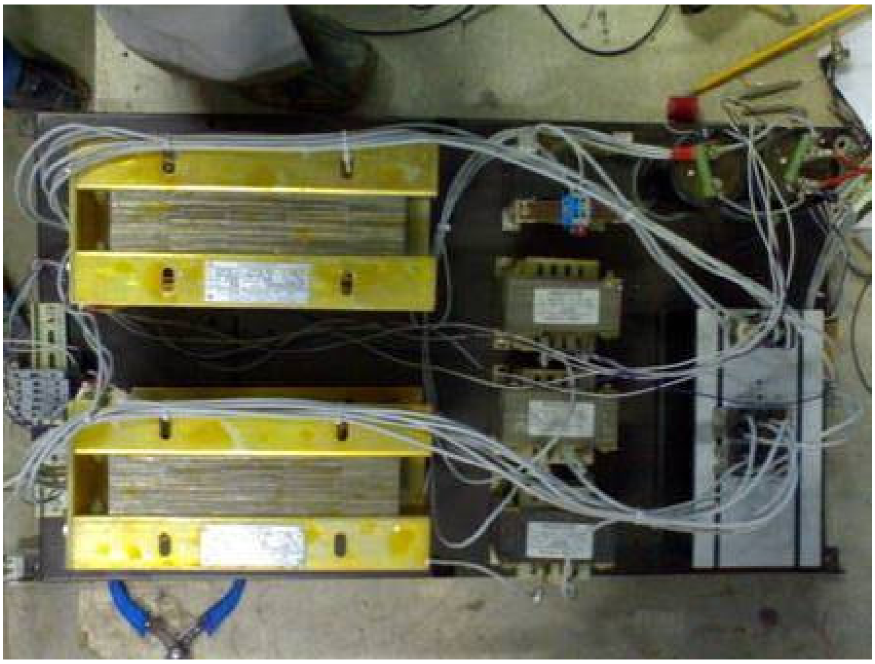
8. Oliveira Jr, D.S., Bissochi Jr, C.A., Tavares, Douglas, Vieira Jr, J.B, Farias, V.J., Freitas, L.C. “Proposal of an Isolated Current Controlled Inverter” IEEE Transactions, 2000 pp 337-340.
9. J. Rodriguez, J. Pontt, R. Huerta, P. Newman “24-Pulse Active Frondt End Rectifier with Low Switching Frequency” IEEE Power Electronics Specialiststs Conference, 2004 pp 3517-3523.
10. Mohan, Undeland, Robbins, Power Electronics, 3rd ed. John Wiley & Sons, Inc.
11. Texas Instrument www. ti.com
12. www.irf.com

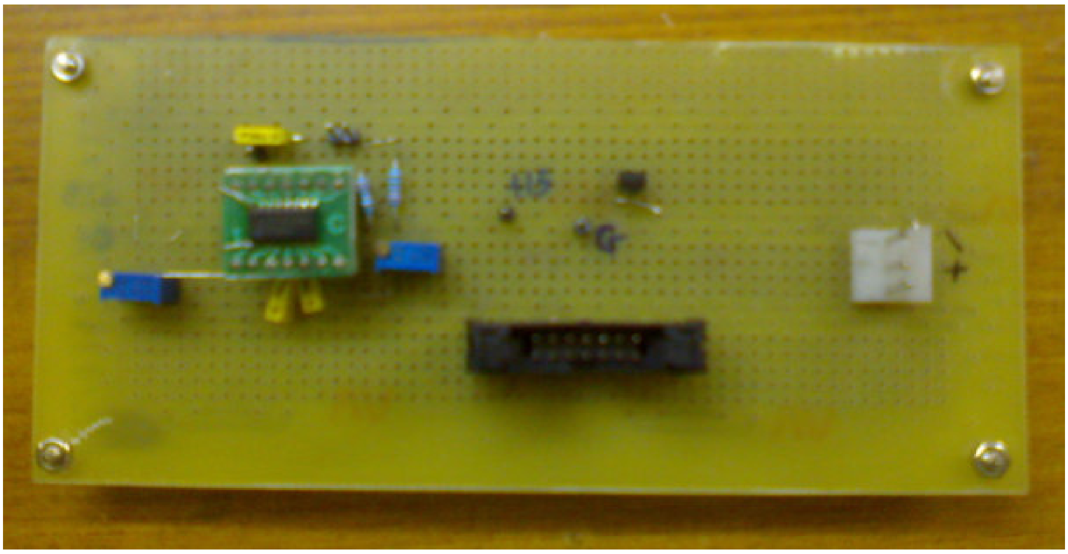
APPENDIX-1

EXPERIMENTAL SET UP

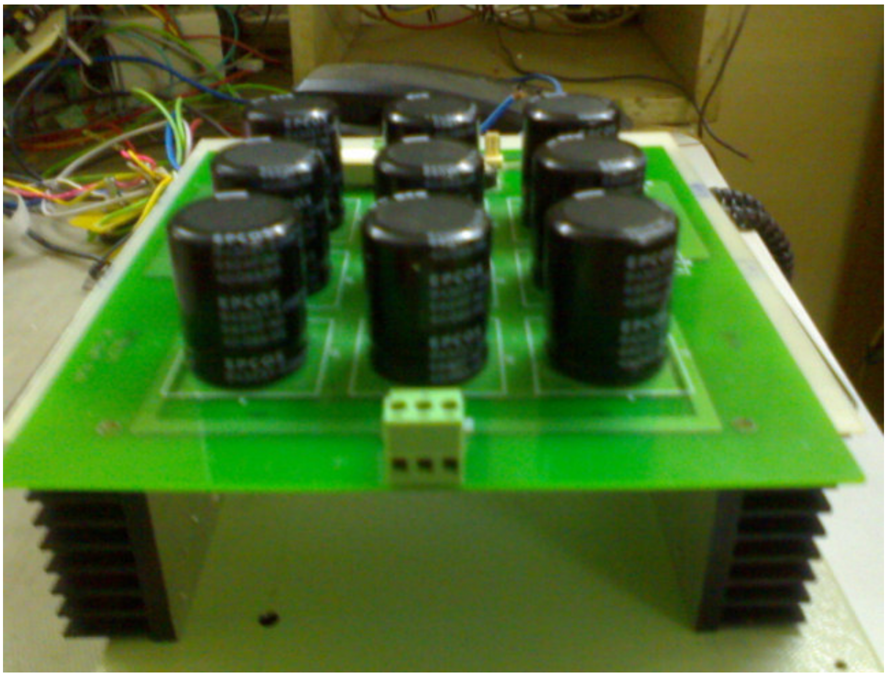
(1) Hardware set up for 24 pulse rectifier for scaled down version



(2) Test circuit PCB for testing of chopper controller



(3) H-bridge Power card



APPENDIX-B

DATSHEET

(1) BS300GB120DLC Data Sheet

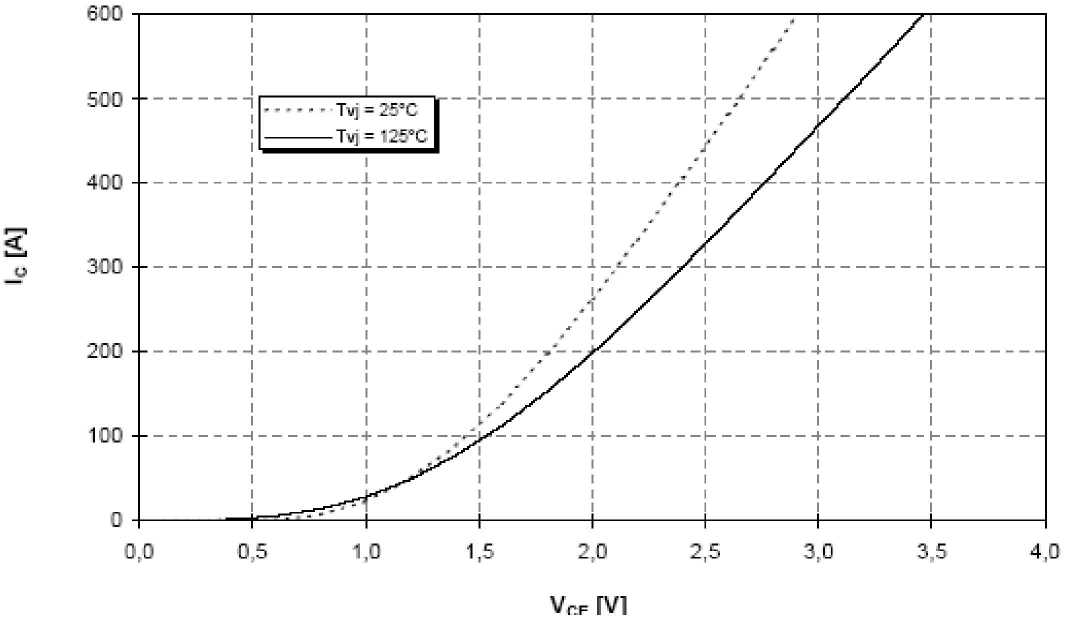


Fig. Collector current Vs Collector to emitter voltage characteristic

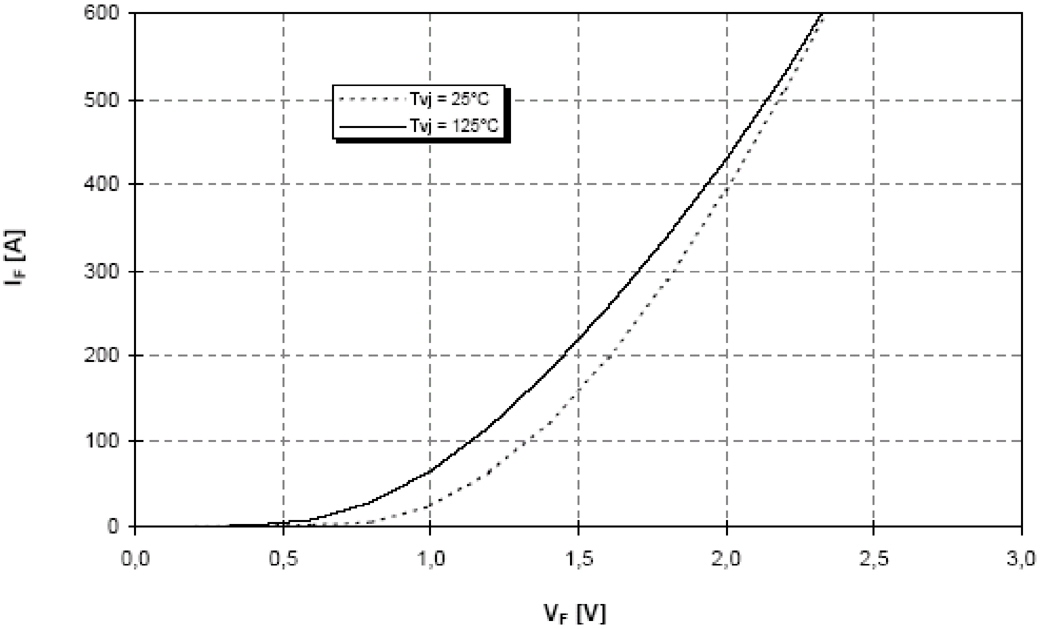


Fig. Forward characteristic of inverse diode

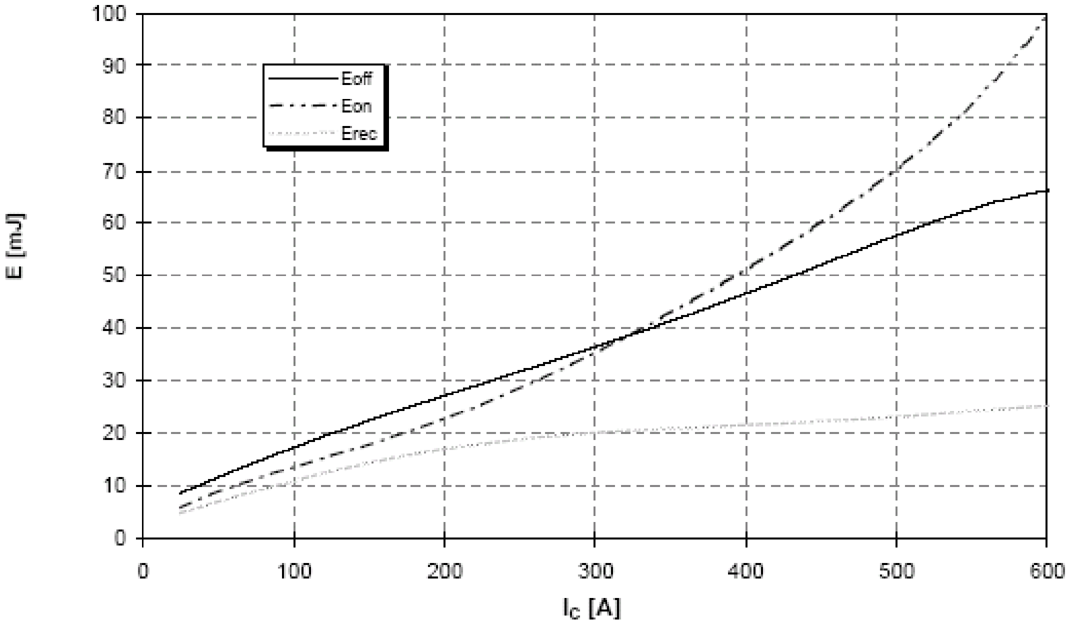


Fig. 3.4 Characteristic for switching losses

(2) 1KW40T120 Data sheet

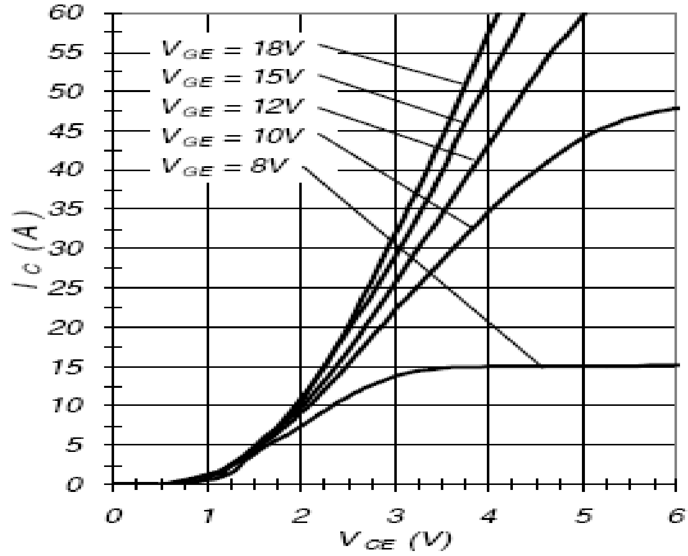


Fig. 5.1 Collector current vs. Collector-Emitter Voltage

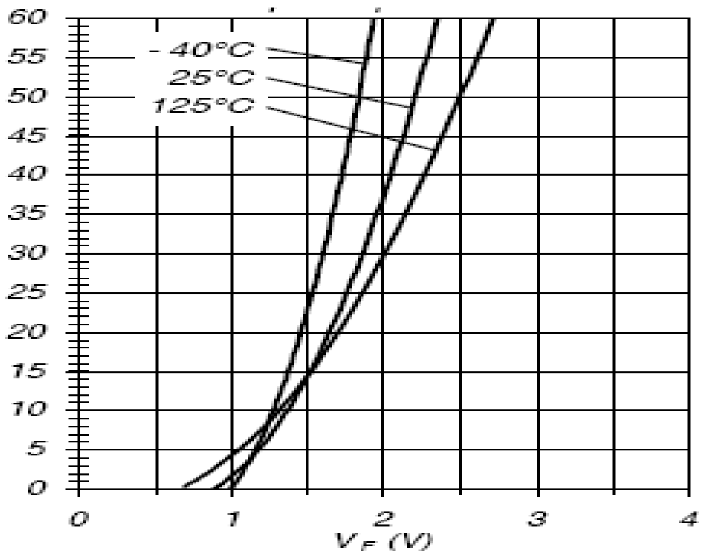


Fig. 5.2 Forward current vs. Forward voltage

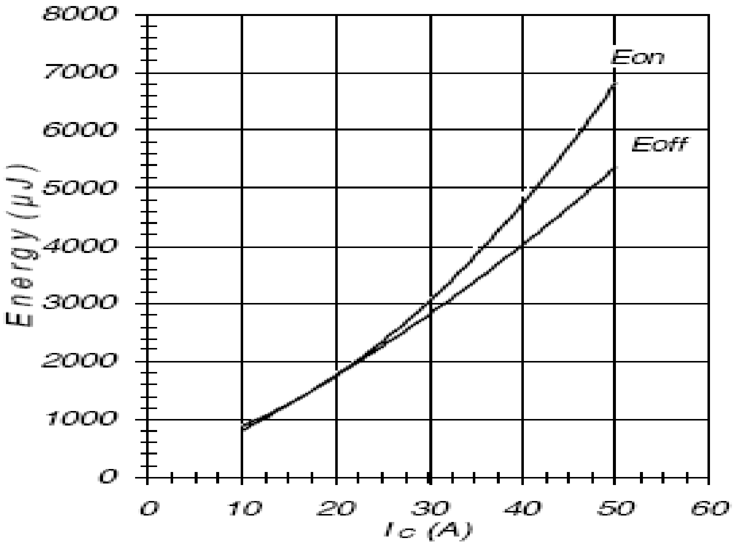
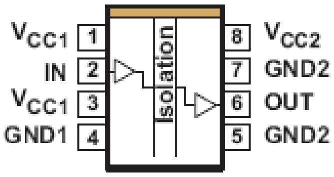


Fig. 5.3 Energy loss vs. Collector current

(3) ISOLATION IC ISO721, 722 M

- **PIN DIAGRAM**

PACKAGE PIN ASSIGMENTS
ISO721D, ISO721MD
(TOP VIEW)

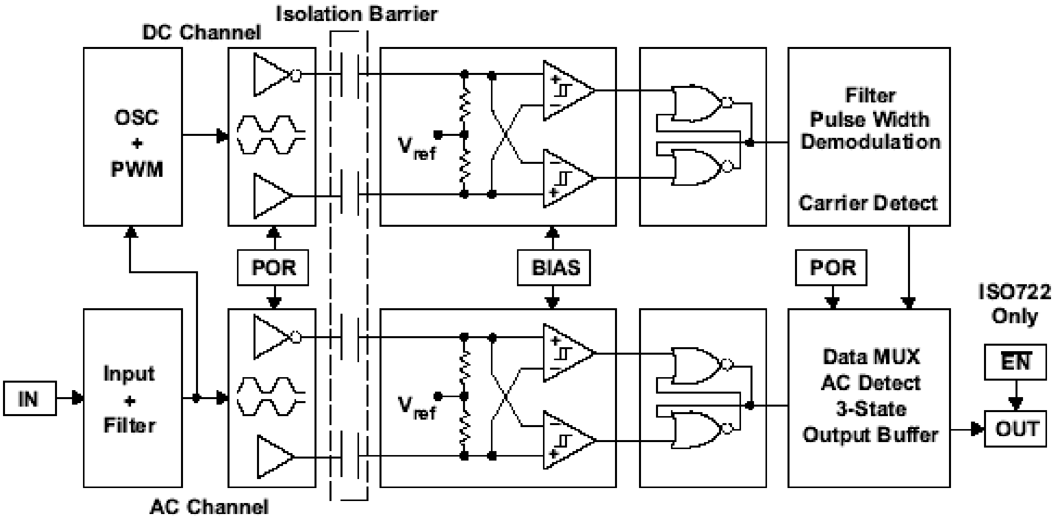


- **SOME FEATURES OF ISO721**
  - 4000-V(peak) Isolation, 560-Vpeak VIORM
  - Signaling Rate 0 Mbps to 150 Mbps
  - Low-Propagation Delay
  - Low-Pulse Skew (Pulse-Width Distortion)
  - Low-Power Sleep Mode
  - High-Electromagnetic Immunity
  - Low-Input Current Requirement
  - Failsafe Output
  - Drop-In Replacement for Most Opto and agnetic Isolators

APPLICATIONS

- Industrial Fieldbus
- Computer Peripheral Interface
- Servo Control Interface
- Data Acquisition

FUNCTIONAL BLOCK DIAGRAM



➢ DESCRIPTION OF ISO721M, ISO722 M

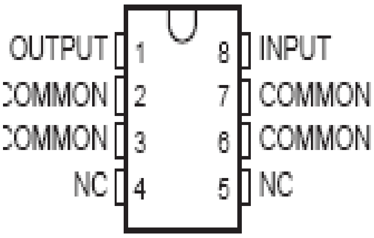
The ISO721, ISO721M, ISO722, and ISO722M are digital isolators with a logic input and output buffer separated by a silicon oxide (SiO2) insulation barrier. This barrier provides galvanic isolation of up to 4000 V. Used in conjunction with isolated power supplies, these devices prevent noise currents on a data bus or other circuits from entering the local ground,. and interfering with or damaging sensitive circuitry. A binary input signal is conditioned, translated to a balanced signal, then differentiated by the capacitive isolation barrier. Across the isolation barrier, a differential comparator receives the logic transition information, then sets or resets a flip flop and output circuit accordingly. A periodic update pulse is sent across the barrier to ensure the proper dc level of the output. If this dc-refresh pulse is not received for more than 4 ms, the input is assumed to be unpowered or not being actively driven and the failsafe circuit drives the output to a logic high state.

The symmetry of the dielectric and capacitor within the integrated circuitry provides for close capacitive matching, and allows fast transient voltage changes between the input and output grounds without corrupting the output. The small capacitance and resulting time constant provide for fast operation with signaling rates(1) from 0 Mbps (dc) to 100 Mbps for the ISO721/ISO722, and 0 Mbps to 150 Mbps with the

ISO721M/ISO722M. These devices require two supply voltages of 3.3-V, 5-V, or any combination. All inputs are 5-V tolerant when supplied from a 3.3-V supply and all outputs are 4-mA CMOS. The ISO722 and ISO722M devices includes an active-low output enable that when driven to a high-logic level, places the output in a high-impedance state, and turns off internal bias circuitry to conserve power. Both the ISO721 and ISO722 have TTL input thresholds and a noise-filter at the input that prevents transient pulses of up to 2 ns in duration from being passed to the output of the device. The ISO721M and ISO722M have CMOS VCC/2 input thresholds, but do not have the noise-filter and the additional propagation delay. These features of the ISO721M also provide for reduced jitter operation. The ISO721, ISO721M, ISO722, and ISO722M are characterized for operation over the ambient temperature range of –40°C to 125°C.

(4) TL750LO5CD

- **PIN DIAGRAM**



- SOME FEATURE OF TL750L05CD
  - Very Low Dropout Voltage, Less Than 0.6 V at 150 mA
  - Very Low Quiescent Current
  - TTL and CMOS-Compatible Enable on TL751L Series
  - 60 V Load-Dump Protection
  - Reverse Transient Protection Down To 50 V
  - Internal Thermal-Overload Protection
  - Over voltage Protection
  - Internal Over current-Limiting Circuitry Less Than 500 μA Disable

- **DESCRIPTION OF TL750L05CD**

The TL750L and TL751L series of fixed-output voltage regulators offer 5-V, 8-V, 10-V, and 12-V options. The TL751L series also has an enable (ENABLE) input. When ENABLE is high, the regulator output is placed in the high-impedance state. This gives the designer complete control over power up, power down, or emergency shutdown. The TL750L and TL751L series are low-dropout positive-voltage regulators specifically designed for battery-powered systems. These devices incorporate overvoltage and current-limiting protection circuitry, along with internal reverse-battery protection circuitry to protect the devices and the regulated system. The series is fully protected against 60-V load-dump and reverse-battery conditions. Extremely low quiescent current during full-load conditions makes these devices ideal for standby power systems.

APPENDIX C

OVEVIEW OF CHOPPER CONTROLLER IC TL 594

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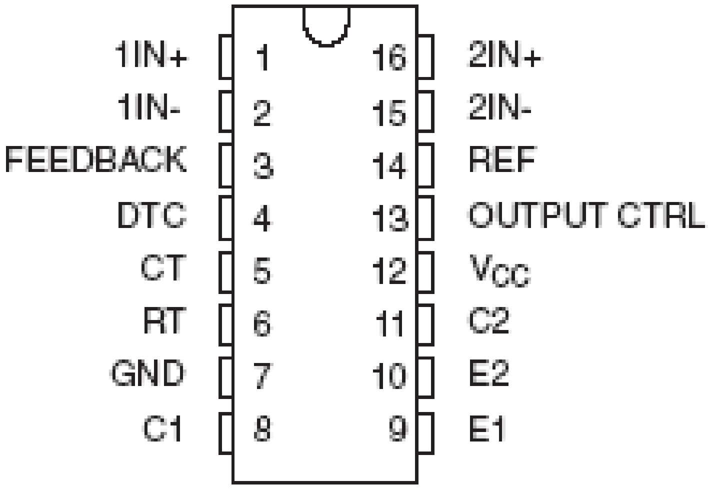
(1) Pin Diagram of TL594

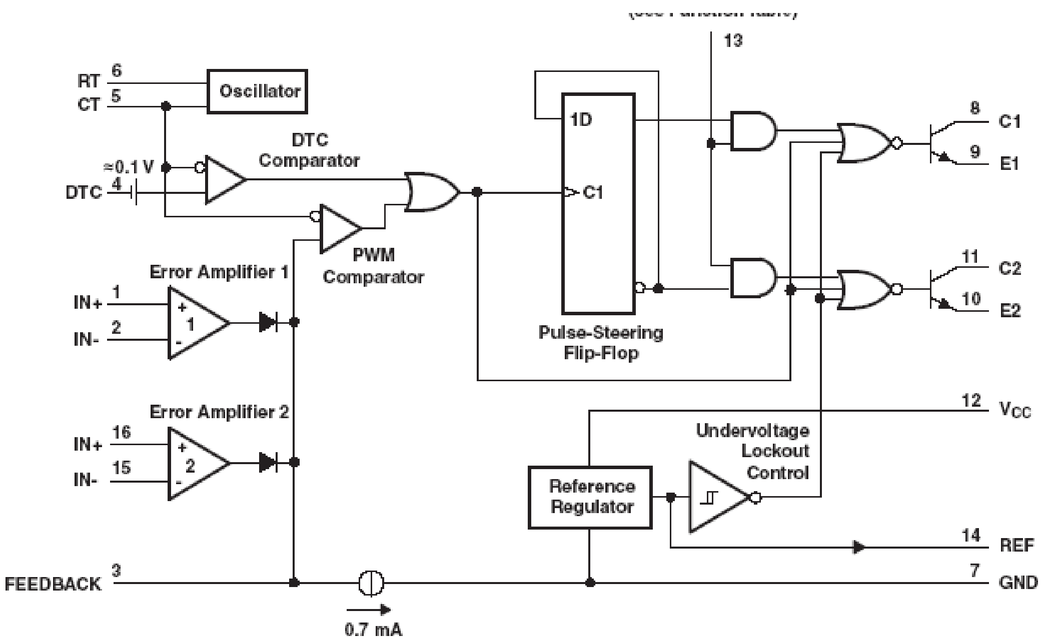


(2) Some Important Features of TL 594

- Complete PWM Power- control Circuitry
- Uncommitted Outputs for 200-mA Sink or Source Current
- Output Control Selects Single-Ended or Push-Pull Operation
- Internal Circuitry Prohibits Double Pulse at either Output
- Variable Dead Time Provides Control Over total Range
- Internal Regulator Provides a Stable 5-V reference Supply Trimmed to 1%

  Circuit Architecture Allows Easy Synchronization

(4) Functional Block Diagram



(5) Description of TL594

The TL594 incorporates all the functions required in the construction of a pulse-width-modulation (PWM) control circuit on a single chip. Designed primarily for power-supply control, this device offers the systems engineer the flexibility to tailor the power-supply control circuitry to a specific application. The TL594 contains two error amplifiers, an on-chip adjustable oscillator, a dead-time control (DTC) comparator, a pulse-steering control flip-flop, a 5-V regulator with a precision of 1%, an under voltage lockout control circuit, and output control circuitry.

The error amplifiers have a common-mode voltage range of –0.3 V to VCC – 2 V. The DTC comparator has a fixed offset that provides approximately 5% dead time. The on-chip oscillator can be bypassed by terminating RT to the reference output and providing a saw tooth input to CT, or it can be used to drive the common circuitry in synchronous multiple-rail power supplies.

The uncommitted output transistors provide either common-emitter or emitter-follower output capability. Each device provides for push-pull or single-ended output operation, with selection by means of the output-control
Function.

The architecture of these devices prohibits the possibility of either output being pulsed twice during push-pull operation. The under voltage lockout control circuit locks the outputs off until the internal circuitry is operational.