NOVEL TECHNIQUE FOR IMPROVING POWER QUALITY WITH VOLTAGE COMPENSATION AND HARMONIC SUPPRESSION

A Dissertation
By

Gupta Vinodkumar Rameshchandra

Submitted to

सत्यं शिवं सुन्दरम्

Department of Electrical Engineering
Faculty of Technology & Engineering
The Maharaja Sayajirao University of Baroda
Vadodara, Gujarat

in partial fulfillment of the requirements
for Degree of

Doctor of Philosophy

November 2011

Major Subject: Electrical Engineering

NOVEL TECHNIQUE FOR IMPROVING POWER QUALITY WITH VOLTAGE COMPENSATION AND HARMONIC SUPPRESSION

A Dissertation

By

Gupta Vinodkumar Rameshchandra

Submitted to

Department of Electrical Engineering

Faculty of Technology & Engineering

The Maharaja Sayajirao University of Baroda

Vadodara, Gujarat

in partial fulfillment of the requirements for Degree of Doctor of Philosophy

Guided By:

Dr. R. B. Kelker, Professor
PhD (Electrical), MIE, MISTE
Department of Electrical Engineering
Faculty of Technology and Engineering
M.S. University
Vadodara

Dr. M. Ramamoorty

Head of Department

Faculty of Technology & Engineering, M. S. University, Vadodara

Dean

Faculty of Technology & Engineering, M. S. University, Vadodara

DECLARATION

I, **Gupta Vinodkumar Rameshchandra** hereby declare that the work reported in the thesis titled "**NOVEL TECHNIQUE FOR IMPROVING POWER QUALITY WITH VOLTAGE COMPENSATION AND HARMONIC SUPPRESSION**", submitted for the award of degree of **DOCTOR OF PHILOSOPHY** in Electrical Engineering is original and was done by me in the "**The Department of Electrical Engineering, Faculty of Technology and Engineering, The Maharaja Sayajirao University of Baroda, Vadodara**". I take the privilege in declaring that the work is developed, structured and prepared by me solely. Further I declare that work is not part of any declared or published work partly or fully for the award of any degree or academic qualifications for this university or any other institutions of examining body in India and abroad.

Place:

Date:

Gupta Vinodkumar Rameshchandra

ABSTRACT

New technology has driven the development of ever more flexible, cost effective, high performance electrical equipments. However, such increasingly sophisticated equipments inject non-sinusoidal current, which distort the power supply voltage. The harmonics, which cause the voltage distortion, lead to malfunctioning, abnormal heating and vibrations in equipment connected to network.

Also Use of 3-ϕ induction motor (IM) in various industries is almost 80% of their total drives requirements. Performance of these induction motors is affected by the unbalance in input voltages due to the flow of negative sequence current, which heats up the rotor and generates breaking torque. Due to the very low value of negative sequence impedance offered by the induction motor as compared to its positive sequence impedance, a very small negative sequence voltage component in the input may give rise to considerable negative sequence current causing deterioration to a great extent in the performance of induction motor. Therefore even low unbalance factor (ratio of negative sequence to positive sequence) in input voltage has to be carefully looked into. Voltage unbalances occur quite often in distribution systems. Fig. 1 shows the schematic of this conventional induction motor drive and the effect of input voltage unbalance.

This research work presents a three-phase series active filter to eliminate voltage harmonics in distribution system along with negative sequence & zero sequence compensator. It is based on instantaneous active reactive power theory (p-q theory). A three phase IGBT based voltage source inverter with DC bus capacitor is used as three-phase series active filter and negative & zero sequence compensator. Reference voltages and distorted voltages are derived through PTs from three-phase system for the computation of desired compensation voltages.

It is well known that the average power flow due to positive sequence voltage and negative sequence current is zero

$$
\begin{aligned}
&(V_m \sin(\omega t) I_m \sin(\omega t) + \\
&(V_m \sin(\omega t - 120) I_m \sin(\omega t + 120)) \\
&+ (V_m \sin(\omega t + 120) I_m \sin(\omega t - 120)) = 0
\end{aligned}
$$

Similarly the negative sequence voltage will generate nonzero power with negative sequence current.

Hence this concept is used for generating the reference signal for negative sequence compensator using instantaneous α-β-0 theory.

Finally using Instantaneous Active Reactive Power Theory the compensating signal was derived, which can compensate the voltage due to negative sequence, zero sequence, & harmonics present. This is the contribution of this thesis.

The three-phase, three legs IGBT based voltage source inverter with DC bus capacitor regulates the compensating voltages in the close vicinity of the desired reference voltages. The performance characteristics of the active filter is simulated and tested experimentally. The experimental results established the superiority of the series type active filter over passive tuned filter.

To my parents

for their sacrifice, trust, and for all the beautiful things they taught me.

To my beloved family:

Sangita

Hetavi and Bhavya

who are always deep in my heart.

AKNOWLEDGEMENT

I wish to express my appreciation & deep thanks to my research guide Dr R. B. Kelker, Professor Electrical Engineering Department, Faculty of Technology and Engineering, M.S. University; Vadodara, who inspired me to work in power quality improvement, for his timely guidance, suggestions without whom this success would not have been possible.

With a deep sense of gratitude, I acknowledge here, the motivation, guidance & inspiration I received from Dr M. Ramamoorty who shows full confidence in my abilities and encouraging me to perform to the best of my capabilities. It is only his inspiration to introduce me to front end research & guide me throughout this research work.

My sincere thanks to Shri U. C. Trivedi, Joint Director & Head, Electrical & Electronics Division, Electrical Research and Development Association, (ERDA), Vadodara for extending all the best possible help throughout the course of this work

With deep sense of gratitude I thank Shri D. A. Sathe, General Manager Engineering Department, Tata Power Co Ltd. Mumbai for sponsoring & taking the initiating to solve the issue related to power quality of their system. I am also grateful to Shri A. M. Mukherji & his colleagues along with my colleague Shri Kamlesh Keharia for his co-operation during the field-testing of the series active filter.

I am grateful to Management of ERDA for providing me the facilities and infrastructure to carry out the work at ERDA, Vadodara. I gratefully acknowledge the assistance received from my colleague Shri Kamlesh Keharia and Dinesh Rohit along with others colleagues of Power Electronics & Instrumentation cell who truly helped me throughout the period.

Without my family's sacrifice and support, this study would not have been possible. I would like to thank my parents, my beloved wife, my daughters Hetavi and my son Bhavya for their unconditional love and constant encouragement in all my endeavors.

My grateful thanks are due to many others who have made it possible for me to complete this work

Vinod Gupta

TABLE OF CONTENTS

List of Figures

Chapter III

List of Tables

CHAPTER-I

OVERVIEW

TABLE OF CONTENTS

1.1 INTRODUCTION

With modern era of information technology & with comfort of living, more and more power electronics equipment are used in our life. Most of the power electronics equipment are nonlinear load which creating the power quality issue even though it is known that with advancement in technology these problem will be solved by power electronics control technology. Because of power quality problem the performance of the other equipment connected in the same bus is affected. Moreover electrical quantities are also affected because of power quality issues. This is explained in subsequent section

Both electric utilities and end users of electric power are becoming increasingly concerned about the quality of electric power. The term *power quality* has become one of the most prolific buzzwords in the power industry since the late 1980s. It is an umbrella concept for a multitude of individual types of power system disturbances. The issues that fall under this umbrella are not necessarily new. What is new is that engineers are now attempting to deal with these issues using a system approach rather than handling them as individual problems. There are four major reasons for the increased concern:

1. Newer-generation load equipment, with microprocessor-based controls and power electronic devices, is more sensitive to power quality variations than was equipment used in the past.

2. The increasing emphasis on overall power system efficiency has resulted in continued growth in the application of devices such as high-efficiency, adjustable-speed motor drives and shunt capacitors for power factor correction to reduce losses. This is resulting in increasing harmonic levels on power systems and has many people concerned about the future impact on system capabilities.

3. End users have an increased awareness of power quality issues. Utility customers are becoming better informed about such issues as interruptions, sags and switching transients and are challenging the utilities to improve the quality of power delivered.

4. Many things are now interconnected in a network. Integrated processes mean that the failure of any component has much more important consequences.

The common thread running through all these reasons for increased concern about the quality of electric power is the continued push for increasing productivity for all utility customers. Manufacturers want faster, more productive, more efficient machinery. Utilities encourage this effort because it helps their customers become more profitable and also helps defer large investments in substations and generation by using more efficient load equipment. Interestingly, the equipment installed to increase the productivity is also often the equipment that suffers the most from common power disruptions. And the equipment is sometimes the source of additional power quality problems. When entire processes are automated, the efficient operation of machines and their controls becomes increasingly dependent on quality power.

1. Throughout the world, many governments have revised their laws regulating electric utilities with the intent of achieving more cost-competitive sources of electric energy. Deregulation of utilities has complicated the power quality problem. In many geographic areas there is no longer tightly coordinated control of the power from generation through end-use load. While regulatory agencies can change the laws regarding the flow of money, the physical laws of power flow cannot be altered. In order to avoid deterioration of the quality of power supplied to customers, regulators are going to have to expand their thinking beyond traditional reliability indices

and address the need for power quality reporting and incentives for the transmission and distribution companies.

2. There has been a substantial increase of interest in distributed generation (DG), that is, generation of power dispersed throughout the power system. There are a number of important power quality issues that must be addressed as part of the overall interconnection evaluation for DG.

3. The globalization of industry has heightened awareness of deficiencies in power quality around the world. Companies building factories in new areas are suddenly faced with unanticipated problems with the electricity supply due to weaker systems or a different climate. There have been several efforts to benchmark power quality in one part of the world against other areas.

4. Indices have been developed to help benchmark the various aspects of power quality. Regulatory agencies have become involved in performance-based rate-making (PBR), which addresses a particular aspect, reliability, which is associated with interruptions. Some customers have established contracts with utilities for meeting a certain quality of power delivery.

1.2 WHAT IS POWER QUALITY?

There can be completely different definitions for power quality, depending on one's frame of reference. For example, a utility may define power quality as reliability and show statistics demonstrating that its system is 99.98 percent reliable. Criteria established by regulatory agencies are usually in this vein. A manufacturer of load equipment may define power quality as those characteristics of the power supply that enable the equipment to work properly. These characteristics can be very different for different criteria.

Power quality is ultimately a consumer-driven issue, and the end user's point of reference takes precedence. Therefore, the following definition of a power quality problem is used in this thesis:

"Any power problem manifested in voltage, current, or frequency deviations that result in failure or mis-operation of customer equipment."

There are many misunderstandings regarding the causes of power quality problems. The charts in Figure 1.2-1 show the results of one survey conducted by the Georgia Power Company in which both utility personnel and customers were polled about what causes power quality problems. While surveys of other market sectors might indicate different splits between the categories, these charts clearly illustrate one common theme that arises repeatedly in such surveys: The utility's and customer's perspectives are often much different. While both tend to blame about two-thirds of the events on natural phenomena (e.g., lightning), customers, much more frequently than utility personnel, think that the utility is at fault. When there is a power problem with a piece of equipment, end users may be quick to complain to the utility of an "outage" or "glitch" that has caused the problem. However, the utility records may indicate no abnormal events on the feed to the customer. It must be realized that there are many events resulting in end-user problems that never show up in the utility statistics. One example is capacitor switching, which is quite common and normal on the utility system, but can cause transient overvoltage that disrupt manufacturing machinery. Another example is a momentary fault elsewhere in the system that causes the voltage to sag briefly at the location of the customer in question. This might because an adjustable-speed drive or a distributed generator to trip off, but the utility will have no indication that anything was amiss on the feeder unless it has a power quality monitor installed.

In addition to real power quality problems, there are also perceived power quality problems that may actually be related to hardware, software, or control system malfunctions. Electronic components can degrade over time due to repeated transient voltages and eventually fail due to a relatively low magnitude event. Thus, it is sometimes difficult to associate a failure with a specific cause. It is becoming more common that designers of control software for microprocessor-based equipment have an incomplete knowledge of how power systems operate and do not anticipate all types of malfunction events. Thus, a device can misbehave because of a deficiency in the embedded software. This is particularly common with early versions of new computer-controlled load equipment.

In response to this growing concern for power quality, electric utilities have programs that help them respond to customer concerns. The philosophy of these programs ranges from reactive, where the utility responds to customer complaints, to proactive, where the utility is involved in educating the customer and promoting services that can help to develop solutions to power quality problems. The regulatory issues facing utilities may play an important role in how their programs are structured. Since power quality problems often involve interactions between the supply system and the customer facility and equipment, regulators should make sure that distribution companies have incentives to work with customers and help customers solve these problems.

The economics involved in solving a power quality problem must also be included in the analysis. It is not always economical to eliminate power quality variations on the supply side. In many cases, the optimal solution to a problem may involve making a particular piece of sensitive equipment less sensitive to power quality variations. The level of power quality required is that level which will result in proper operation of the equipment at a particular facility.

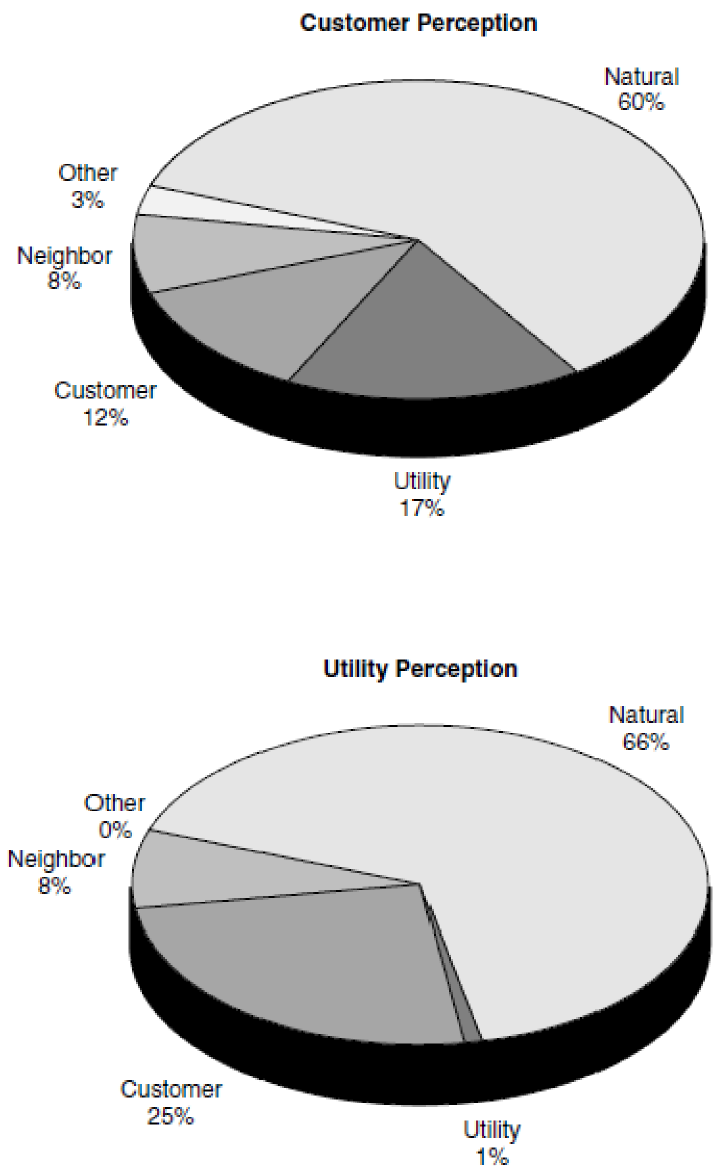


Figure 1.2-1: Results of a survey on the causes of power quality problems (Courtesy of Georgia Power Co.).

Power quality, like quality in other goods and services, is difficult to quantify. There is no single accepted definition of quality power. There are standards for voltage and other technical criteria that may be measured, but the ultimate measure of power quality is determined by the performance and productivity of end-user equipment. If the electric power is inadequate for those needs, then the “quality” is lacking. Perhaps nothing has been more symbolic of a mismatch in the power delivery system and consumer technology than the “blinking clock” phenomenon. Clock designers created the blinking display of a digital clock to warn of possible incorrect time after loss of power and inadvertently created one of the first power quality monitors. It has made the homeowner aware that there are numerous minor disturbances occurring throughout the power delivery system that may have no ill effects other than to be detected by a clock. Many appliances now have a built-in clock, so the average household may have about a dozen clocks that must be reset when there is a brief interruption. Older-technology motor-driven clocks would simply lose a few seconds during minor disturbances and then promptly come back into synchronism.

1.3 POWER QUALITY = VOLTAGE QUALITY

In most of the cases the terms power quality is nothing but it is actually the quality of the voltage that is being addressed in most cases. Power is the rate of energy delivery and is proportional to the product of the voltage and current. It would be difficult to define the quality of this quantity in any meaningful manner. The power supply system can only control the quality of the voltage; it has no control over the currents that particular loads might draw. Therefore, the standards in the power quality area are devoted to maintaining the supply voltage within certain limits. AC power systems are designed to operate at a sinusoidal voltage of a given frequency [typically 50 or 60 hertz (Hz)] and magnitude. Any significant deviation in the waveform magnitude, frequency, or purity is a potential power quality problem. Of course, there is always a close relationship between voltage and current in any practical power system. Although the generators may provide a near-perfect sine-wave voltage, the current passing through the impedance of the system can cause a variety of disturbances to the voltage. For example,

1. *The current resulting from a short circuit causes the voltage to sag or disappear completely, as the case may be.*
2. *Currents from lightning strokes passing through the power system cause high-impulse voltages that frequently flash over insulation and lead to other phenomena, such as short circuits.*
3. *Distorted currents from harmonic-producing loads also distort the voltage as they pass through the system impedance. Thus a distorted voltage is presented to other end users. Therefore, ultimately the voltage is more concerned. Hence the phenomena in the current to understand the basis of many power quality problems must be addressed.*

1.4 POWER SYSTEM QUANTITIES UNDER NONSINUSOIDAL CONDITIONS

Traditional power system quantities such as rms, power (reactive, active, apparent), power factor, and phase sequences are defined for the fundamental frequency context in a pure sinusoidal condition. In the presence of harmonic distortion the power system no longer operates in a sinusoidal condition, and unfortunately many of the simplifications power engineers use for the fundamental frequency analysis do not apply.

1.4.1 ACTIVE, REACTIVE, AND APPARENT POWER

There are three standard quantities associated with power:

■ *Apparent power S* [*voltampere* (VA)]. The product of the rms voltage and current.

■ *Active power P* [*watt* (W)]. The average rate of delivery of energy.

■ *Reactive power Q* [*voltampere-reactive*] (VAr)]. The portion of the apparent power that is out of phase, or in quadrature, with the active power.

The apparent power *S* applies to both sinusoidal and nonsinusoidal conditions. The apparent power can be written as follows:

$$S = V_{rms} \, X \, I_{rms} \qquad (1.4.1\text{-}1)$$

where V_{rms} and I_{rms} are the rms values of the voltage and current. In a sinusoidal condition both the voltage and current waveforms contain only the fundamental frequency component; thus the rms values can be expressed simply as

$$V_{rms} = \frac{1}{\sqrt{2}} V_1 \quad and \quad I_{rms} = \frac{1}{\sqrt{2}} I_1 \qquad (1.4.1\text{-}2)$$

where V_1 and I_1 are the amplitude of voltage and current waveforms, respectively. The subscript “1” denotes quantities in the fundamental frequency. In a nonsinusoidal condition a harmonically distorted waveform is made up of sinusoids of harmonic frequencies with different amplitudes as shown in Figure 1.4.1-1. The rms values of the waveforms are computed as the square root of the sum of rms squares of all individual components, i.e.

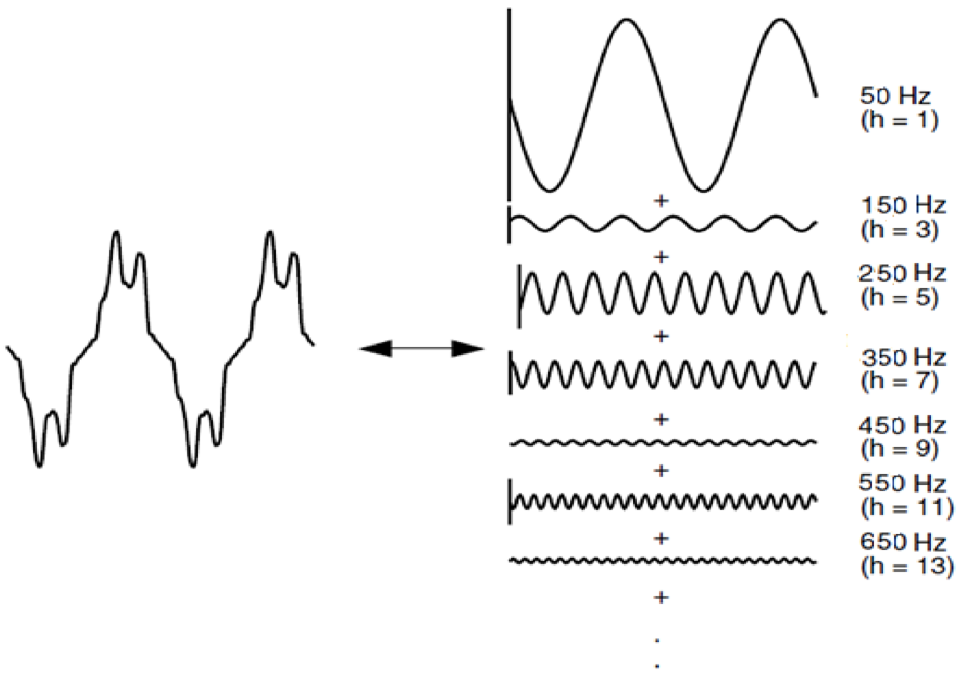


Figure 1.4.1-1: Fourier series representation of distorted waveform

$$V_{rms} = \sqrt{\sum_{h=1}^{h_{max}} \left(\frac{1}{\sqrt{2}} V_h\right)^2} = \frac{1}{\sqrt{2}} \sqrt{V_1^2 + V_2^2 + V_3^2 + \ldots + V_{h_{max}}^2} \quad (1.4.1\text{-}3)$$

$$I_{rms} = \sqrt{\sum_{h=1}^{h_{max}} \left(\frac{1}{\sqrt{2}} I_h\right)^2} = \frac{1}{\sqrt{2}} \sqrt{I_1^2 + I_2^2 + I_3^2 + \ldots + I_{h_{max}}^2} \quad (1.4.1\text{-}4)$$

where V_h and I_h are the amplitude of a waveform at the harmonic component *h*. In the sinusoidal condition, harmonic components of V_h and I_h are all zero, and only $V1$ and $I1$ remain. Equations (1.4.1-3) and (1.4.1-4) simplify to Equation (1.4.1-2). The active power *P* is also commonly referred to as the average power, real power, or true power. It represents useful power expended by loads to perform real work, i.e., to convert electric energy to other forms of energy. Real work performed by an incandescent light bulb is to convert electric energy into light and heat. In electric power, real work is performed for the portion of the current that is in phase with the voltage. No real work will result from the portion where the current is not in phase with the voltage. The active power is the rate at which energy is expended, dissipated, or consumed by the load and is measured in units of watts. *P* can be computed by averaging the product of the instantaneous voltage and current, i.e.

$$P = \frac{1}{T}\int_0^T v(t)\, i(t)\, dt \qquad (1.4.1\text{-}5)$$

Equation (1.4.1-5) is valid for both sinusoidal and nonsinusoidal conditions.

For the sinusoidal condition, *P* resolves to the familiar form,

$$P = \frac{V_1 I_1}{2}\cos\theta_1 \; = \; V_{1rms}\, I_{1rms}\, \cos\theta_1 = S\, \cos\theta_1 \qquad (1.4.1\text{-}6)$$

where θ_1 is the phase angle between voltage and current at the fundamental frequency. Equation (1.4.1-6) indicates that the average active power is a function only of the fundamental frequency quantities. In the nonsinusoidal case, the computation of the active power must include contributions from all harmonic components; thus it is the sum of active power at each harmonic. Furthermore, because the voltage distortion is generally very low on power systems (less than 5 percent), Equation (1.4.1-6) is a good approximation regardless of how distorted the current is. This approximation cannot be applied when computing the apparent and reactive power. These two quantities are greatly influenced by the distortion. The apparent power *S* is a measure of the potential impact of the load on the thermal capability of the system. It is proportional to the rms of the distorted current, and its computation is straightforward, although slightly more complicated than the sinusoidal case. Also, many current probes can now directly report the true rms value of a distorted waveform. The reactive power is a type of power that does no real work and is generally associated with reactive elements (inductors and capacitors).

For example, the inductance of a load such as a motor causes the load current to lag behind the voltage. Power appearing across the inductance sloshes back and forth between the inductance itself and the power system source, producing no net effective work. For this reason it is called imaginary or reactive power since no power is dissipated or expended. It is expressed in units of vars. In the sinusoidal case, the reactive power is simply defined as

$$Q = \frac{V_1 I_1}{2} \sin\theta_1 \ = V_{1rms}\, I_{1rms}\ \sin\theta_1 = S\ \sin\theta_1 \qquad (1.4.1\text{-}7)$$

Which is the portion of power in quadrature with the active power shown in Equation (1.4.1-6). Figure 1.4.1-2 summarizes the relationship between *P, Q,* and *S* in sinusoidal condition.

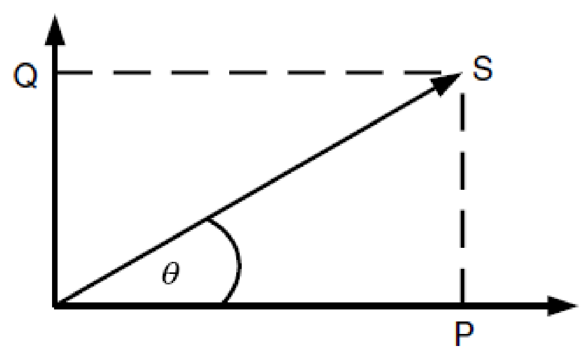


Figure 1.4.1-2: Relationship between P, Q and S in sinusoidal condition

There is some disagreement among harmonics analysts on how to define *Q* in the presence of harmonic distortion. If it were not for the fact that many utilities measure *Q* and compute demand billing from the power factor computed by *Q,* it might be a moot point. It is more important to determine *P* and *S*; *P* defines how much active power is being consumed, while *S* defines the capacity of the power system required to deliver *P. Q* is not actually very useful by itself. However, $Q1$, the traditional reactive power component at fundamental frequency, may be used to size shunt capacitors.

The reactive power when distortion is present has another interesting peculiarity. In fact, it may not be appropriate to call it reactive *power.* The concept of VAr flow in the power system is deeply ingrained in the minds of most power engineers. What many do not realize is that this concept is valid only in the sinusoidal steady state. When distortion is present, the component of S that remains after P is taken out is not conserved—that is, it does not sum to zero at a node. Power quantities are presumed to flow around the system in a conservative manner.

This does not imply that P is not conserved or that current is not conserved because the conservation of energy and Kirchhoff's current laws are still applicable for any waveform. The reactive components actually sum in quadrature (square root of the sum of the squares). This has prompted some analysts to propose that Q be used to denote the reactive components that are conserved and introduce a new quantity for the components that are not. Many call this quantity *D,* for *distortion power* or, simply, *distortion voltamperes.* It has units of voltamperes, but it may not be strictly appropriate to refer to this quantity as *power,* because it does not flow through the system as power is assumed to do. In this concept, Q consists of the sum of the traditional reactive power values at each frequency. D represents all cross products of voltage and current at different frequencies, which yield "NO AVERAGE POWER". *P, Q, D,* and S are related as follows, using the definitions for S and P previously given in Equation (1.4.1-1) and Equation (1.4.1-5) as a starting point:

$$S = \sqrt{P^2 + Q^2 + D^2} \tag{1.4.1-8-a}$$

$$Q = \sum_k V_k I_k \sin\theta_k \tag{1.4.1-8-b}$$

Therefore, *D* can be determined after *S, P,* and *Q* by

$$D = \sqrt{S^2 - P^2 - Q^2} \qquad (1.4.1\text{-}9)$$

Some prefer to use a three-dimensional vector chart to demonstrate the relationships of the components as shown in Figure 1.4.2-1. *P* and *Q* contribute the traditional sinusoidal components to *S,* while *D* represents the additional contribution to the apparent power by the harmonics.

1.4.2 POWER FACTOR: DISPLACEMENT AND TRUE

Power factor (PF) is a ratio of useful power to perform real work (active power) to the power supplied by a utility (apparent power), i.e.

$$\mathrm{PF} = \frac{P}{S} \qquad (1.4.2\text{-}1)$$

In other words, the power factor ratio measures the percentage of power expended for its intended use. Power factor ranges from zero to unity. A load with a power factor of 0.9 lagging denotes that the load can effectively expend 90 percent of the apparent power supplied (voltamperes) and convert it to perform useful work (watts). The term *lagging* denotes that the fundamental current lags behind the fundamental voltage by 25.84°. In the sinusoidal case there is only one phase angle between the voltage and the current (since only the fundamental frequency is present; the power factor can be computed as the cosine of the phase angle and is commonly referred as the *displacement power factor:*

$$\mathrm{PF} = \frac{P}{S} = \cos\theta \qquad (1.4.2\text{-}2)$$

In the nonsinusoidal case the power factor cannot be defined as the cosine of the phase angle as in Equation (1.4.2-2). The power factor that takes into account the contribution from all active power, including both fundamental and harmonic frequencies, is known as the *true power factor.* The true power factor is simply the ratio of total active power for all frequencies to the apparent power delivered by the utility as shown in Equation (1.4.2-1).

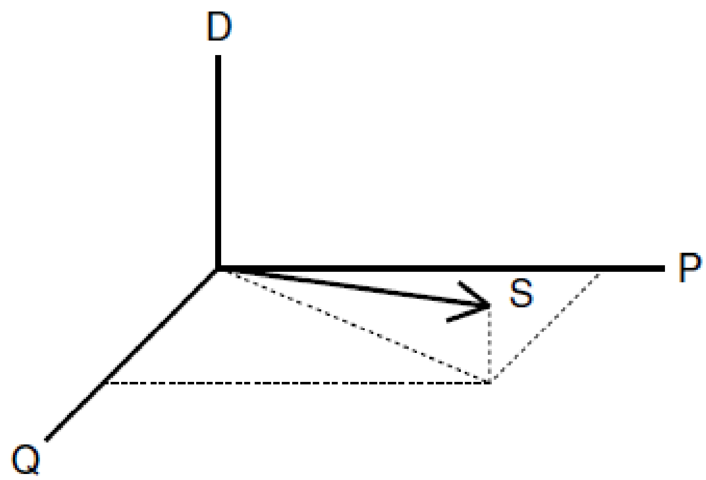


Figure 1.4.2-1: Relationship of components of the apparent power

Power quality monitoring instruments now commonly report both displacement and true power factors. Many devices such as switch- mode power supplies and PWM adjustable-speed drives have a near unity displacement power factor, but the true power factor may be 0.5 to 0.6. An ac-side capacitor will do little to improve the true power factor in this case because $Q1$ is zero.

In fact, if it results in resonance, the distortion may increase, causing the power factor to degrade. The true power factor indicates how large the power delivery system must be built to supply a given load. In this example, using only the displacement power factor would give a false sense of security that all is well. The bottom line is that distortion results in additional current components flowing in the system that do not yield any net energy except that they cause losses in the power system elements they pass through. This requires the system to be built to a slightly larger capacity to deliver the power to the load than if no distortion were present.

1.4.3 HARMONIC PHASE SEQUENCES

Power engineers have traditionally used symmetrical components to help describe three-phase system behavior. The three-phase system is transformed into three single-phase systems that are much simpler to analyze. The method of symmetrical components can be employed for analysis of the system's response to harmonic currents provided care is taken not to violate the fundamental assumptions of the method. The method allows any unbalanced set of phase currents (or voltages) to be transformed into three balanced sets. The *positive-sequence* set contains three sinusoids displaced 120° from each other, with the normal A-B-C phase rotation (e.g., 0°, -120°, 120°). The sinusoids of the *negative-sequence* set are also displaced 120°, but have opposite phase rotation (A-C-B, e.g., 0°, 120°, -120°). The sinusoids of the *zero sequence* are in phase with each other (e.g., 0°, 0°, 0°).

In a perfect balanced three-phase system, the harmonic phase sequence can be determined by multiplying the harmonic number h with the normal positive-sequence phase rotation. For example, for the second harmonic, h=2, hence 2 is multiplied with normal phase rotation i.e. 2 x (0, -120°, 120°) or (0°, 120°, -120°), which is the negative sequence. For the third harmonic, h = 3, hence 3 is multiplied with normal phase rotation i.e. 3 x (0°, -120°, 120°) or (0°, 0°, 0°), which is the zero sequence. Phase sequences for all other harmonic orders can be determined in the same fashion. Since a distorted waveform in power systems contains only odd-harmonic components (see Sec. 5.1), only odd-harmonic phase sequence rotations are summarized here:

■ Harmonics of order h = 1, 7, 13,… are generally positive sequence.

■ Harmonics of order h = 5, 11, 17,… are generally negative sequence.

■ Triplens (h = 3, 9, 15,…) are generally zero sequence. Impacts of sequence harmonics on various power system components are detailed in Sec. 1.9.

1.5 HARMONIC INDICES

The two most commonly used indices for measuring the harmonic content of a waveform are the total harmonic distortion and the total demand distortion. Both are measures of the effective value of a waveform and may be applied to either voltage or current.

1.5.1 TOTAL HARMONIC DISTORTION

The THD is a measure of the *effective value* of the harmonic components of a distorted waveform. That is, it is the potential heating value of the harmonics relative to the fundamental. This index can be calculated for either voltage or current:

$$THD = \frac{\sqrt{\sum_{h>1}^{h_{max}} M_h^2}}{M_1} \tag{1.5.1-1}$$

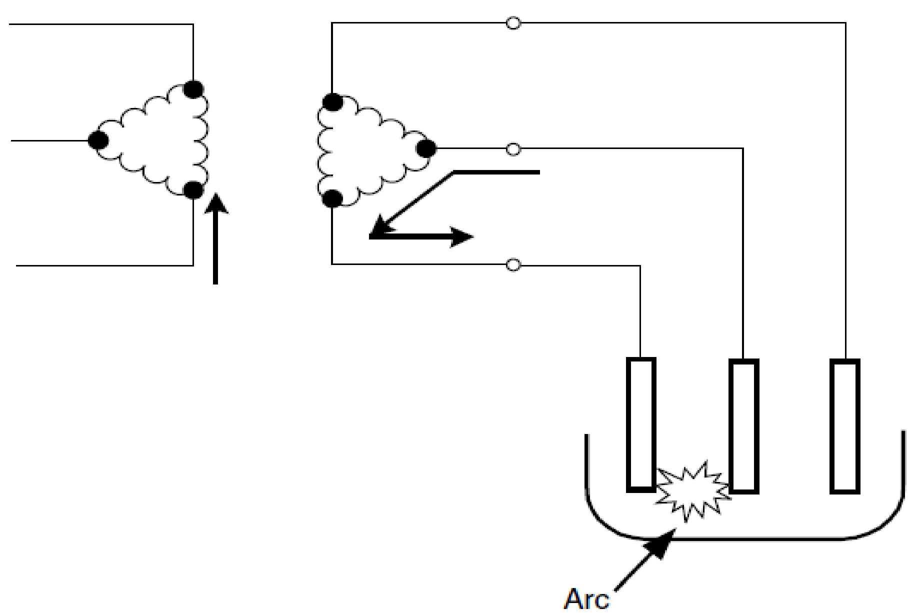


Figure 1.5.1-1: Arc furnace operation in an unbalanced mode allows triplen harmonics to reach the power system despite a delta connected transformer.

where M_h is the rms value of harmonic component h of the quantity M. The rms value of a distorted waveform is the square root of the sum of the squares as shown in Equation (1.4.1-3) and (1.4.1-4). The THD is related to the rms value of the waveform as follows:

$$RMS = \sqrt{\sum_{h=1}^{h_{max}} M_h^2} = M_1 \sqrt{1 + THD^2} \tag{1.5.1-2}$$

The THD is a very useful quantity for many applications, but its limitations must be realized. It can provide a good idea of how much extra heat will be realized when a distorted voltage is applied across a resistive load. Likewise, it can give an indication of the additional losses caused by the current flowing through a conductor. However, it is not a good indicator of the voltage stress within a capacitor because that is related to the peak value of the voltage waveform, not its heating value.

The THD index is most often used to describe voltage harmonic distortion. Harmonic voltages are almost always referenced to the fundamental value of the waveform at the time of the sample. Because fundamental voltage varies by only a few percent, the voltage THD is nearly always a meaningful number. Variations in the THD over a period of time often follow a distinct pattern representing nonlinear load activities in the system.

1.5.2 TOTAL DEMAND DISTORTION

Current distortion levels can be characterized by a THD value, as has been described, but this can often be misleading. A small current may have a high THD but not be a significant threat to the system. For example, many adjustable-speed drives will exhibit high THD values for the input current when they are operating at very light loads. This is not necessarily a significant concern because the magnitude of harmonic current is low, even though its relative current distortion is high. Some analysts have attempted to avoid this difficulty by referring THD to the fundamental of the peak demand load current rather than the fundamental of the present sample. This is called total demand distortion and serves as the basis for the guidelines in IEEE Standard 519-1992, *Recommended Practices and Requirements for Harmonic Control in Electrical Power Systems.* It is defined as follows:

$$TDD = \frac{\sqrt{\sum_{h=2}^{h_{max}} I_h^2}}{I_L} \tag{1.5.2-1}$$

IL is the peak, or maximum, demand load current at the fundamental frequency component measured at the point of common coupling (PCC). There are two ways to measure *IL.* With a load already in the system, it can be calculated as the average of the maximum demand current for the preceding 12 months. The calculation can simply be done by averaging the 12-month peak demand readings. For a new facility, *IL* has to be estimated based on the predicted load profiles.

1.6 HARMONIC SOURCES FROM COMMERCIAL LOADS

Commercial facilities such as office complexes, department stores, hospitals, and Internet data centers are dominated with high-efficiency fluorescent lighting with electronic ballasts, adjustable-speed drives for the heating, ventilation, and air conditioning (HVAC) loads, elevator drives, and sensitive electronic equipment supplied by single-phase switch-mode power supplies. Commercial loads are characterized by a large number of small harmonic-producing loads. Depending on the diversity of the different load types, these small harmonic currents may add in phase or cancel each other. The voltage distortion levels depend on both the circuit impedances and the overall harmonic current distortion. Since power factor correction capacitors are not typically used in commercial facilities, the circuit impedance is dominated by the service entrance transformers and conductor impedances. Therefore, the voltage distortion can be estimated simply by multiplying the current by the impedance adjusted for frequency. Characteristics of typical nonlinear commercial loads are detailed in the following sections.

1.6.1 SINGLE-PHASE POWER SUPPLIES

Electronic power converter loads with their capacity for producing harmonic currents now constitute the most important class of nonlinear loads in the power system. Advances in semiconductor device technology have fueled a revolution in power electronics over the past decade, and there is every indication that this trend will continue. Equipment includes adjustable-speed motor drives, electronic power supplies, dc motor drives, battery chargers, electronic ballasts, and many other rectifier and inverter applications.

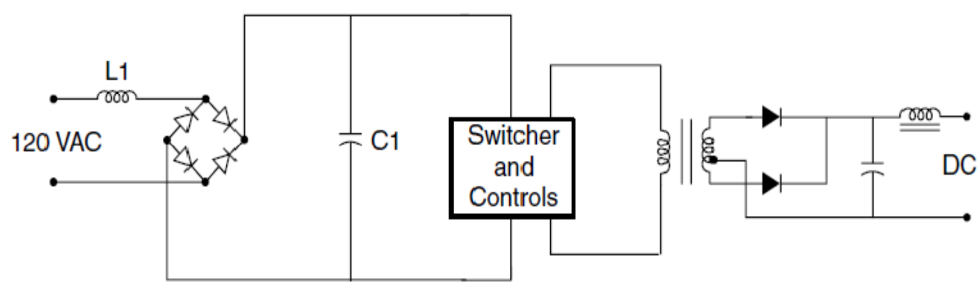


Figure 1.6.1-1: Switch-mode power supply.

A major concern in commercial buildings is that power supplies for single-phase electronic equipment will produce too much harmonic current for the wiring. DC power for modern electronic and microprocessor based office equipment is commonly derived from single-phase full-wave diode bridge rectifiers. The percentage of load that contains electronic power supplies is increasing at a dramatic pace, with the increased utilization of personal computers in every commercial sector. There are two common types of single-phase power supplies. Older technologies use ac-side voltage control methods, such as transformers, to reduce voltages to the level required for the dc bus. The inductance of the transformer provides a beneficial side effect by smoothing the input current waveform, reducing harmonic content.

Newer-technology switch-mode power supplies (see Figure 1.6.1-1) use dc-to-dc conversion techniques to achieve a smooth dc output with small, lightweight components.

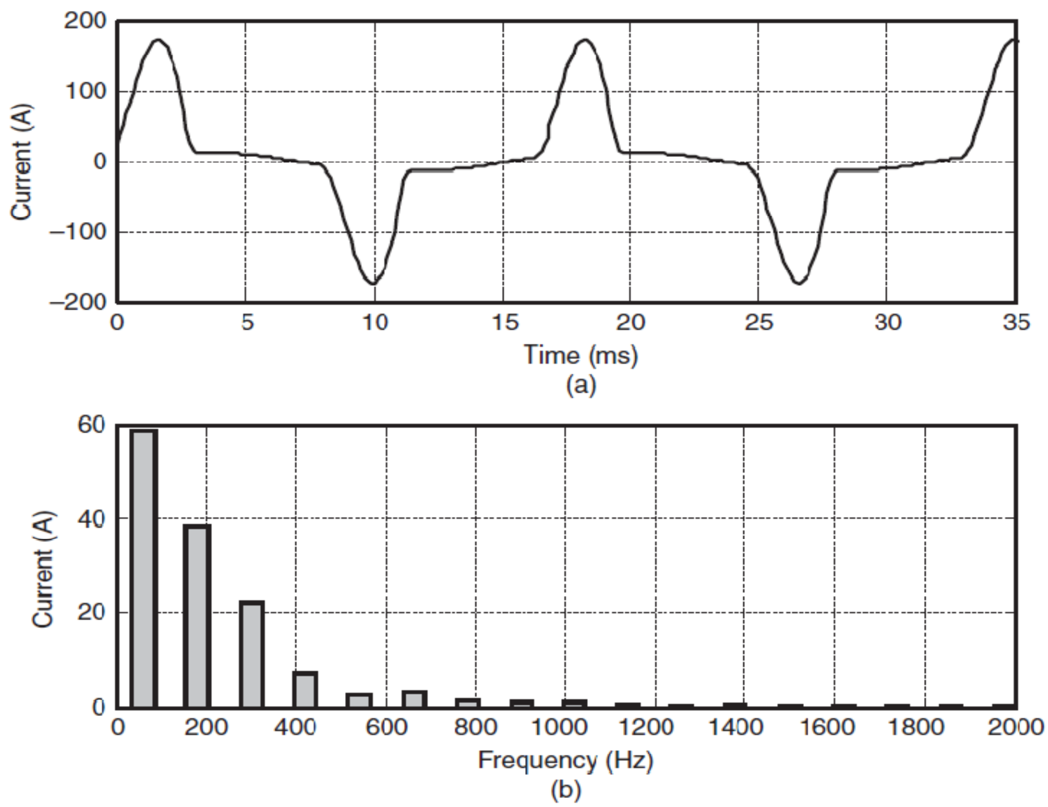


Figure 1.6.1-2: SMPS current and harmonic spectrum.

The input diode bridge is directly connected to the ac line, eliminating the transformer. This results in a coarsely regulated dc voltage on the capacitor. This direct current is then converted back to alternating current at a very high frequency by the switcher and subsequently rectified again. Personal computers, printers, copiers, and most other single-phase electronic equipment now almost universally employ switch-mode power supplies. The key advantages are the light weight, compact size, efficient operation, and lack of need for a transformer. Switch-mode power supplies can usually tolerate large variations in input voltage. Because there is no large ac-side inductance, the input current to the power supply comes in very short pulses as the capacitor $C1$ regains its charge on each half cycle. Figure 1.6.1-2 illustrates the current waveform and spectrum for an entire circuit supplying a variety of electronic equipment with switch-mode power supplies. A distinctive characteristic of switch-mode power supplies is a very high third-harmonic content in the current. Since third-harmonic current components are additive in the neutral of a three-phase system, the increasing application of switch-mode power supplies causes concern for overloading of neutral conductors, especially in older buildings where an undersized neutral may have been installed. There is also a concern for transformer overheating due to a combination of harmonic content of the current, stray flux, and high neutral currents.

1.6.2 FLUORESCENT LIGHTING

Lighting typically accounts for 40 to 60 percent of a commercial building load. According to the 1995 Commercial Buildings Energy Consumption study conducted by the U.S. Energy Information Administration, fluorescent lighting was used on 77 percent of commercial floor spaces; while only 14 percent of the spaces used incandescent lighting.1 Fluorescent lights are a popular choice for energy savings.

Fluorescent lights are discharge lamps; thus they require a ballast to provide a high initial voltage to initiate the discharge for the electric current to flow between two electrodes in the fluorescent tube. Once the discharge is established, the voltage decreases as the arc current increases. It is essentially a short circuit between the two electrodes, and the ballast has to quickly reduce the current to a level to maintain the specified lumen output. Thus, ballast is also a current-limiting device in lighting applications.

There are two types of ballasts, magnetic and electronic. Standard magnetic ballast is simply made up of an iron-core transformer with a capacitor encased in an insulating material. Single magnetic ballast can drive one or two fluorescent lamps, and it operates at the line fundamental frequency, i.e., 50 or 60 Hz. The iron-core magnetic ballast contributes additional heat losses, which makes it inefficient compared to electronic ballast.

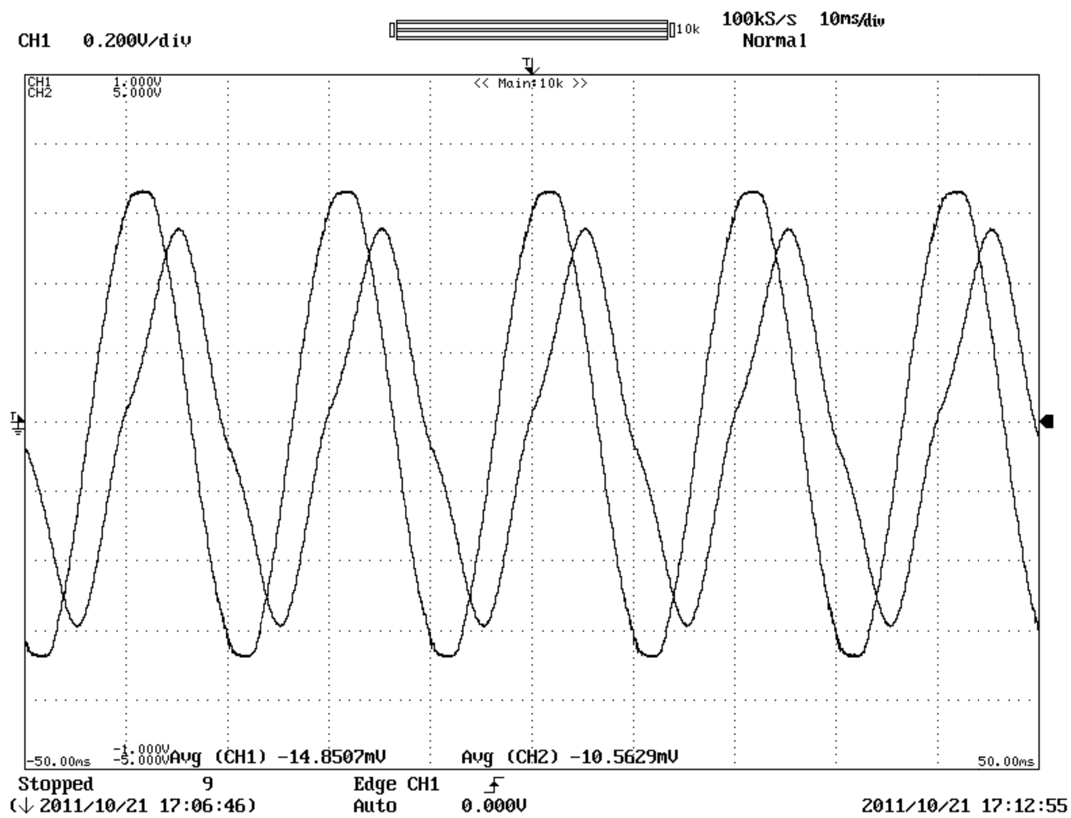


Figure 1.6.2-1: Input current & voltage waveform for fluorescent lamp with magnetic ballast

Electronic ballast employs a switch-mode–type power supply to convert the incoming fundamental frequency voltage to a much higher frequency voltage typically in the range of 25 to 40 kHz.

Table 1.6.2-1: Input current & voltage harmonic spectrum for fluorescent lamp with magnetic ballast

Harm No.	Current IR in %	Current IR in Amps	Voltage VRY in %
1	100.00	0.34	100.00
2	2.37	0.01	0.00
3	13.65	0.05	1.49
4	0.00	0.00	0.00
5	0.92	0.00	0.62
6	0.00	0.00	0.00
7	0.56	0.00	0.71
8	0.00	0.00	0.00
9	0.24	0.00	0.46
10	0.00	0.00	0.00
11	0.15	0.00	0.33
12	0.00	0.00	0.00
13	0.06	0.00	0.04
14	0.00	0.00	0.04
15	0.06	0.00	0.25
16	0.00	0.00	0.04
17	0.03	0.00	0.08
18	0.00	0.00	0.00
19	0.00	0.00	0.04
20	0.00	0.00	0.00
21	0.03	0.00	0.08
22	0.00	0.00	0.00
23	0.00	0.00	0.00
24	0.00	0.00	0.00
25	0.00	0.00	0.00
THD %	13.90		1.87
Parameter measured Amp/Volt	0.34		241.14

This high frequency has two advantages. First, a small inductor is sufficient to limit the arc current. Second, the high frequency eliminates or greatly reduces the 100- or 120-Hz flicker associated with iron-core magnetic ballast.

Standard magnetic ballasts are usually rather benign sources of additional harmonics themselves since the main harmonic distortion comes from the behavior of the arc. Figure 1.6.2-1 & table 1.6.2-1 shows a measured fluorescent lamp input current & voltage waveform and their harmonic spectrum respectively. The current THD is a moderate 15 percent. As a comparison, electronic ballasts, which employ switch-mode power supplies, can produce double or triple the standard magnetic ballast harmonic output. Figure 1.6.2-2 shows a fluorescent lamp with electronic ballast that has a current THD of 144.

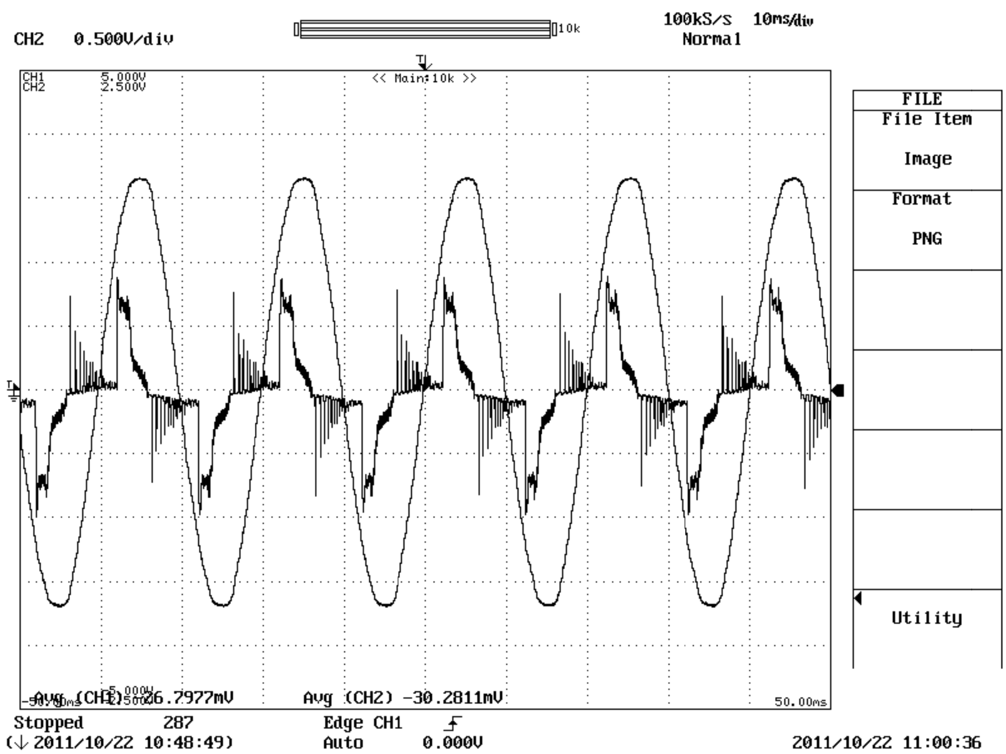


Figure 1.6.2-2: Input current & voltage waveform for fluorescent lamp with electronic ballast

Table 1.6.2-2: Input current & voltage harmonic spectrum for fluorescent lamp with electronic ballast.

Harm No.	Current I_R in %	Current I_R in Amps	Voltage V_{RY} in %
1	100.00	0.23	100.00
2	0.00	0.00	0.83
3	62.50	0.15	1.67
4	0.00	0.00	0.00
5	34.91	0.08	0.50
6	0.43	0.00	0.00
7	26.72	0.06	0.67
8	0.43	0.00	0.00
9	17.24	0.04	0.46
10	0.43	0.00	0.08
11	4.31	0.01	0.25
12	0.00	0.00	0.00
13	9.05	0.02	0.21
14	0.00	0.00	0.00
15	7.33	0.02	0.21
16	0.43	0.00	0.04
17	9.48	0.02	0.13
18	0.43	0.00	0.00
19	9.05	0.02	0.13
20	0.43	0.00	0.00
21	4.74	0.01	0.08
22	0.43	0.00	0.00
23	5.17	0.01	0.04
24	0.43	0.00	0.00
25	1.72	0.00	0.04
THD %	80.72		2.14
Parameter measured Amp/Volt	0.30		239.95

Other electronic ballasts have been specifically designed to minimize harmonics and may actually produce less harmonic distortion than the normal magnetic ballast-lamp combination. Electronic ballasts typically produce current THDs in the range of between 10 and 32 percent. A current THD greater than 32 percent is considered excessive according to ANSI C82.11-1993, *High-Frequency Fluorescent Lamp Ballasts.* Most electronic ballasts are equipped with passive filtering to reduce the input current harmonic distortion to less than 20 percent.

Since fluorescent lamps are a significant source of harmonics in commercial buildings, they are usually distributed among the phases in a nearly balanced manner. With a delta-connected supply transformer, this reduces the amount of triplen harmonic currents flowing onto the power supply system. However, it should be noted that the common wye-wye supply transformers will not impede the flow of triplen harmonics regardless of how well balanced the phases are to match the application requirement such as slowing a pump or fan. ASDs also find many applications in industrial loads.

1.6.3 ADJUSTABLE-SPEED DRIVES FOR HVAC AND ELEVATORS

Common applications of adjustable-speed drives (ASDs) in commercial loads can be found in elevator motors and in pumps and fans in HVAC systems. An ASD consists of an electronic power converter that converts ac voltage and frequency into variable voltage and frequency. The variable voltage and frequency allows the ASD to control motor speed

1.7 HARMONIC SOURCES FROM INDUSTRIAL LOADS

Modern industrial facilities are characterized by the widespread application of nonlinear loads. These loads can make up a significant portion of the total facility loads and inject harmonic currents into the power system, causing harmonic distortion in the voltage. This harmonic problem is compounded by the fact that these nonlinear loads have a relatively low power factor. Industrial facilities often utilize capacitor banks to improve the power factor to avoid penalty charges. The application of power factor correction capacitors can potentially magnify harmonic currents from the nonlinear loads, giving rise to resonance conditions within the facility. The highest voltage distortion level usually occurs at the facility's low-voltage bus where the capacitors are applied. Resonance conditions cause motor and transformer overheating, and mis-operation of sensitive electronic equipment. Nonlinear industrial loads can generally be grouped into three categories: three-phase power converters, arcing devices, and saturable devices. Sections 1.7.1 to 1.7.3 detail the industrial load characteristics.

1.7.1 THREE-PHASE POWER CONVERTERS

Three-phase electronic power converters differ from single-phase converters mainly because they do not generate third-harmonic currents. This is a great advantage because the third-harmonic current is the largest component of harmonics. However, they can still be significant sources of harmonics at their characteristic frequencies, as shown in Figure 1.7.1-1. This is a typical current source type of adjustable-speed drive. The harmonic spectrum given in Figure 1.7.1-1 would also be typical of a dc motor drive input current. Voltage source inverter drives (such as PWM-type drives) can have much higher distortion levels as shown in Figure 1.7.1-2.

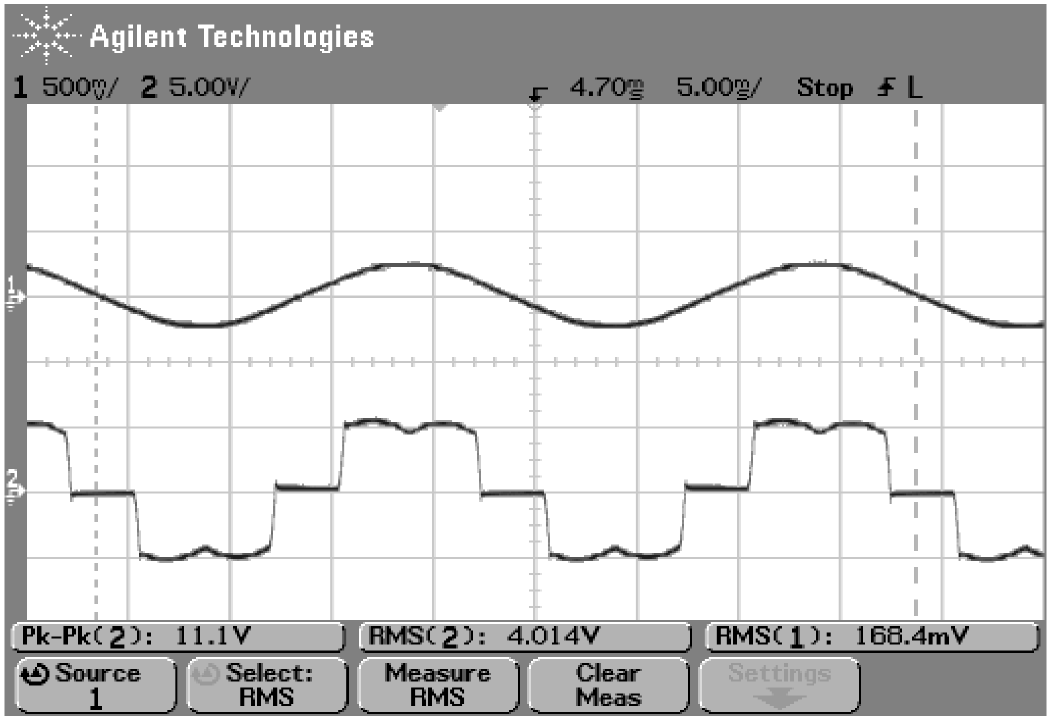


Figure 1.7.1-1: Voltage &Current and waveform for CSI-type ASD.

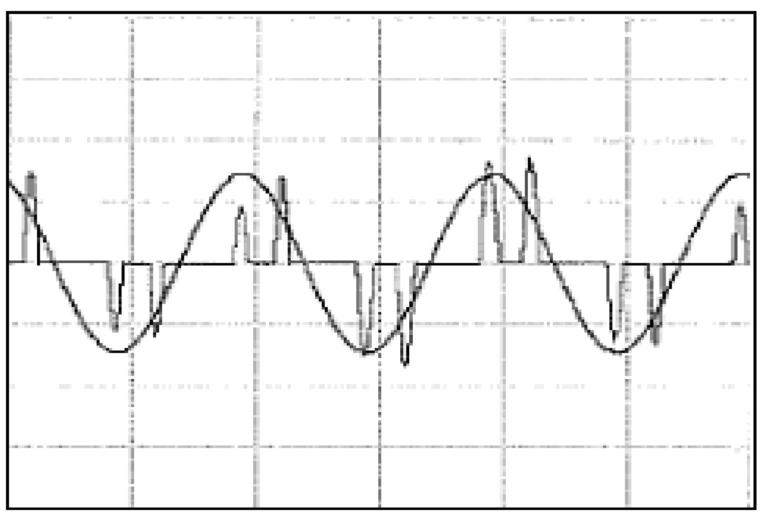


Figure 1.7.1-2: Voltage & current waveform for PWM-type ASD.

Table 1.7.1-2 Harmonic analysis of current & voltage waveform for PWM-type ASD as shown in Figure 1.7.1-2.

Harm No.	Current I_R in %	Current I_R in Amps	Voltage V_{RY} in %	Current I_Y in %	Current I_Y in Amps	Voltage V_{YB} in %	Current I_B in %	Current I_B in Amps	Voltage V_{BR} in %
1	100.00	3.39	100.00	100.00	3.37	100.00	100.00	3.14	100.00
2	2.06	0.07	0.00	3.86	0.13	0.00	1.27	0.04	0.00
3	5.60	0.19	0.00	2.37	0.08	0.24	3.82	0.12	0.27
4	2.06	0.07	0.00	6.53	0.22	0.00	2.23	0.07	0.00
5	84.66	2.87	0.24	90.50	3.05	0.00	91.72	2.88	0.27
6	1.18	0.04	0.00	6.53	0.22	0.00	1.59	0.05	0.00
7	79.06	2.68	0.24	77.74	2.62	0.48	79.30	2.49	0.48
8	1.47	0.05	0.00	0.00	0.00	0.00	1.59	0.05	0.00
9	7.67	0.26	0.24	4.75	0.16	0.00	4.78	0.15	0.12
10	1.47	0.05	0.00	5.04	0.17	0.00	2.87	0.09	0.00
11	46.31	1.57	0.24	56.97	1.92	0.24	59.55	1.87	0.24
12	0.88	0.03	0.00	5.04	0.17	0.00	1.91	0.06	0.00
13	38.94	1.32	0.24	40.65	1.37	0.24	43.31	1.36	0.46
14	0.59	0.02	0.00	2.37	0.08	0.00	1.91	0.06	0.00
15	5.01	0.17	0.00	4.45	0.15	0.00	3.50	0.11	0.00
16	0.59	0.02	0.00	1.78	0.06	0.00	2.23	0.07	0.00
17	13.27	0.45	0.00	21.07	0.71	0.24	24.20	0.76	0.22
18	0.59	0.02	0.00	3.26	0.11	0.00	0.96	0.03	0.00
19	7.96	0.27	0.00	11.28	0.38	0.00	13.06	0.41	0.00
20	0.29	0.01	0.00	1.78	0.06	0.00	1.59	0.05	0.00
21	1.18	0.04	0.00	2.08	0.07	0.00	1.27	0.04	0.00
22	0.00	0.00	0.00	1.19	0.04	0.00	1.27	0.04	0.00
23	1.47	0.05	0.00	1.78	0.06	0.00	2.87	0.09	0.00
24	0.29	0.01	0.00	2.08	0.07	0.00	0.32	0.01	0.00
25	2.65	0.09	0.00	3.26	0.11	0.00	0.96	0.03	0.00
THD %	132.17		0.55	141.34		0.68	144.91		0.84
Parameter measured	5.62 A		410.01 V	5.83 A		414.01 V	5.53 A		413.01 V

The input to the PWM drive is generally designed like a three-phase version of the switch-mode power supply in computers. The rectifier feeds directly from the ac bus to a large capacitor on the dc bus. With little intentional inductance, the capacitor is charged in very short pulses, creating the distinctive “rabbit ear” ac-side current waveform with very high distortion. Whereas the switch-mode power supplies are generally for very small loads, PWM drives are now being applied for loads up to 500 horsepower (hp). This is a justifiable cause for concern from power engineers.

DC drives. Rectification is the only step required for dc drives. Therefore, they have the advantage of relatively simple control systems. Compared with ac drive systems, the dc drive offers a wider speed range and higher starting torque. However, purchase and maintenance costs for dc motors are high, while the cost of power electronic devices has been dropping year after year. Thus, economic considerations limit use of the dc drive to applications that require the speed and torque characteristics of the dc motor. Most dc drives use the six-pulse rectifier shown in Figure 1.7.1-3. Large drives may employ a 12-pulse rectifier. This reduces thyristor current duties and reduces some of the larger ac current harmonics.

The two largest harmonic currents for the six-pulse drive are the fifth and seventh. They are also the most troublesome in terms of system response. A 12-pulse rectifier in this application can be expected to eliminate about 90 percent of the fifth and seventh harmonics, depending on system imbalances. The disadvantages of the 12-pulse drive are that there is more cost in electronics and another transformer is generally required.

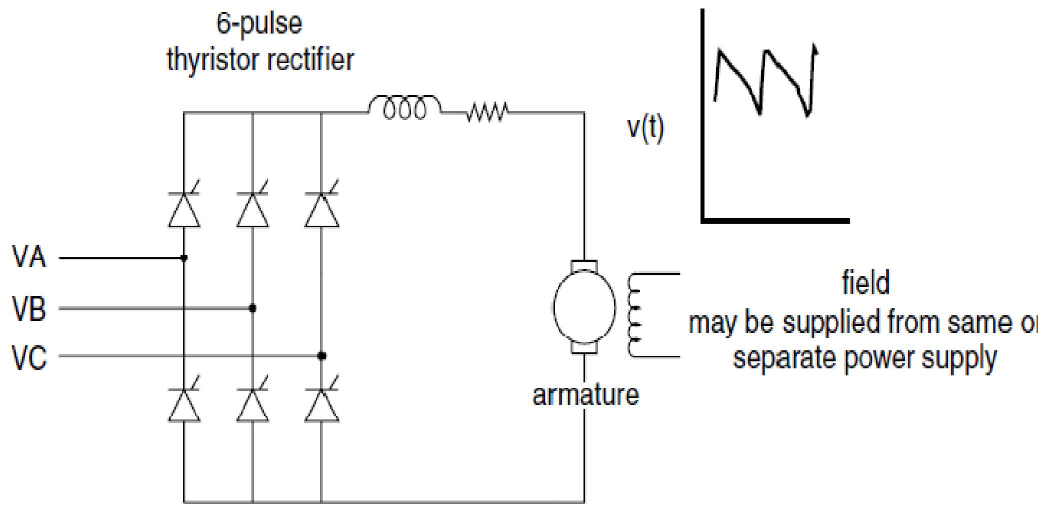


Figure 1.7.1-3: Six-pulse dc ASD.

AC drives. In ac drives, the rectifier output is inverted to produce a variable-frequency ac voltage for the motor. Inverters are classified as voltage source inverters (VSIs) or current source inverters (CSIs). A VSI requires a constant dc (i.e., low-ripple) voltage input to the inverter stage. This is achieved with a capacitor or *LC* filter in the dc link. The CSI requires a constant current input; hence, a series inductor is placed in the dc link. AC drives generally use standard squirrel cage induction motors. These motors are rugged, relatively low in cost, and require little maintenance. Synchronous motors are used where precise speed control is critical. A popular ac drive configuration uses a VSI employing PWM techniques to synthesize an ac waveform as a train of variable-width dc pulses (see Figure 1.7.1-4).

The inverter uses either SCRs, gate turnoff (GTO) thyristors, or power transistors for this purpose. Currently, the VSI PWM drive offers the best energy efficiency for applications over a wide speed range for drives up through at least 500 hp. Another advantage of PWM drives is that, unlike other types of drives, it is not necessary to vary rectifier output voltage to control motor speed. This allows the rectifier thyristors to be replaced with diodes, and the thyristor control circuitry to be eliminated.

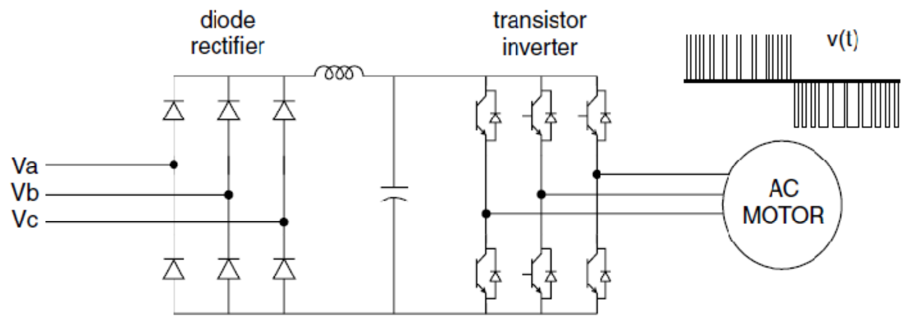


Figure 1.7.1-4: PWM ASD

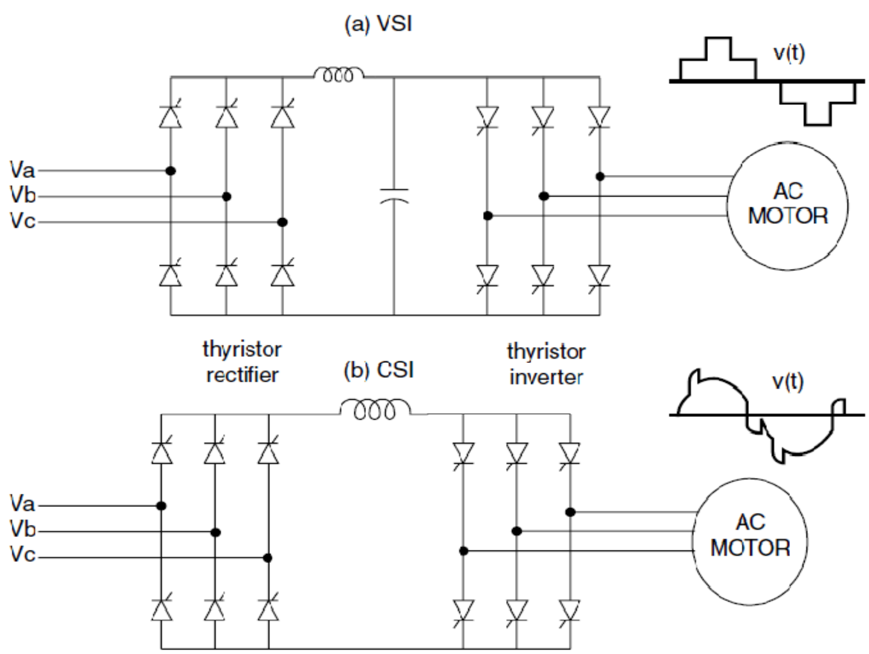


Figure 1.7.1-5: Large ac ASDs

Very high power drives employ SCRs and inverters. These may be 6- pulse, as shown in Figure 1.7.1-5, or like large dc drives, 12-pulse. VSI drives (Figure 1.7.1-5 a) are limited to applications that do not require rapid changes in speed. CSI drives (Figure 1.7.1-5 *b*) have good acceleration/deceleration characteristics but require a motor with a leading power factor (synchronous or induction with capacitors) or added control circuitry to commutate the inverter thyristors. In either case, the CSI drive must be designed for use with a specific motor. Thyristors in current source inverters must be protected against inductive voltage spikes, which increases the cost of this type of drive.

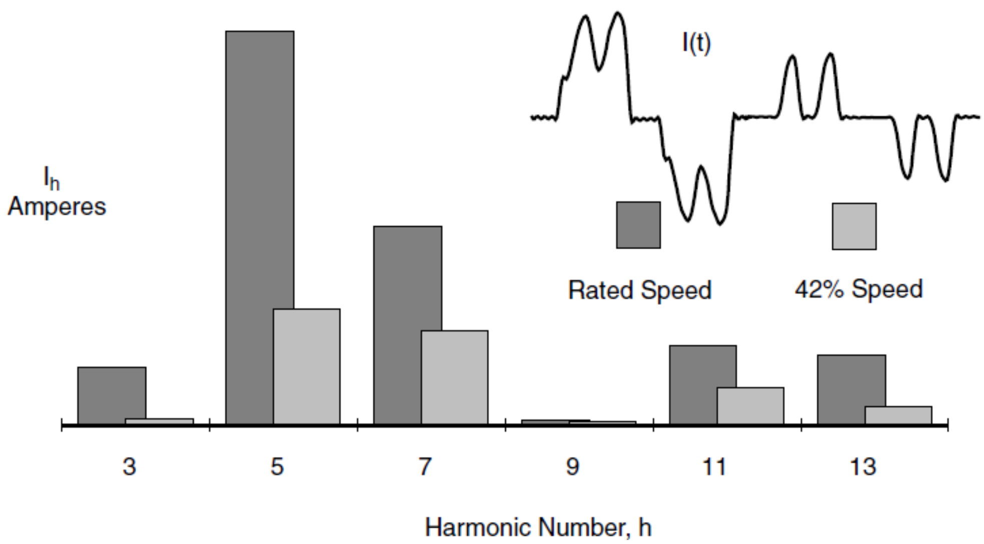


Figure 1.7.1-6: Effect of PWM ASD speed on ac current harmonics.

Impact of operating condition. The harmonic current distortion in adjustable-speed drives is not constant. The waveform changes significantly for different speed and torque values. Figure 1.7.1-6 shows two operating conditions for a PWM adjustable speed drive. While the waveform at 42 percent speed is much more distorted proportionately, the drive injects considerably higher magnitude harmonic currents at rated speed. The bar chart shows the amount of current injected. This will be the limiting design factor, not the highest THD. Engineers should be careful to understand the basis of data and measurements concerning these drives before making design decisions.

1.7.2 ARCING DEVICES

This category includes arc furnaces, arc welders, and discharge-type lighting (fluorescent, sodium vapor, mercury vapor) with magnetic (rather than electronic) ballasts. As shown in Figure 1.7.2-1, the arc is basically a voltage clamp in series with a reactance that limits current to a reasonable value. The voltage-current characteristics of electric arcs are nonlinear.

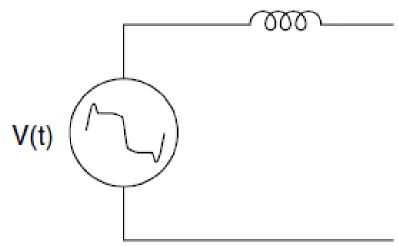


Figure 1.7.2-1: Equivalent circuit for an arcing device.

Following arc ignition, the voltage decreases as the arc current increases, limited only by the impedance of the power system. This gives the arc the appearance of having a negative resistance for a portion of its operating cycle such as in fluorescent lighting applications. In electric arc furnace applications, the limiting impedance is primarily the furnace cable and leads with some contribution from the power system and furnace transformer. Currents in excess of 60,000 Amp are common. The electric arc itself is actually best represented as a source of voltage harmonics. If a probe were to be placed directly across the arc, one would observe a somewhat trapezoidal waveform. Its magnitude is largely a function of the length of the arc. However, the impedance of ballasts or furnace leads acts as a buffer so that the supply voltage is only moderately distorted. The arcing load thus appears to be a relatively stable harmonic current source, which is adequate for most analyses. The exception occurs when the system is near resonance and a Thevenin equivalent model using the arc voltage waveform gives more realistic answers. The harmonic content of an arc furnace load and other arcing devices is similar to that of the magnetic ballast shown in Figure 1.6.2-1. Three phase arcing devices can be arranged to cancel the triplen harmonics through the transformer connection. However, this cancellation may not work in three-phase arc furnaces because of the frequent unbalanced operation during the melting phase. During the refining stage when the arc is more constant, the cancellation is better.

1.7.3 SATURABLE DEVICES

Equipment in this category includes transformers and other electromagnetic devices with a steel core, including motors. Harmonics are generated due to the nonlinear magnetizing characteristics of the steel (see Figure 1.7.3-1). Power transformers are designed to normally operate just below the “knee” point of the magnetizing saturation characteristic. The operating flux density of a transformer is selected based on a complicated optimization of steel cost, no-load losses, noise, and numerous other factors. Many electric utilities will penalize transformer vendors by various amounts for no-load and load losses, and the vendor will try to meet the specification with a transformer that has the lowest evaluated cost. A high-cost penalty on the no-load losses or noise will generally result in more steel in the core and a higher saturation curve that yields lower harmonic currents.

Although transformer exciting current is rich in harmonics at normal operating voltage (see Figure 1.7.3-2a & Figure 1.7.3-2b), it is typically less than 1 percent of rated full load current. Transformers are not as much of a concern as electronic power converters and arcing devices which can produce harmonic currents of 20 percent of their rating, or higher. However, their effect will be noticeable, particularly on utility distribution systems, which have hundreds of transformers. It is common to notice a significant increase in triplen harmonic currents during the early morning hours when the load is low and the voltage rises.

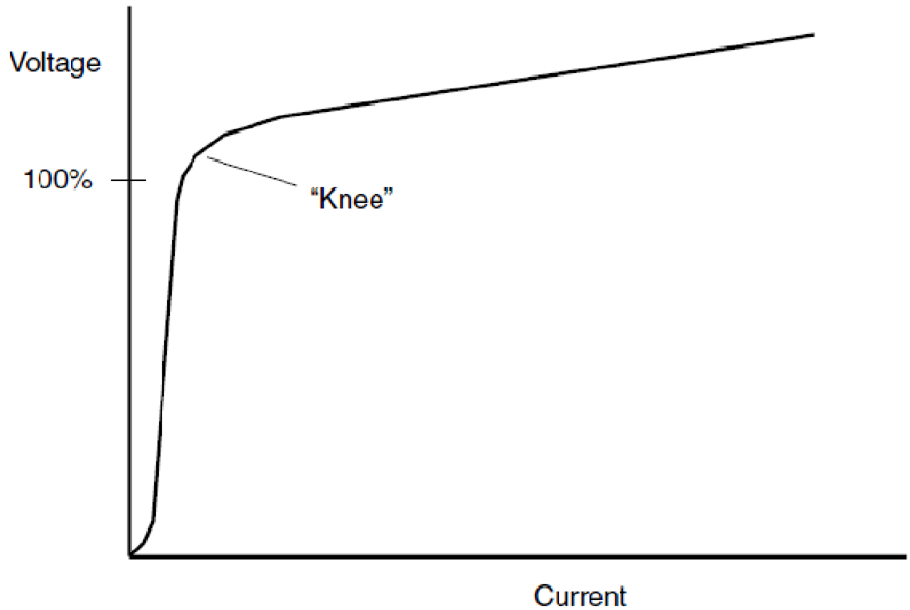


Figure 1.7.3-1: Transformer magnetizing characteristic

Transformer exciting current is more visible then because there is insufficient load to obscure it and the increased voltage causes more current to be produced. Harmonic voltage distortion from transformer over excitation is generally only apparent under these light load conditions. Some transformers are purposefully operated in the saturated region. One example is a triplen transformer used to generate 150 Hz for induction furnaces.

Motors also exhibit some distortion in the current when overexcited, although it is generally of little consequence. There are, however, some fractional horsepower, single-phase motors that have a nearly triangular waveform with significant third-harmonic currents. The waveform shown in see Figure 1.7.3-2 is for single-phase or wye grounded three-phase transformers. The current obviously contains a large amount of third harmonic. Delta connections and ungrounded wye connections prevent the flow of zero-sequence harmonic, which triplens tend to be. Thus, the line current will be void of these harmonics unless there is an imbalance in the system.

Table 1.7.3-2: FFT Analysis of Transformer magnetizing current.

Harm No.	Current I_R in %	Current I_R in Amps	Voltage V_{RY} in %
1	100.00	2.59	100.00
2	0.31	0.01	0.11
3	11.27	0.29	2.70
4	0.39	0.01	0.06
5	26.77	0.69	7.89
6	0.08	0.00	0.00
7	6.33	0.16	3.70
8	0.00	0.00	0.00
9	0.85	0.02	0.66
10	0.00	0.00	0.00
11	1.47	0.04	0.83
12	0.00	0.00	0.00
13	1.00	0.03	0.61
14	0.00	0.00	0.00
15	0.54	0.01	0.33
16	0.00	0.00	0.00
17	0.39	0.01	0.28
18	0.00	0.00	0.00
19	0.23	0.01	0.17
20	0.00	0.00	0.00
21	0.15	0.00	0.17
22	0.00	0.00	0.00
23	0.08	0.00	0.06
24	0.00	0.00	0.00
25	0.00	0.00	0.00
THD %	29.81		9.22
Parameter measured	2.70 A		36.39 kV

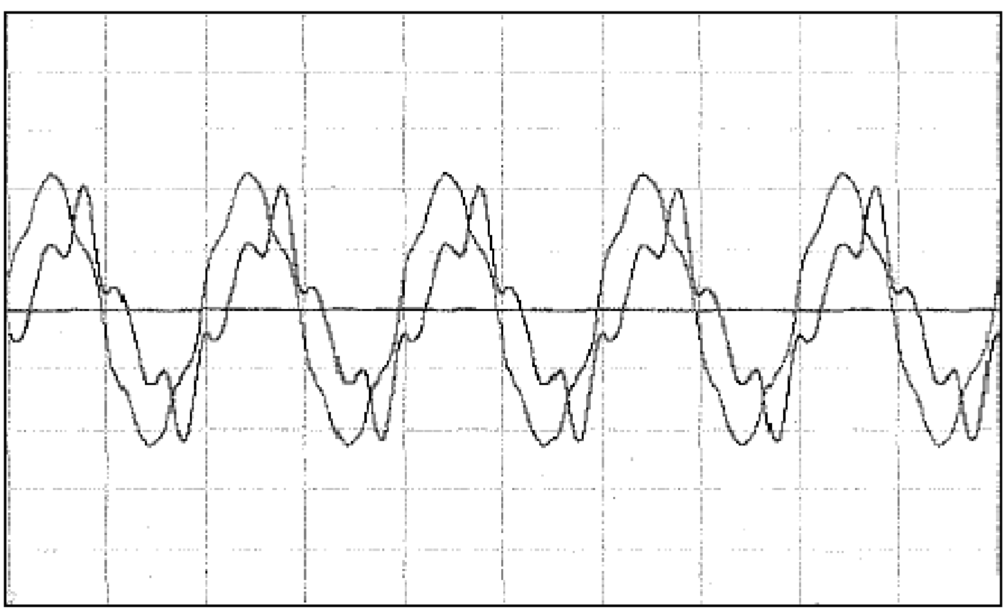


Figure 1.7.3-2a: Transformer magnetizing current waveforms and applied Voltage

1.8 SYSTEM RESPONSE CHARACTERISTICS

In power systems, the response of the system is equally as important as the sources of harmonics. In fact, power systems are quite tolerant of the currents injected by harmonic-producing loads unless there is some adverse interaction with the impedance of the system. Identifying the sources is only half the job of harmonic analysis. The response of the power system at each harmonic frequency determines the true impact of the nonlinear load on harmonic voltage distortion. There are three primary variables affecting the system response characteristics, i.e., the system impedance, the presence of a capacitor bank, and the amount of resistive loads in the system. Sections 1.8.1 through 1.8.4 detail these variables.

1.8.1 SYSTEM IMPEDANCE

At the fundamental frequency, power systems are primarily inductive, and the equivalent impedance is sometimes called simply the short-circuit reactance. Capacitive effects are frequently neglected on utility distribution systems and industrial power systems. One of the most frequently used quantities in the analysis of harmonics on power systems is the short-circuit impedance to the point on a network at which a capacitor is located. If not directly available, it can be computed from short-circuit study results that give either the short-circuit MegaVoltAmpere (MVA) or the short-circuit current as follows:

$$Z_{SC} = R_{SC} + jX_{SC} = \frac{kV^2}{MVA_{SC}} = \frac{kV\ X\ 1000}{\sqrt{3}\, I_{SC}} \qquad (1.8.1\text{-}1)$$

where Z_{SC} = short-circuit impedance

R_{SC} = short-circuit resistance

X_{SC} = short-circuit reactance

kV = phase-to-phase voltage, kV

MVA_{SC} = three-phase short-circuit MVA

I_{SC} = short-circuit current, A

Z_{SC} is a phasor quantity, consisting of both resistance and reactance.

However, if the short-circuit data contain no phase information, one is usually constrained to assuming that the impedance is purely reactive. This is a reasonably good assumption for industrial power systems for buses close to the mains and for most utility systems. When this is not the case, an effort should be made to determine a more realistic resistance value because that will affect the results once capacitors are considered. The inductive reactance portion of the impedance changes linearly with frequency. One common error made by novices in harmonic analysis is to forget to adjust the reactance for frequency. The reactance at the h^{th} harmonic is determined from the fundamental impedance reactance X_1 by:

$$X_h = hX_1 \tag{1.8.1-2}$$

In most power systems, one can generally assume that the resistance does not change significantly when studying the effects of harmonics less than the ninth. For lines and cables, the resistance varies approximately by the square root of the frequency once skin effect becomes significant in the conductor at a higher frequency. The exception to this rule is with some transformers. Because of stray eddy current losses, the apparent resistance of larger transformers may vary almost proportionately with the frequency. This can have a very beneficial effect on damping of resonance as will be shown later. In smaller transformers, less than 100 kVA, the resistance of the winding is often so large relative to the other impedances that it swamps out the stray eddy current effects and there is little change in the total apparent resistance until the frequency reaches about 500 Hz. Of course, these smaller transformers may have an X/R ratio of 1.0 to 2.0 at fundamental frequency, while large substation transformers might typically have a ratio of 20 to 30. Therefore, if the bus that is being studied is dominated by transformer impedance rather than line impedance, the system impedance model should be considered more carefully. Neglecting the resistance will generally give a conservatively high prediction of the harmonic distortion. At utilization voltages, such as industrial power systems, the equivalent system reactance is often dominated by the service transformer impedance. A good approximation for X_{SC} may be based on the impedance of the service entrance transformer only:

$$X_{SC} \equiv X_{tr} \tag{1.8.1-3}$$

While not precise, this is generally at least 90 percent of the total impedance and is commonly more. This is usually sufficient to evaluate whether or not there will be a significant harmonic resonance problem. Transformer impedance in ohms can be determined from the percent impedance Z_{tx} found on the nameplate by

$$X_{tr} = \left(\frac{kV^2}{MVA_{3\emptyset}}\right) Z_{tr}\ (\%) \tag{1.8.1-4}$$

where $MVA_{3\emptyset}$ is the kVA rating of the transformer in MVA.

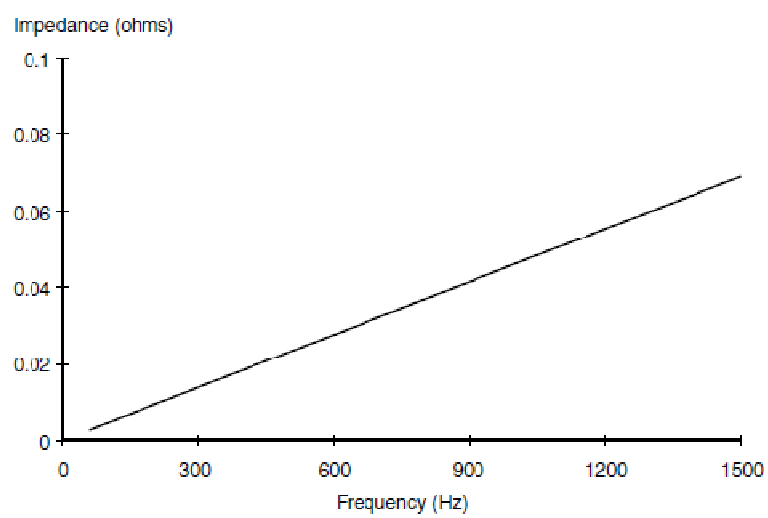


Figure 1.8.2-1: Impedance versus frequency for inductive system.

This assumes that the impedance is predominantly reactive. For example for a 1500-kVA, 6 percent transformer, the equivalent impedance on the 440-V side is A plot of impedance versus frequency for an inductive system (no capacitors installed) would look like Figure 1.8.2-1. This simple model neglects capacitance, which cannot be done for harmonic analysis.

1.8.2 CAPACITOR IMPEDANCE

Shunt capacitors, either at the customer location for power factor correction or on the distribution system for voltage control, dramatically alter the system impedance variation with frequency.

Capacitors do not create harmonics, but severe harmonic distortion can sometimes be attributed to their presence. While the reactance of inductive components increases proportionately to frequency, capacitive reactance *XC* decreases proportionately:

$$X_C = \frac{1}{2f\Pi C} \tag{1.8.2-1}$$

C is the capacitance in farads. This quantity is seldom readily available for power capacitors, which are rated in terms of kVAr or MVAr at a given voltage. The equivalent line-to-neutral capacitive reactance at fundamental frequency for a capacitor bank can be determined by

$$X_C = \frac{kV^2}{Mvar} \tag{1.8.2-2}$$

For three-phase banks, use phase-to-phase voltage and the three phase reactive power rating. For single-phase units, use the capacitor voltage rating and the reactive power rating. For example, for a three phase, 1200-kvar, 13.8-kV capacitor bank, the positive-sequence reactance in ohms would be

$$X_C = \frac{kV^2}{Mvar} = \frac{13.8^2}{1.2} = 158.7\Omega \tag{1.8.2-3}$$

1.8.3 PARALLEL RESONANCE

All circuits containing both capacitances and inductances have one or more natural frequencies. When one of those frequencies lines up with a frequency that is being produced on the power system, a resonance may develop in which the voltage and current at that frequency continue to persist at very high values. This is the root of most problems with harmonic distortion on power systems. Figure 1.8.3-1 shows a distribution system with potential parallel resonance problems. From the perspective of harmonic sources the shunt capacitor appears in parallel with the equivalent system inductance (source and transformer inductances) at harmonic frequencies as depicted in Figure 1.8.3-2 b. Furthermore, since the power system is assumed to have an equivalent voltage source of fundamental frequency only, the power system voltage source appears short circuited in the figure. Parallel resonance occurs when the reactance of *XC* and the distribution system cancel each other out. The frequency at which this phenomenon occurs is called the parallel resonant frequency. It can be expressed as follows:

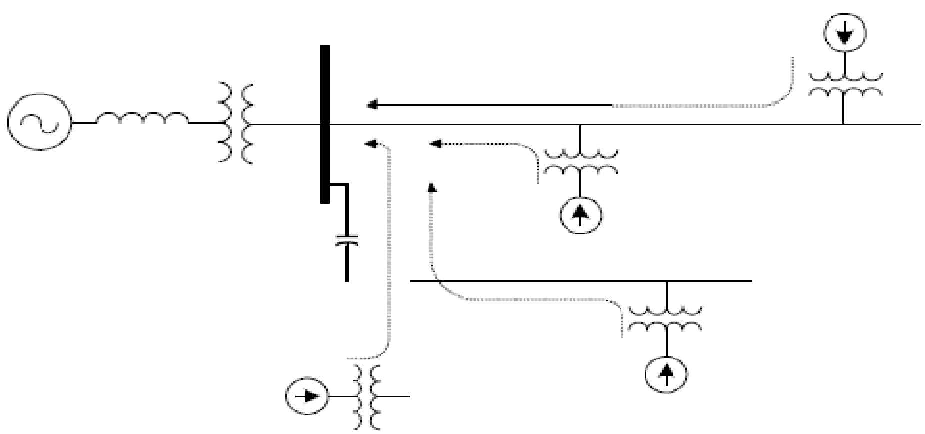


Figure 1.8.3-1: System with parallel resonance problem

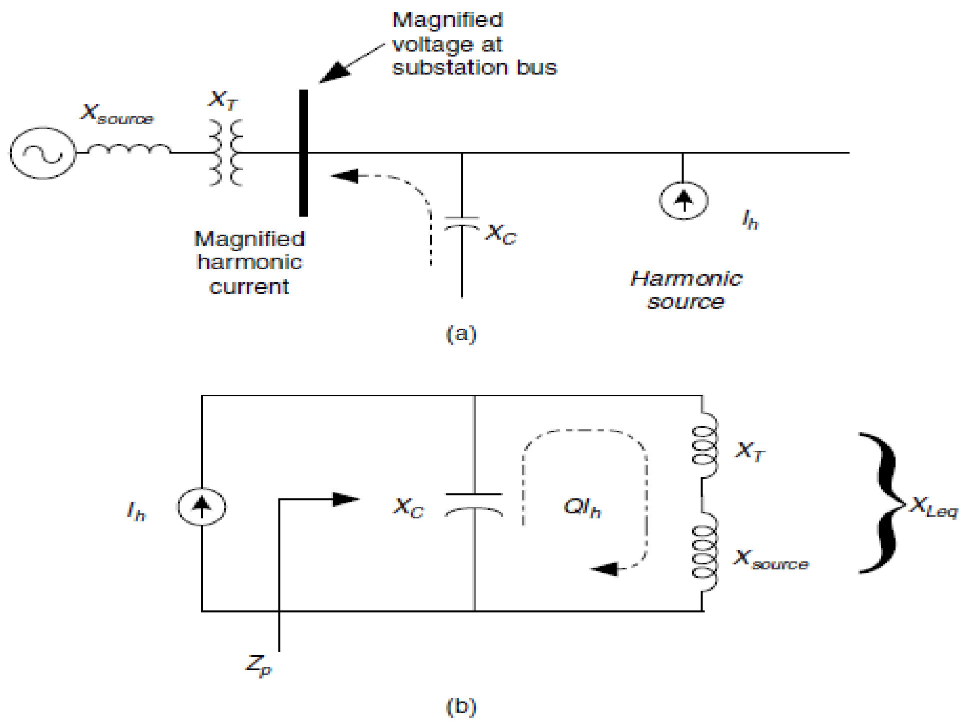


Figure 1.8.3-2: At harmonic frequencies, the shunt capacitor bank appears in parallel with the system inductance. (*a*) Simplified distribution circuit; (*b*) parallel resonant circuit as seen from the harmonic source

$$f_p = \frac{1}{2\Pi}\sqrt{\frac{1}{L_{eq}\,C} - \frac{R^2}{4L_{eq}^2}} \equiv \frac{1}{2\Pi}\sqrt{\frac{1}{L_{eq}\,C}} \qquad (1.8.3\text{-}1)$$

where R = resistance of combined equivalent source and transformer (not shown in Figure 1.8.3-2)

L_{eq} = inductance of combined equivalent source and transformer

C = capacitance of capacitor bank

At the resonant frequency, the apparent impedance of the parallel combination of the equivalent inductance and capacitance as seen from the harmonic current source becomes very large, i.e.,

$$Z_p = \frac{X_C\,(X_{Leq} + R)}{X_C + X_{Leq} + R} = \frac{X_C\,(X_{Leq} + R)}{R}$$

$$\equiv \frac{X_{Leq}^2}{R} = \frac{X_C^2}{R} = QX_{Leq} = QX_C \qquad (1.8.3\text{-}2)$$

where $Q = XL/R = XC/R$ and $R \ll X_{L\text{Equation}}$ Keep in mind that the reactance in this equation is computed at the resonant frequency. Q often is known as the quality factor of a resonant circuit that determines the sharpness of the frequency response. Q varies considerably by location on the power system. It might be less than 5 on a distribution feeder and more than 30 on the secondary bus of a large step-down transformer.

From Equation (1.8.3-2), it is clear that during parallel resonance, a small harmonic current can cause a large voltage drop across the apparent impedance, i.e., $V_p = Q\, X_{L\text{eq}}\, I_h$. The voltage near the capacitor bank will be magnified and heavily distorted. Let us now examine current behavior during the parallel resonance. Let the current flowing in the capacitor bank or into the power system be $I_{\text{resonance}}$; thus,

$$I_{resonance} = \frac{V_p}{X_C} = \frac{QX_C I_h}{X_C} = QI_h \tag{1.8.3-3-a}$$

or

$$I_{resonance} = \frac{V_p}{X_{Leq}} = \frac{QX_{Leq} I_h}{X_{Leq}} = QI_h \tag{1.8.3-3-b}$$

From Equation (1.8.3-3), it is clear that currents flowing in the capacitor bank and in the power system (i.e., through the transformer) will also be magnified Q times. These phenomenon will likely cause capacitor failure, fuse blowing, or transformer overheating.

The extent of voltage and current magnification is determined by the size of the shunt capacitor bank. Figure 1.8.3-3 shows the effect of varying capacitor size in relation to the transformer on the impedance seen from the harmonic source and compared with the case in which there is no capacitor. The following illustrates how the parallel resonant frequency is computed. Power systems analysts typically do not have *L* and *C* readily available and prefer to use other forms of this relationship.

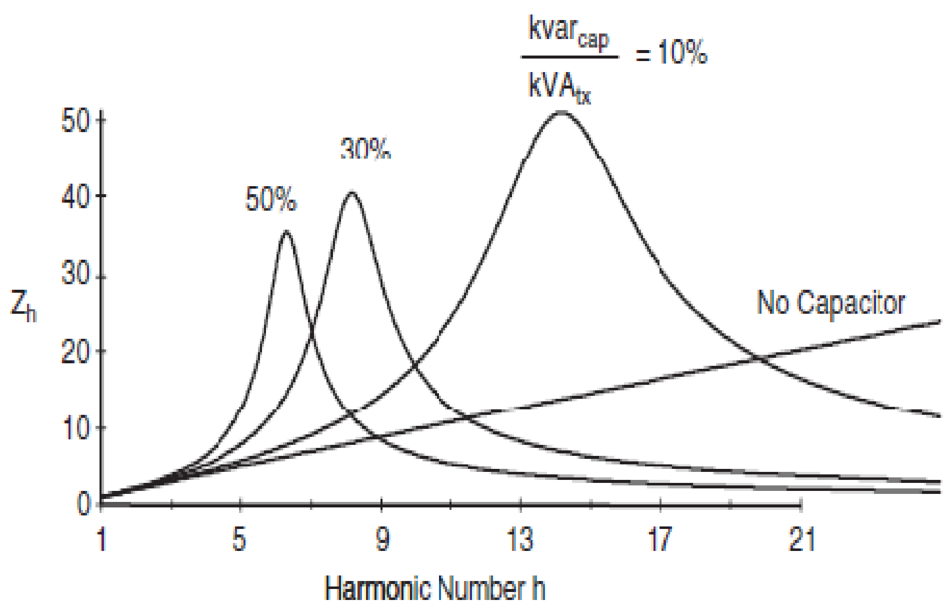


Figure 1.8.3-3: System frequency response as capacitor size is varied in relation to transformer

They commonly compute the resonant harmonic *hr* based on fundamental frequency impedances and ratings using one of the following:

$$h_r = \sqrt{\frac{X_C}{X_{SC}}} = \sqrt{\frac{MVA_{SC}}{Mvar_{cap}}} \equiv \sqrt{\frac{kVA_{tx}\ X\ 100}{kvar_{cap}\ X\ Z_{tx}\ (\%)}} \qquad (1.8.3\text{-}4)$$

where h_r = resonant harmonic

X_C = capacitor reactance

X_{SC} = system short-circuit reactance

MVA_{SC} = system short-circuit MVA

MVA_{cap} = MVAr rating of capacitor bank

kVA_{tx} = kVA rating of step-down transformer

Z_{tx} = step-down transformer impedance

$kVAr_{cap}$ = kVAr rating of capacitor bank

For example, for an industrial load bus where the transformer impedance is dominant, the resonant harmonic for a 1500-kVA, 6 percent transformer and a 500-kvar capacitor bank is approximately

$$h_r \equiv \sqrt{\frac{kVA_{tx}\ X\ 100}{kVAr_{cap}\ X\ Z_{tx}\ (\%)}} = \sqrt{\frac{1500\ X\ 100}{500\ X\ 6}} = 7.07 \qquad (1.8.3\text{-}5)$$

1.8.4 SERIES RESONANCE

There are certain instances when a shunt capacitor and the inductance of a transformer or distribution line may appear as a series *LC* circuit to a source of harmonic currents. If the resonant frequency corresponds to a characteristic harmonic frequency of the nonlinear load, the *LC* circuit will attract a large portion of the harmonic current that is generated in the distribution system. A customer having no nonlinear load, but utilizing power factor correction capacitors, may in this way experience high harmonic voltage distortion due to neighboring harmonic sources. This situation is depicted in Figure 1.8.4-1.

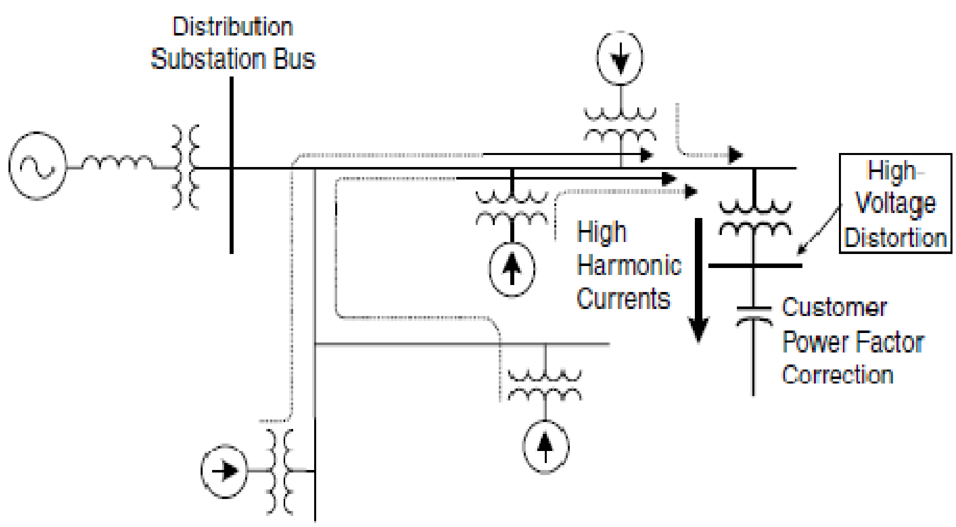


Figure 1.8.4-1: System with potential series resonance problems

During resonance, the power factor correction capacitor forms a series circuit with the transformer and harmonic sources. The simplified circuit is shown in Figure 1.8.4-2. The harmonic source shown in this Figure represents the total harmonics produced by other loads. The inductance in series with the capacitor is that of the service entrance transformer. The series combination of the transformer inductance and the capacitor bank is very small (theoretically zero) and only limited by its resistance. Thus the harmonic current corresponding to the resonant frequency will flow freely in this circuit. The voltage at the power factor correction capacitor is magnified and highly distorted. This is apparent from the following equation:

$$V_s\ (at\ power\ factor\ capacitor\ bank) = \frac{X_C}{X_T + X_C + R} V_h \equiv \frac{X_C}{R} V_h \qquad (1.8.4\text{-}1)$$

where V_h and V_s are the harmonic voltage corresponding to the harmonic current I_h and the voltage at the power factor capacitor bank, respectively. The resistance R of the series resonant circuit is not shown in Figure 1.8.4-2, and it is small compared to the reactance. The negligible impedance of the series resonant circuit can be exploited to absorb desired harmonic currents.

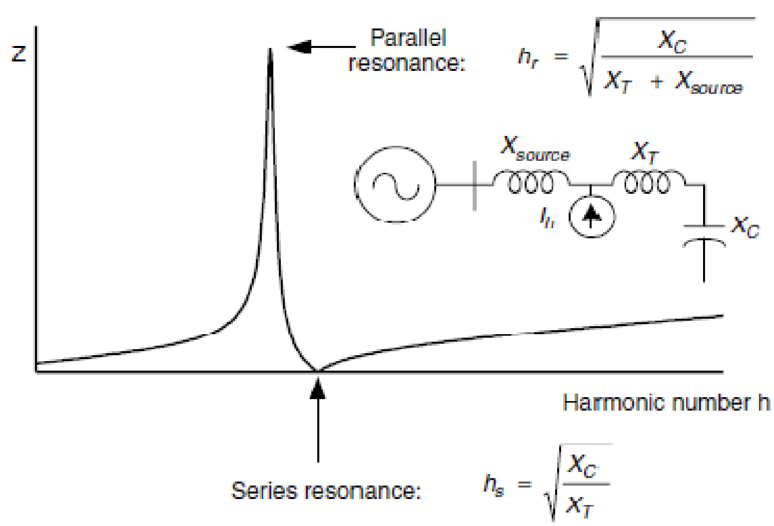


Figure 1.8.4-2: Frequency response of a circuit with series resonance

This is indeed the principle in designing a notch filter. In many systems with potential series resonance problems, parallel resonance also arises due to the circuit topology. One of these is shown in Figure 1.8.4-2 where the parallel resonance is formed by the parallel combination between X_{source} and a series between X_T and X_C. The resulting parallel resonant frequency is always smaller than its series resonant.

$$h_r = \sqrt{\frac{X_C}{X_T + X_{source}}} \qquad (1.8.4\text{-}2)$$

1.8.5 EFFECTS OF RESISTANCE AND RESISTIVE LOAD

Determining that the resonant harmonic aligns with a common harmonic source is not always cause for alarm. The damping provided by resistance in the system is often sufficient to prevent catastrophic voltages and currents. Figure 1.8.5-1 shows the parallel resonant circuit impedance characteristic for various amounts of resistive load in parallel with the capacitance. As little as 10 percent resistive loading can have a significant beneficial impact on peak impedance.

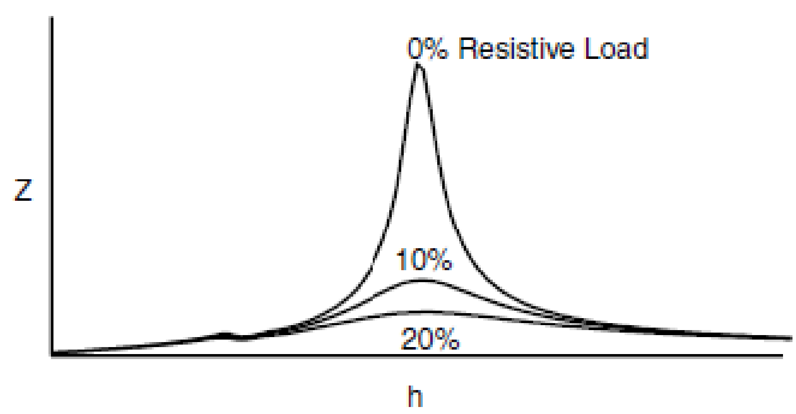


Figure 1.8.5-1: Effect of resistive loads on parallel resonance

Likewise, if there is a significant length of lines or cables between the capacitor bus and the nearest up line transformer, the resonance will be suppressed. Lines and cables can add a significant amount of the resistance to the equivalent circuit. Loads and line resistances are the reasons why catastrophic harmonic problems from capacitors on utility distribution feeders are seldom seen. That is not to say that there will not be any harmonic problems due to resonance, but the problems will generally not cause physical damage to the electrical system components. The most troublesome resonant conditions occur when capacitors are installed on substation buses, either utility substations or in industrial facilities. In these cases, where the transformer dominates the system impedance and has a high X/R ratio, the relative resistance is low and the corresponding parallel resonant impedance peak is very sharp and high. This is a common cause of capacitor, transformer, or load equipment failure.

While utility distribution engineers may be able to place feeder banks with little concern about resonance, studies should always be performed for industrial capacitor applications and for utility substation applications. Utility engineers familiar with the problems indicate that about 20 percent of industrial installations for which no studies are performed have major operating disruptions or equipment failure due to resonance. In fact, selecting capacitor sizes from manufacturers' tables to correct the power factor based on average monthly billing data tends to result in a combination that tunes the system near the fifth harmonic. This is one of the worst harmonics to which to be tuned because it is the largest frequency component in the system.

It is a misconception that resistive loads damp harmonics because in the absence of resonance, loads of any kind will have little impact on the harmonic currents and resulting voltage distortion. Most of the current will flow back into the power source. However, it is very appropriate to say that resistive loads will damp *resonance,* which will lead to a significant reduction in the harmonic distortion.

Motor loads are primarily inductive and provide little damping. In fact, they may increase distortion by shifting the system resonant frequency closer to a significant harmonic. Small, fractional-horsepower motors may contribute significantly to damping because their apparent X/R ratio is lower than that of large three-phase motors.

1.9 EFFECTS OF HARMONIC DISTORTION

Harmonic currents produced by nonlinear loads are injected back into the supply systems. These currents can interact adversely with a wide range of power system equipment, most notably capacitors, transformers, and motors, causing additional losses, overheating, and overloading. These harmonic currents can also cause interference with telecommunication lines and errors in power metering. Sections 1.9.1 through 1.9.5 discuss impacts of harmonic distortion on various power system components.

1.9.1 IMPACT ON CAPACITORS

Problems involving harmonics often show up at capacitor banks first. As discussed in Sections. 1.8.3 & 1.8.4, a capacitor bank experiences high voltage distortion during resonance. The current flowing in the capacitor bank is also significantly large and rich in a monotonic harmonic. Figure 1.9.1-1 shows a current waveform of a capacitor bank in resonance with the system at the 11th harmonic. The harmonic current shows up distinctly, resulting in a waveform that is essentially the 11th harmonic riding on top of the fundamental frequency. This current waveform typically indicates that the system is in resonance and a capacitor bank is involved. In such a resonance condition, the rms current is typically higher than the capacitor rms current rating.

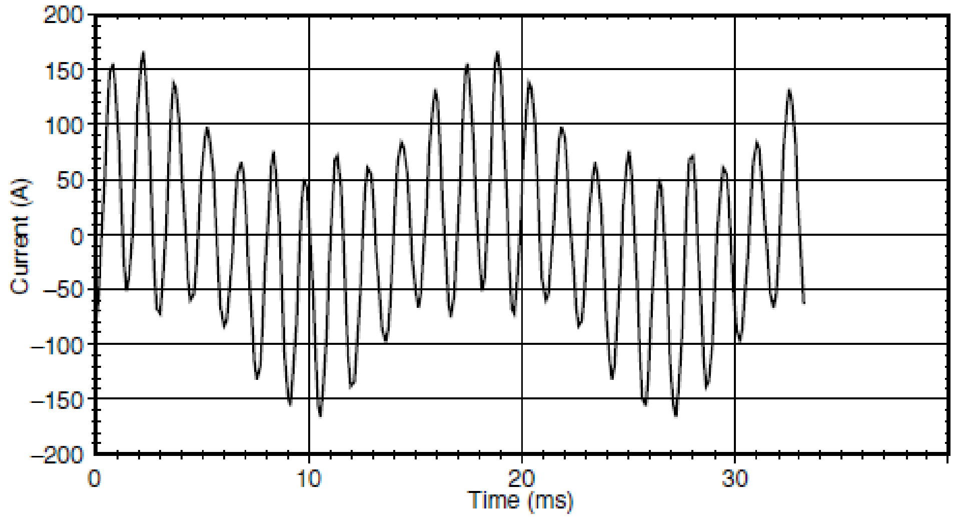


Figure 1.9.1-1: Typical capacitor current from a system in 11th-harmonic resonance.

IEEE Standard for Shunt Power Capacitors (IEEE Standard 18-1992) specifies the following continuous capacitor ratings:

■ 135 percent of nameplate kVAr

■ 110 percent of rated rms voltage (including harmonics but excluding transients)

■ 180 percent of rated rms current (including fundamental and harmonic current)

■ 120 percent of peak voltage (including harmonics)

Table 1.9.2-1 summarizes an example capacitor evaluation using a computer spreadsheet that is designed to help evaluate the various capacitor duties against the standards. The fundamental full-load current for the 1200-kvar capacitor bank is determined from

$$I_C = \frac{kvar_{3\emptyset}}{\sqrt{3}\ X\ kV_{LL}} = \frac{1200}{\sqrt{3}\ X\ 11.0} = 62.98\ A$$

The capacitor is subjected principally to two harmonics: the fifth and the seventh. The voltage distortion consists of 5 percent fifth and 4 percent seventh. This results in 25 percent fifth harmonic current and 28 percent seventh harmonic current. The resultant values all come out well below standard limits in this case, as shown in the box at the bottom of Table 1.9.1-1.

Table 1.9.1-1: Example Capacitor Evaluation

Recommended Practice for Establishing Capacitor Capabilities when Supplied by Nonsinusoidal Voltage
IEEE Std 18-2002

Capacitor Bank Data:

Bank Rating:	1200	kVAr
Voltage Rating:	11000	V (L-L)
Operating Voltage:	11000	V (L-L)
Supplied Compensation:	1200	kVAr
Fundamental Current Rating:	62.98	Amps
Fundamental Frequency:	50	Hz
Capacitive Reactance:	100.8	Ohm

Harmonic Distribution of Bus Voltage:

Harmonic Number	Frequency (Hz)	Voltage Mag. Vh (% of Fund.)	Voltage Mag. Vh (Volts)	Line Current Ih (% of Fund.)	Line Current Ih (Amp)
1	50	100.00	6198.0	100.0	61.47
3	150	0.00	0.0	0.0	0.00
5	250	5.00	310.0	25.0	15.37
7	350	4.00	248.0	28.0	17.22
9	450	0.00	0.0	0.0	0.00
11	550	0.00	0.0	0.0	0.00
13	650	0.00	0.0	0.0	0.00
15	750	0.00	0.0	0.0	0.00
17	850	0.00	0.0	0.0	0.00
19	950	0.00	0.0	0.0	0.00
21	1050	0.00	0.0	0.0	0.00
23	1150	0.00	0.0	0.0	0.00
25	1250	0.00	0.0	0.0	0.00

Voltage Distortion (THD):	6.41	%
RMS Capacitor Voltage:	6210.70	Volts
Capacitor Current Distortion:	37.55	%
RMS Capacitor Current:	65.66	Amps

Capacitor Bank Data:

	Calculated	Limit	Exceeds Limits
Peak Voltage:	101.0 %	120%	No
RMSVoltage:	97.8 %	110%	No
RMS Current:	104.2 %	180%	No
kVAr:	101.95 %	135%	No

1.9.2 IMPACT ON TRANSFORMERS

Transformers are designed to deliver the required power to the connected loads with minimum losses at fundamental frequency. Harmonic distortion of the current, in particular, as well as of the voltage will contribute significantly to additional heating. To design a transformer to accommodate higher frequencies, designers make different design choices such as using continuously transposed cable instead of solid conductor and putting in more cooling ducts. As a general rule, a transformer in which the current distortion exceeds 5 percent is a candidate for derating for harmonics. The effect of Harmonics on transformer losses is given in Annexure-I-A

There are three effects that result in increased transformer heating when the load current includes harmonic components:

1. *RMS current.* If the transformer is sized only for the kVA requirements of the load, harmonic currents may result in the transformer rms current being higher than its capacity. The increased total rms current results in increased conductor losses.
2. *Eddy current losses. These are induced currents in a transformer caused by the magnetic fluxes. These induced currents flow in the windings, in the core, and in other conducting bodies subjected to the magnetic field of the transformer and cause additional heating. This component of the transformer losses increases with the square of the frequency of the current causing the eddy currents. Therefore, this becomes a very important component of transformer losses for harmonic heating.*
3. *Core losses. The increase in core losses in the presence of harmonics will be dependent on the effect of the harmonics on the applied voltage and the design of the transformer core. Increasing the voltage distortion may increase the eddy currents in the core laminations. The net impact that this will have depends on the thickness of the core laminations and*

the quality of the core steel. The increase in these losses due to harmonics is generally not as critical as the previous two.

Exceptions: There are often cases with transformers that do not appear to have a harmonics problem from the criteria given in Table 1.9.2-1, yet are running hot or failing due to what appears to be overload. One common case found with grounded-wye transformers is that the line currents contain about 8 percent third harmonic, which is relatively low, and the transformer is overheating at less than rated load. Why would this transformer pass the heat run test in the factory and perhaps, an overload test also, and fail to perform as expected in practice? Discounting mechanical cooling problems, chances are good that there is some conducting element in the magnetic field that is being affected by the harmonic fluxes. Three of several possibilities are as follows:

■ Zero-sequence fluxes will "escape" the core on three-legged core designs (the most popular design for utility distribution substation transformers). This is illustrated in Figure 1.9.2-1. The 3d, 9th, 15th, etc., harmonics are predominantly zero-sequence. Therefore, if the winding connections are proper to allow zero-sequence current flow, these harmonic fluxes can cause additional heating in the tanks, core clamps, etc., that would not necessarily be found under balanced three-phase or single-phase tests. The 8 percent line current previously mentioned translates to a neutral third-harmonic current of 24 percent of the phase current. This could add considerably to the leakage flux in the tank and in the oil and air space. Two indicators are charred or bubbled paint on the tank and evidence of heating on the end of a bayonet fuse tube (without blowing the fuse) or bushing end.

DC offsets in the current can also cause flux to "escape" the confines of the core. The core will become slightly saturated on, for example, the positive half cycle while remaining normal for the negative half cycle. There are a number of electronic power converters that produce current waveforms that are nonsymmetrical either by accident or by design. This can result in a small dc offset on the load side of the transformer (it can't be measured from the source side). Only a small amount of dc offset is required to cause problems with most power transformers.

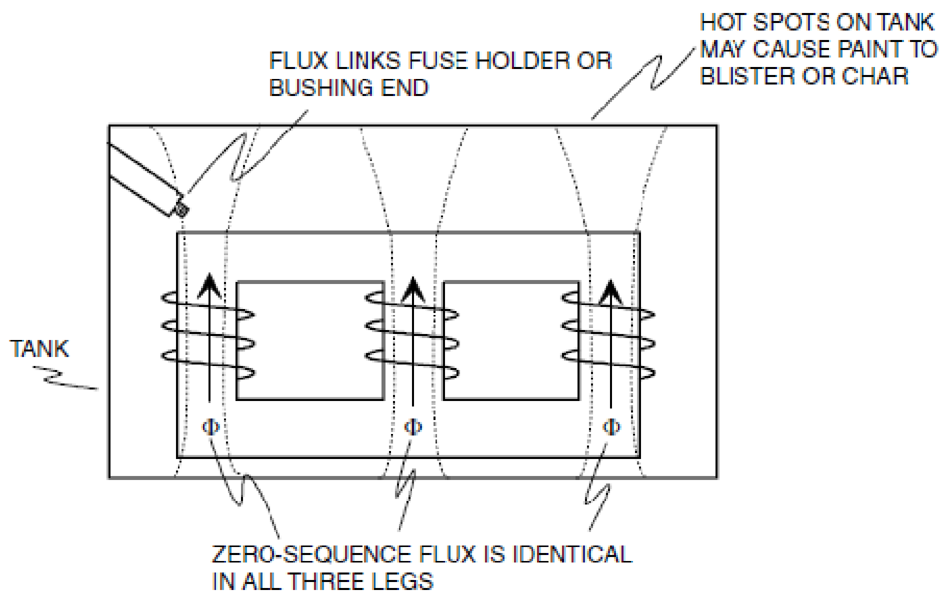


Figure 1.9.2-1: Zero Sequence flux in three legged core transformers enters the tank and the air and oil space

■ There may be a clamping structure, bushing end, or some other conducting element too close to the magnetic field. It may be sufficiently small in size that there is no notable effect in stray losses at fundamental frequency but may produce a hot spot when subjected to harmonic fluxes.

1.10 POWER IN DISTORTED AC NETWORKS

The active and reactive power components for electric circuits with sinusoidal and linear loads are well established. In the case of non-linear loads, however the use of the reactive and the harmonic power is an actual necessity for accomplishing reactive power compensation and/or harmonic filtering. The components of the electric power are shown in Figure 1.10-1 assuming a sinusoidal voltage supply and a non-linear load [8]. In this case, the power factor *($\cos\Phi$)* is the product of the distortion factor *($I_1/I= \cos\gamma$)* by displacement factor *($\cos\Phi_l$)* :

Power Factor = Distortion factor **x** Displacement Factor

The displacement factor corresponds to the power factor of systems without harmonics. This factor may be called fundamental power factor, as it depends only on the current fundamental component. On the other hand, the power factor, as defined above, may be called total power factor, as it depends on fundamental and al1 harmonic components. Hence, harmonics cause lower power factor. The power components in distorted networks are represented in power tetrahedron, instead of the triangle as in the linear case, is shown in Figure 1.10-2.

From Figure 1.10-1 and Figure 1.10-2, various important factors are determined on the use of reactive power and harmonic compensation:

(i) The reactive component is dependent only on the current component at fundamental frequency. The reactive component can be eliminated by using a conveniently chosen capacitor or inductor. The connection of an inductor or a capacitor component in parallel with load allows the generation of a current at fundamental frequency that absorbs or generates the reactive power required by the load;

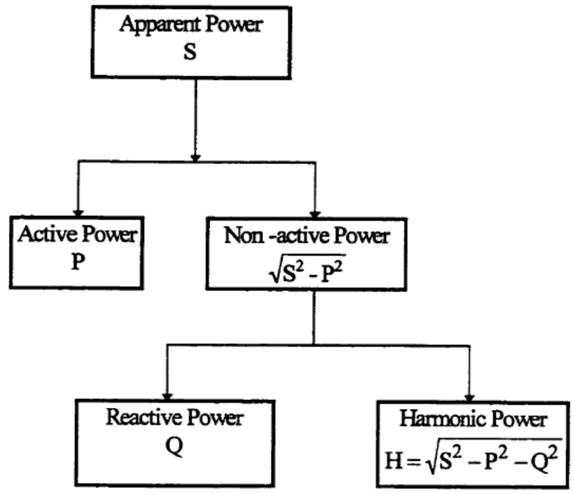


Figure 1.10-1: Components of electric power

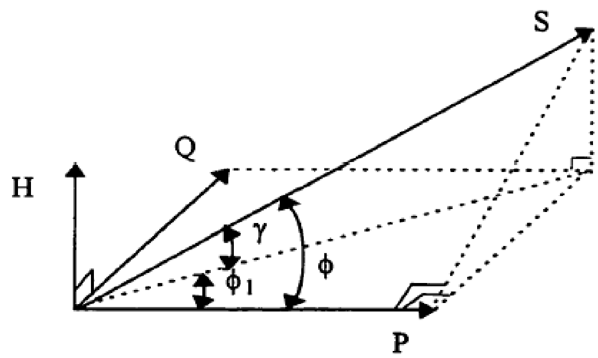


Figure 1.10-2: Power Tetrahedron.

(ii) Harmonic power is reactive power. There will be real power due to harmonics also this is included in "P". The distorting power is dependent on the current components with frequencies different from fundamental frequency (harmonics) and cannot be eliminated by a single capacitor or inductor. The elimination of harmonic power depends on filters that work as short-circuit for the harmonic current generated by the load.

1.11 HARMONIC STANDARDS AND RECOMMENDED PRACTICES.

In view of the proliferation of the power electronic equipment connected to the utility distribution system, various international agencies have proposed limits on the magnitude of harmonic current injected into the supply to maintain acceptable power quality. The resulting guidelines and standards specify limits on the magnitudes of harmonic currents and harmonic voltage distortion at various harmonic frequencies.

The most widely known are the IEEE-519 guidelines [2] in North America and the IEC-61000 Standard (formerly IEC 555) [3] prepared by the International Electrical Commission (in effect since 1996). However, the approach taken in these documents is drastically different. The IEC Standard imposes limits on individual equipment (up to 15 A, 220 V) connected to the supply, whereas the IEEE Recommended Practice addresses the issue of harmonic distortion at the point of common of coupling (PCC). Complying with the IEC Standard usually requires special design of the equipment itself *[5]* [6] [7]. However, meeting the IEEE guidelines can be achieved by means of filter, particularly active filters. Therefore, reference in this work will only be made to the IEEE guideline. IEEE 519 proposes to designers of industrial plants harmonic limits as given in Table 1.11-1 and 1.11-2. For existing installations, harmonic mitigation techniques may have to be used to reduce distortion to the specified limits.

Table 1.11-1: IEEE-519 Voltage distortions limits

Bus Voltage @PCC	HDv(%)	THDv(%)
69 KV and below	3.0	5.0

HDv=Individual harmonic voltage distortion.

Table 1.11-2: IEEE-519 Maximum odd-harmonic current distortions.

% Limits of Harmonic Currents (Bus voltage @ PCC<69 KV)

I_{SC}/I_1	h<11	11<h<17	17<h<23	23<h<35	35<h	THD
<20	4.0	2.0	1.5	0.3	0.3	5.0
20-50	7.0	3.5	2.5	1.0	0.5	8.0
50-100	10.0	4.5	4.0	1.5	0.7	12.0
100-1000	12.0	5.5	5.0	2.0	1.0	15.0
>1000	15.0	7.0	6.0	2.5	1.4	20.0

$\mathbf{I_{sc}}$ is the maximum short-circuit current @PCC.

I_1 is the maximum fundamental frequency load current @PCC.

1.12 LITERATURE SURVEY

Most of the research work in the three-phase power converter area was for balanced three-phase systems, such as motor drive applications, where three-legged power converters are used. When the neutral connection is needed, typically split DC link capacitors are used. The first four-legged current source thyristor inverter was presented in [E2] in 1979, where the fourth leg was used for commutation of thyristors. A four-legged power inverter was also proposed in [B19] to eliminate the common mode noise. The first voltage source four-legged inverter used for an aircraft power generation application was presented in [E1] in 1993 to provide the neutral connection and to handle the neutral current. Since it is a resonant DC link inverter (RDCL inverter), it is controlled with a pulse density modulation scheme. The same topology was proposed in [G2] in 1992 for active filter applications to deal with a zero-sequence component in a power system. The current source four-legged inverter was also proposed for active filter applications in [G3] in the same year to handle a zero-sequence component. Another variation of four-legged inverter was proposed in [G23], where the neutral point is still provided by two split capacitors, and the fourth leg is really an active filter independent from the three phase legs. The fourth leg is controlled to nullify the zero-sequence current due to unbalanced load or nonlinear load so that the neutral current does not flow through the DC link capacitors. This approach allows a smaller DC link capacitance to be used for the same voltage ripple; however, it still suffers from insufficient utilization of the DC link voltage. In all the previous research, there are no space vector modulation schemes proposed for four-legged power converters, nor are the modeling and control aspects of four-legged power converters discussed.

Pulse width modulation (PWM) techniques have been widely used in the field since 1960s. With intensive research activities over the last decades, many PWM schemes have been proposed for single-phase and three-phase applications, including sinusoidal PWM, harmonic elimination or optimal PWM [B24-B25], hysteresis and bang-bang type modulation [B28-B32], random modulation [B33-B35], and space vector modulation [B2-B17]. Space vector modulation was first proposed in [B2] in 1982 and became more and more popular due to its merits of high utilization of the DC link voltage, possible optimized output distortion and switching losses, and compatibility with a digital controller. It has been widely used for high performance three-phase drive systems [B3-B7] and PFC rectifiers [B8] with success. Several research focuses for space vector modulation can be identified:

(1) optimized space vector modulation schemes in terms of harmonic distortion [B13-B14] and switching losses [B10-B12];

(2) digital implementation of space vector modulation schemes [B4, B14, B16];

(3) over modulation operation [B9]; and [B4] adaptive sequencing schemes for varying modulation index and load power factor [B15, B17].

All the existing space vector modulation schemes are implemented in a two-dimensional space, and are therefore unable to deal with the zero sequence component caused by unbalanced or nonlinear loads.

Unbalanced source and load have been analyzed extensively in power systems. The symmetrical component representation was proposed by C. L. Fortescue in 1918, and became a textbook method when analyzing unbalance in power systems. By decomposing an arbitrarily unbalanced three-phase variable into three sets of balanced three-phase variables, namely, positive-sequence, negative-sequence and zero-sequence, not only is the analysis greatly simplified, but also more physical meaning can be obtained from the unbalanced conditions. In power systems, there are several passive means to correct the negative-sequence and zero-sequence components caused by unbalanced load, such as zero-sequence trap [A4], zigzag transformer [A5], and passive balancing network [A6]. In a power electronics system, there are three ways to correct the negative-sequence component caused by a unbalanced load:

(1) Large passive filter to reduce the 2w ripple;

(2) High bandwidth feedback control so that the disturbance caused by the 2w ripple can be suppressed;

(3) Feed forward control to counterbalance the disturbance of the 2w ripple; or combination of (2) and (3).

Using three-legged power converters to deal with unbalanced source has been addressed in [F1-F3]. By engaging a feed forward control, the negative-sequence component caused by an unbalanced source can be canceled out so that the input power becomes a constant and the DC link voltage is free of low frequency even harmonic ripples. The same concept was used for active power filter application in [G1]. However, a three-legged power converter is incapable of dealing with zero-sequence unbalance. To solve the limitation, normally split DC link capacitors are used. The zero-sequence current path is provided by tying the neutral point to the middle point of the two DC link capacitors. [G1] presented a scheme using split DC link capacitors to handle the zero-sequence for active power filter application. The drawback of this scheme is that excessively large DC link capacitors are needed; therefore cost is high especially for high voltage applications. To handle the zero-sequence component, the four-legged inverter proposed in [E1] can substantially reduce the DC link capacitance. Since it is a soft-switching inverter, the correction of unbalance is achieved by a fast feedback control loop with a high switching frequency.

Modeling of three-phase three-legged power converters was presented in [H2] by representing the switching networks with controlled voltage and current sources. The "inplace" circuit averaging method was adopted to derive the average circuit model directly from the switching circuit model. Large-signal and small-signal models are derived in d-q rotating coordinate. The models are convenient for direct analysis and simulation with circuit simulation software. Effects of circuit parasitic are also discussed in [H2]. Due to three-legged topologies, the zero-sequence component is not revealed in the models. Small-signal models of PWM modulators are also presented in [H1]. It has been shown that a PWM modulator will add additional time delay, and the gain of the PWM modulator could change with respect to the modulation index and the vector position. Both the gain and phase delay of the PWM modulator cannot be expressed in a simple closed form.

Control of three-phase three-legged power converters has been extensively investigated. As with DC/DC converters, control for three-phase power converters can also be classified into voltage mode control and current mode control. Current mode control has been used widely for PFC rectifier, motor drive applications and high performance UPS due to its fast dynamic response and inherent current protection capability. Three independent hysteresis current controllers can provide fast current regulation. However, in unbalanced cases, the performance is degraded due to a continuous fight among the three independent hysteresis controllers. Deadbeat control was proposed to make the controlled variable equal to the reference at the end of each switching cycle so that a response within one discretigation time is obtained. Deadbeat control was used for both single-phase applications [H5] and three-phase applications [H6-H8]. It can be applied to both the current control loop and voltage control loop [H7-H8].

For hard-switching power converters, a dead time needs to be added for each switching commutation to prevent shoot-through problem. The dead time will cause duty ratio loss, and thus lead to an inaccurate control. The control inaccuracy may greatly degrade performance especially when the ratio of the dead time to the switching period is large. [H14] proposed a method to compensate the duty ratio loss caused by the dead time. It adjusts the pulse width by the amount of dead time according to the load current direction.

As Digital Signal Processor (DSP), microcontroller, and other peripherals such as A/D conversion and D/A conversion become faster and faster, the digital controller becomes more and more popular in power electronics [H9-H13] [H22]. A digital controller can easily realize a complicated control algorithm. It is insensible to parameter changes and temperature variations, and thus more robust. The drawback with a digital controller is an additional time delay caused by sensing, sampling and computation, which is a major limitation for extending control bandwidth. Predictive current control may help to reduce the delay to some extent [B1].

To achieve a high performance with nonlinear load, a low output impedance is needed. The output impedance may be reduced by (1) parallel power converters, or (2) a high control bandwidth. The parallel power converters have to deal with current sharing problem [D5-D14]. A high control bandwidth is always desired to achieve a compact and high performance system. The control bandwidth can be extended with a high switching frequency. Soft-switching power converters [C1-C14] enable the use of a high switching frequency by reducing switching losses. However, there is a trade-off in using soft switching power converters. A control inaccuracy may occur due to duty ratio losses caused by interventions of a soft-switching network.

Active filters were investigated in the past decades to deal with nonlinear loads in power systems. The basic principle was first proposed in [G5] in 1971, and was generalized in [G6] in 1976. Active filters can be classified into two categories: shunt (parallel) active filter and series active filter [G7]. Parallel active filters are controlled to be a current source, and used to deal with nonlinear loads with a harmonic current source characteristic; a series active filter is controlled to be a voltage source, and used to deal with nonlinear loads with a harmonic voltage source characteristic [G8]. Active filters can be used not only to handle the harmonics caused by nonlinear loads, but also to handle reactive power [G9-G10]. Design aspects of hard switching active filters were discussed in [G14-G16], while an example of a soft switching active filter was given in [G17].

There were three hybrid approaches in the active filter applications aiming at a high performance and cost-effective solution. First, a hybrid of a shunt active filter with a passive filter. The shunt active filter compensates low-order harmonic currents, and the passive filter compensates high-order harmonic currents [G7]; second, hybrid of a series active filter with a passive filter [G7], [G1-G12]. The series active filter does not compensate harmonic currents, it only isolates the harmonics for the passive filter to handle. This approach can eliminate the sensitivity of the passive filter to the power source impedance, and results in a low cost system; third, a hybrid of a shunt active filter with a series active filter. The so-called universal active can deal with nonlinear loads with both harmonic current and voltage source characteristics [G13].

Harmonic current detection is the major focus in designing an active power filter. Instantaneous reactive power theory was developed in [G19] to extract harmonic currents. Both three-phase voltage and current are sensed. If the three-phase voltage is known, instantaneous reactive power theory is the same as applying d-q transformation to the load current to extract the harmonic currents. Instantaneous reactive power theory is more appropriate for a balanced source. When three-phase systems are unbalanced, [G20] proposed three methods to calculate the current references by using a synchronous detection technique. Under an unbalanced source voltage, the current reference for the active power filter can be given to realize either of the following three: (1) equal power among phases; (2) equal current among phases; (3) equal resistance among phases. The source could be not only unbalanced, it could also be distorted due to heavy nonlinear loads, that makes the harmonic current detection even more difficult. It was also argued in [G21] that when the source voltage is heavily distorted under a heavy nonlinear load, the instantaneous reactive power theory does not give satisfactory results. It could even make the harmonic currents larger. A compensation strategy, which aims at making the nonlinear load/active filter system a constant linear resistor, is concluded to be a better choice in this situation. Another compensation concept suitable for nonlinear voltage and current can be found in [G22], where fictitious power was defined to include reactive power, effect of harmonic voltage and current and sub harmonic voltage and current. However, the computation is more complicated.

With the ever faster evolution of power electronics technologies, power converters tend to be connected to form a power converter system. Series and parallel connected power converters are the two basic forms of power converter systems. A series connected AC/DC PFC rectifier and a DC/AC inverter can provide a high performance motor drive system. Multiple power converters in parallel can provide high power capacity and N+1 redundancy. In recent years, more sophisticated power converter systems have been proposed. Among them, multi-functional power converters, passive circuit combined with active power converter, and active power converter combined with active power converter can be found. One example of a multi-functional power converter was given in [D17], where a three-phase voltage source inverter, used in a battery energy storage system and connected to a utility line, can perform charger, inverter and active filter functions. Hybrid approaches are mostly found in power conditioning applications. A combined system with a shunt passive filter and series active filter was presented in [G11-G12] to compensate harmonic currents in power systems. The series active filter isolates the harmonic currents from the power source so that the passive filter can be effective for harmonic currents. The passive-active hybrid system is more effective than a passive filter and practical due to its low cost. An active-active hybrid approach can be found in [G13], where a shunt active filter is combined with a series active filter. Since a shunt active filter is effective in dealing with nonlinear load with harmonic current source characteristics, and a series active filter is effective in dealing with nonlinear loads with a harmonic voltage source characteristics, the hybrid system renders an overall better performance. This kind of active-active power converter system is a hybrid of power converters for different loads. Another kind of active-active power converter system is presented in [D18-G20], which partitions the energy flow into two distinguished parts: high-power/low-frequency and low-power/high-frequency, and then handles them separately by two power converters. [D18] presented a high-power magnet power supply, where a low-frequency thyristor rectifier is used to handle the main output power, while a high-frequency PWM converter is in series with the thyristor rectifier to cancel harmonics and compensate for errors. [D19] presented a hybrid system with a similar concept for active power filter application. A GTO high-power low frequency inverter is combined with a low-power high-frequency IGBT inverter. [D20] was also for power conditioning applications. A high-power multi-step inverter was adopted to handle reactive power, and PWM inverters are adopted for harmonic compensation. The common essential aspect of a power converter system is the control strategy it employs. Since more than one power converter and/or loads are involved, the analysis and control design of the whole system can be extremely difficult. In designing a power converter system, special attention should also be paid to the circulating current at switching frequency [D1-D2] and grounding issue [D16].

Using power converters to perform active damping function was suggested in [H20] and [H21]. A distributed UPS system was presented in [H20]. The loop formed by UPS output filter capacitors and the interconnecting line inductance is highly under damped. The derivative of the interconnecting current was used as a current reference for each UPS to serve a damping purpose. No active power is processed to perform the damping function since only harmonic currents were included in the derivative. Simulation results showed that active damping can suppress the sub harmonic oscillation on the line. However, this approach is noise sensitive due to the derivative. A current type PWM rectifier was presented in [H21] with active damping function. The motivation was to damp harmonic currents amplified by the LC filter of the current type rectifier. There is no prior research on using active damping function for the purpose of extending the control bandwidth of the other power converter.

Large-scale power electronic systems have been analyzed for solar power systems and parallel UPS systems. System stability is of a major concern when many power converters are connected together. The forbidden region concept was developed for test of small-signal stability margin. A large-scale DC distribution power system with PFC rectifier, motor drive, utility power supply, and other power converters, has not been investigated before. There is no large-signal stability issue discussed for this large-scale power electronics system.

1.13 MOTIVATION FOR THE THESIS

The study for power quality problems highlights that the shunt active filter can compensate for current harmonics, improve power factor & control the reactive power flow from system to the load. The study also shows that the combined active filter systems have the same objectives as the independent series and shunt active filters. That is, the series active filter isolates the utility voltage disturbances and the shunt active filter provides compensation for reactive power, harmonies current, and unbalance in load. But elimination of voltage harmonics for low rating having imported linear load which is connected to same bus is not reported in the literature.

In addition, the study finds the following unresolved issues

The shunt active filter alone provides the load reactive power compensation, harmonics compensation so that bus voltage does not distort and affect other small but important linear load for this its ratings are very high.

Online as well as instantaneous detection of voltage harmonics so that it can be used as reference for VSI to compensate voltage harmonics

Online as well as instantaneous detection of voltage unbalance (negative sequence as well as zero sequence) so that it can be used as reference for VSI to compensate unbalance.

The motivation for the thesis stems from the fact that there is a need for a novel power quality tool to solve the problem of voltage harmonics & compensate negative sequence and zero sequence voltage which is the normal power quality issue in real world to reduce the ratings of the shunt active filter.

In this thesis, a series active filter, negative and zero sequence voltage compensator using Instantaneous Active Reactive power Theory (IARP Theory) is presented. Instantaneous Active Reactive power Theory (IARP Theory) is used for shunt active filter for current harmonics elimination & reactive power compensation but in this thesis IARP theory is modified in such a way so that it can detect online, instantaneous voltage harmonics, negative sequence and zero sequence voltage.

1.14 ORGANIZATION OF THE THESIS

The thesis is organized into six chapters. The outlines of these chapters are as follows:

Chapter 1 gives overview on power quality & its effect.

Chapter 2 is devoted to the different tools or methods for improvement of power quality. In this chapter different methods along with control theory for improvement of power quality are discussed

For improving reliability & quality of power following methods are discussed.

a) Different type of passive filter & its analysis along with its advantage & disadvantage
b) Different type of shunt active filter, their control, simulation as well as experimental results
c) Unity power factor boost converter along with control, simulation & experimental results

Chapter 3 deals with series active filter which not only isolate the linear load from distortion but also improve the voltage profile of the bus. Instantaneous Active Reactive power Theory (IARP Theory) which is used for detection of current is modified so that it can be used to detect the voltage harmonic of the system. Extensive simulation results along with experimental results were presented.

Main contribution of this chapter is

a) Active series filter to eliminate voltage harmonics for linear load using modified IARP theory.
b) Extensive simulation results along with experimental results were presented. Experiments were done at 415 volts, 3 phase systems.
c) Experiments were done at 415 volts, 3 phase systems
d) Analogue based control systems was discussed along with it power circuit.

Chapter 4 illustrated new techniques to detect negative sequence & zero sequence voltage and compensate the same so that load experience the balance voltage profile. Same Instantaneous Active Reactive power Theory (IARP Theory) is used in different manner so that desired quantity i.e. negative sequence as well as zero sequence can be detected. This signal will be used as reference signal for SPWM inverter so that it will compensate the system voltage profile for unbalance. Main contributions of this chapter are:

a) Novel technique for online, instantaneous detection of negative sequence & zero sequence components
b) Using this techniques a novel power quality improvement tool which can compensate for unbalance in the systems bus voltage

Chapter 5 describe about combine harmonic & negative sequence compensator which is useful where for unbalance voltage having distortion. In this chapter, with slide modification in IARP theory used for series active filter this novel combine series compensator is simulated & simulation results are presented.

Chapter 6 lists the important contributions of this thesis work and identifies the future challenges that need to be addressed.

1.15 PROBLEM STATEMENTS

The system impedance plays an important role in distorting the bus voltages due to presence of harmonics in the line currents. As the system impedance increases, there are more chances of generating distortion in the voltage bus. It is observed in practice that the bus voltages get distorted due to present of current harmonics. Sensitive load being fed from such a distorted bus are liable to malfunction. One of the examples, being the railways signaling load, where in the signaling is affected due to the bus distortion which is generated by rectifier load connected on same bus. This signaling load is of fraction capacity compared to total load on the bus. In such cases it is economical to use a series active filter to filter out the bus voltages feeding a sensitive load instead of the eliminating current harmonics in the bus.

Due to unbalance loads connected to the bus having higher impedance, there is unbalance generated in bus voltages. This unbalance results in generation of zero and negative sequence voltages. 80% of the industrial loads are motor loads. These negative sequence voltages generate negative torque which in turn decreases the effective output torque of the motor and hence the efficiency of the drive system. The zero sequence voltages produce heat loss in the motor.

The work presented here deals with a voltage source converter, configured as a series active filter, which not only eliminates the voltage harmonics but also compensates for the negative sequence voltages and the zero sequence voltages of the bus. It also improves the voltage profile of the system.

CHAPTER-II

PASSIVE & ACTIVE FILTER

TABLE OF CONTENTS

2.1 INTRODUCTION

Over the year there has been a continuous proliferation of nonlinear type of loads due to intensive use of power electronic control in all branches of industry as well as by the general consumers of electric energy. These modern equipments are designed to offer optimum performance at low running costs, but in turn they play havoc with the supply system and hence affect the performance of other equipment connected in the system. These disturbances include harmonic distortion, voltage unbalance, voltage surges, increased reactive power demand and power system fluctuations etc. Harmonic contamination has become a major concern of power system specialist due to its effects on sensitive loads and on power distribution system. Harmonic current component increase power system losses, causes excessive heating in rotating machinery, can create significant interference with the commutation circuits that shared common right of ways with AC power lines, can generate noise on regulating and control circuits causing erroneous operation of such equipment. The effect of voltage and current harmonics can be noted at far of places in equipment connected to the same circuit.

2.2 HARMONIC SOURCES:

Two types of harmonics exist in the AC system voltage and current waveforms, characteristic harmonics and non-characteristic harmonics. Characteristic harmonics are generated due to characteristic of the harmonic producing equipment or subsystem and magnitude of these harmonics is much higher compared to magnitude of non-characteristic harmonics. The magnitude of characteristic harmonics decreases with increase in the harmonic order. Non-characteristic harmonics are generated due to all sorts of asymmetries in various parameters of the subsystem and the AC system.

Following equipment or subsystems are source of harmonics in the AC system. These sources of harmonics are already discussed in detail in chapter-I.

- Thyristor controlled devices, e.g.
- Rectifier and inverter stations of HVDC scheme.
- Thyristor Controlled Reactor (TCR)
- Thyristor Switched Capacitor (TSC)
- Cyclo-Converters
- Non-linear loads, e.g.
- Induction furnace and arc furnace
- Steel mills and rolling mills
- Welding machine
- Single phase uncompensated railways loads
- Switching equipment and various electronic circuits.
- Different type of loads at low level, e.g. house hold equipment and computers
- Non-linearity and core saturation of transformers and reactors.
- Combined effect of various harmonic producing equipment and subsystems.
- Cross modulation across the converters.

Even though magnitude of non-characteristic harmonics are very small, however parallel resonance within the AC system at any harmonic frequency can increase the magnitude significantly.

2.3 DIFFERENT METHODS OF DEALING WITH POWER QUALITY:

Harmonic present in the AC network causes following problems in the power system equipment.

1) Overheating in transformers and reactors
2) Overheating in capacitor bank and rotating machines
3) Additional losses in the transmission lines
4) Telephonic noise due to higher order harmonics
5) Higher stresses on the circuit breaker
6) Malfunctioning of control and protection equipment because of displacement of the zero crossing and increase in peak voltages
7) Third harmonic, which is of zero phase sequence, can flow through neutral wires into electricity distribution network causing overheating in the neutral wires.

As an alternative to harmonic reduction, it has been suggested to derate transformers and oversize cables. This could be an adequate solution in some cases, but since system configurations change, and thereby the harmonic profile of the system, the required amount of derating or over sizing could be difficult to predict. Also, this option does not prevent harmonies from entering the supply system. Therefore, a more appropriate method for harmonic mitigation is the use of filters.

Harmonic injected by the consumer loads into the AC system are required to be limited to the acceptable values. The simplest means of limiting harmonic at the AC bus is by installing harmonic filter at the load itself where harmonics are generated. Installation can be done either by the consumer or by the utilities.

Harmonic filters in power systems are used primarily to mitigate the harmonics generated by various kind of non-linear load. Depending on the configuration, filters are either of passive type or active type.

Conventionally, passive LC filters have been used to eliminate line current harmonics and to increase the load power factor.

2.4 PASSIVE FILTERS:

Filters are either series connected or shunt connected in the AC system. The concept of series connected filter is parallel resonating electrical circuit, which offers very high impedance at tuning frequencies. The high impedance offered by the filter allows very little harmonics to pass through it. The disadvantage of this type of connection is that all the filter components are required to be rated for full line current, which makes installation very expensive.

Most commonly the filters are shunt connected to the AC system. This type of filters use series resonating electrical circuit offering negligible impedance compared to the AC system harmonic impedance at tuning frequencies. The low impedance path, provided by the filters, attract major portion of the harmonics and allows very small portion of the harmonics to flow into the AC systems. Components in the shunt connected filter branch are designed for graded insulation levels, which make component cheaper than those used for series connected filters.

Combination of series and shunt-connected branch is used in the design of Power Line Carrier (PLC) and Radio Interference (RI) filters. The series connected branch blocks the harmonics and the shunt-connected branch allows the harmonics to flow in to the ground. The combination cannot be used for low order harmonics because high blocking impedance at low order tuning frequency will have significant voltage drop at fundamental frequency as well, which will reduce the AC bus voltage.

2.4.1. TYPES OF PASSIVE FILTERS:

Passive filters are series resonating or parallel resonating electrical circuit, which offer very high or very low impedance at tuning frequency. The filters are resistive at tuning frequency, capacitive below tuning frequency and inductive beyond tuning frequency. Filters have two important characteristics: impedance and bandwidth. Low impedance is required to ensure that harmonic voltages have a low magnitude and certain bandwidth is needed to limit the consequences of filter detuning.

2.4.1.1 SINGLE TUNED FILTER:

Single tuned filters as name suggests, are tuned to only one frequency and are simplest of all filters.

CONFIGURATION:

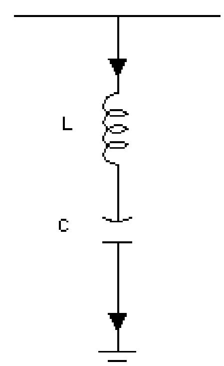


Figure 2.4.1.1-1: Single Tuned Filter

ADVANTAGE:

1) Simple configuration with only two components, capacitor and reactor.
2) Quality factor of the filter is high which provides maximum attenuation of one harmonic
3) Negligible losses as there are no resistor for damping etc.
4) Low maintenance requirements because of fewer components.

DISADVANTAGE:

1) Many filters will be required to filter several harmonics since one filter can be used for one harmonic only
2) High quality factor of the filter gives low bandwidth, which makes filter sensitive to variations in the fundamental frequency as well as the component values.
3) Accurate tuning is required at site because of which provision of taps on the reactor is essential. This increases the cost of the reactor.

FORMULAE FOR COMPONENT VALUE CALCULATION:

$$C_H = \frac{Q}{\left(V^2 * 2\pi f\right)} \qquad (2.4.1.2\text{-}1)$$

$$L = \frac{1}{\left\{\left(2\pi f_r\right)^2 * C_H\right\}} \qquad (2.4.1.2\text{-}2)$$

$$R = q * 2\pi f_r * L \qquad (2.4.1.2\text{-}3)$$

$$C_L = \frac{1}{\left\{\left(2\pi f\right)^2 * L\right\}} \qquad (2.4.1.2\text{-}4)$$

Where Q= reactive power to be generated by the filter at fundamental frequency (assumed)

V=voltage level at which filters are to be installed

f= Fundamental frequency

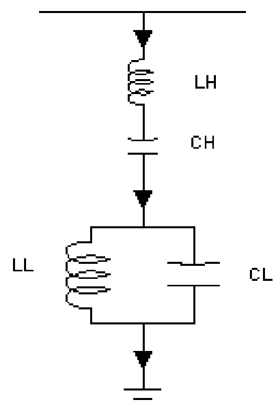
f_r= Tuning frequency (assumed)

q= Quality factor of the filter (assumed)

2.4.1.2 DOUBLE TUNED FILTERS :

Double tuned filters are equivalent to two single tuned filters connected in parallel. These filters use series as well as parallel resonating circuits, which makes it complex. C_H & L_H correspond to first tuning frequency and high voltage section whereas C_L& L_L corresponds to second tuning frequency and low voltage section.

CONFIGURATION:



Double Tuned Filter

Figure 2.4.1.2-1: Double Tuned Filter

ADVANTAGES:

1) Quality factor the filter is high which provides maximum attenuation of two harmonic
2) Negligible losses as there are no resistors for damping etc.
3) There is only one high voltage capacitor and high voltage reactor because of which it is cheaper than two single tuned filters connected in parallel.

DISADVANTAGES:

1) High quality factor of the filter give low bandwidth, which makes filter sensitive to variations in the fundamental frequency as well as the component values.
2) Accurate tuning is required at site because of which provision of taps on the reactor is essential. This increases the cost of the reactor.
3) Rating of low voltage component is decided mainly by transient behavior of the filter circuit.

FORMULA FOR COMPONENT VALUE CALCULATION

$$C_H = \frac{Q}{\left(V^2 * 2\pi f\right)} \tag{2.4.1.3-1}$$

$$L_H = \frac{1}{\left\{(2\pi f_1)^2 * C_H\right\}} \tag{2.4.1.3-2}$$

$$C_L = \frac{1}{\left\{(2\pi f_L)^2 * L_L\right\}} \tag{2.4.1.3-3}$$

Where Q= reactive power to be generated by the filter at fundamental frequency (assumed)

V=voltage level at which filters are to be installed

f= Fundamental frequency

f_1= First tuning frequency (assumed)

f_2= Second tuning frequency (assumed)

The filter circuit being complex in nature, the component values are finalized only after analysis of impedance versus frequency plots. Adjustment in the component values are made based on the desired filter characteristics.

2.4.2. GUIDELINES FOR PASSIVE FILTER SELECTION:-

The choice of techno-economical filter circuit depends on site conditions and AC system parameters. Detailed performance and rating studies only can establish an optimum solution. From the advantages and disadvantages described in section 2.4.1 following can be used for selection of filters

Damped filters are not very sensitive to frequency variations. When variation in the AC system fundamental frequency is large then damped filter shall be preferred to tuned filters.

1) Combination of damped filters can be used for limiting telephonic interference problem
2) Double tuned and triple tuned filters can be used for economical solutions in high voltage low MVAR filters.
3) For accurate tuning at sites, taps on the reactor shall be specified.
4) Where there is a wide ambient temperature range, tuned filters may not be right choice. However if seasonal tuning is permitted then taps on the reactor can be used for returning in summer and winter conditions.
5) Second order-damped filters can provide the optimum solution for characteristic harmonic groups.
6) Low harmonic order filter shall be realized either tuned or damped filters.
7) When limitation of voltage distortion can achieved by either tuned or damped filters.
8) When limitation on reactive power exchange in conjunction with TIF limitation is required double tuned high pass filters may be the optimum solution.

However, in practical application these passive filters present following disadvantages.

1) The source impedance strongly affects filtering characteristics.
2) As both the harmonics and the fundamental current components flow into the filter, the capacity of the filter must be rated by taking into account both currents.
3) When the harmonic current components increase, the capacity of the filter can be overloaded.
4) Parallel resonance between the power system and the passive filter causes amplification of the harmonic currents on the source side at a specific frequency.
5) The passive filter may fall into series resonance with the power system so that voltage distortion produces excessive harmonic currents flowing into the passive filter.

In order to overcome these problems, active filters have been researched and developed. Since then basic compensation were proposed around 1970 much research has been done on active filters and their practical application [I1]-[I3]. In addition, state of art power electronics technology has enabled engineers to put active filters into practical use.

2.5 ACTIVE FILTERING

Passive filters can prevent harmonics from entering a supply system, and are also useful in increasing power factor, but there are a number of considerations that are critical in this approach [10] [11]. For instance parallel and series resonance with ac line impedance may produce amplification in the harmonic current and voltage of the line at certain frequencies. Also the effectiveness in attenuating harmonics is not dictated only by the passive filter itself, but depends on the source impedance. Moreover, it is difficult to decouple the effects of one load from those of other loads connected to the line and compensate only for harmonic currents produced by this one load. Thus, it is often necessary to oversize passive filters in order to avoid overheating. The classic solution of tuned passive filters therefore has serious limitations. Furthermore, with the widespread use of variable distorting electronic loads, harmonic tracking is not adequately achieved. Also the probability of encountering resonant conditions increases with the number of installed filtering units. Active filtering techniques have therefore been proposed to overcome the problems associated with passive filters.

Much research has been performed on active filters for power conditioning and their practical applications since their basic principles of compensation were proposed around 1970 (Bird et al., 1969; Gyugyi and Strycula, 1976; Kawahira et al., 1983). In particular, recent remarkable progress in the capacity and switching speed of power semiconductor devices such as insulated-gate bipolar transistors (IGBTs) has spurred interest in active filters for power conditioning. In addition, state-of-the-art power electronics technology has enabled active filters to be put into practical use. More than one thousand sets of active filters consisting of voltage-fed pulse-width-modulation (PWM) inverters using IGBTs or gate-turn-off (GTO) thyristors are operating successfully in Japan.

Active filters for power conditioning provide the following functions:

1) Reactive-power compensation
2) Harmonic compensation,
3) harmonic isolation,
4) harmonic damping and harmonic termination,
5) negative-sequence current and voltage compensation,
6) Voltage regulation.

The term ‘‘active filters’’ is also used in the field of signal processing. In order to distinguish active filters in power processing from active filters in signal processing, the term ‘‘active power filters’’ is used in normal practice.

2.5.1. CLASSIFICATION OF ACTIVE FILTER:

Various types of active filters have been proposed in many technical literatures. Classification of active filter is made from different point view. Active filters are divided into AC and DC filters. Active DC filters have been designed to compensate for current and/or voltage harmonics on the DC side of thyristor converters for HVDC systems [4] and on the DC link of a PWM rectifier/inverter for traction system. But the term "active filter" refers to the AC filters in most cases.

2.5.1.1 CLASSIFICATION BY SYSTEM CONFIGURATION

2.5.1.1.1 Shunt active filters:

Figure 2.5.1-1 shows a system configuration of a shunt active filter used alone, which is one of the most fundamental system configurations. The shunt active filter is controlled to draw a compensating current i_{AF}, from the utility, so that it cancels current harmonics on the AC side of a general-purpose thyristor rectifier with a DC link inductor or a PWM rectifier with DC link capacitor for traction systems. The shunt active filter has the capability of damping harmonic resonance between an existing passive filter and the supply impedance.

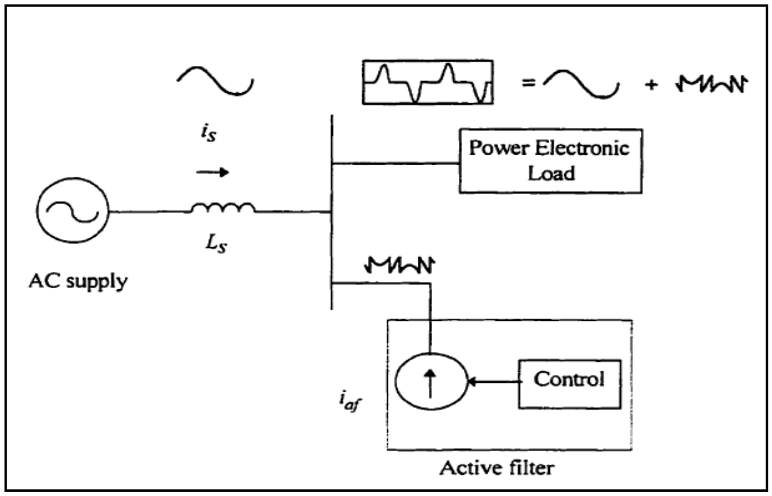


Figure 2.5.1-1: Shunt Active Filter

2.5.1.1.2 Series active filters:

Figure 2.5.1-2 shows a system configuration of a series active filter used alone. The series active filter is connected in series with a utility through a matching transformer, so that it is applicable to harmonic compensation of a large capacity diode rectifier with a DC link capacitor. Comparisons between the shunt and series active filters show that series active filter has a "dual" relationship in each item with the shunt active filter.

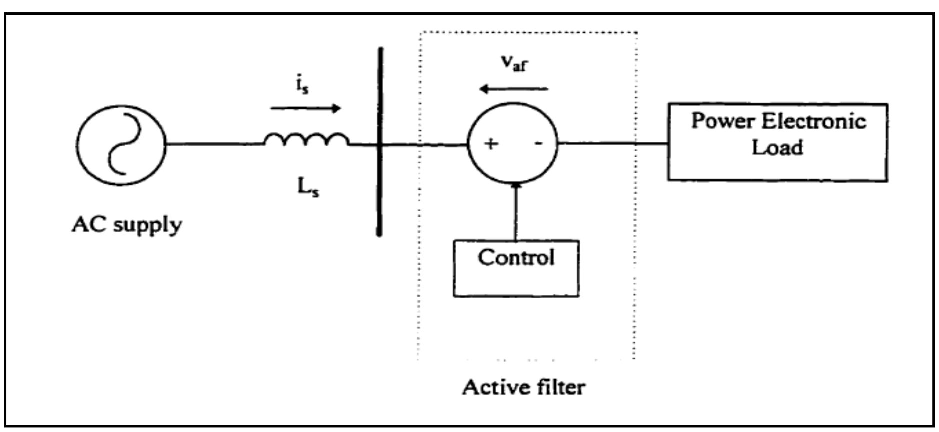


Figure 2.5.1-2: Series Active Filter

Comparison of shunt and series active filter used alone:

	Shunt active filter	**Series active filter**
System configuration	Figure 2.5.1-1	Figure 2.5.1-2
Power circuit of	Voltage-fed PWM inverter with	Voltage-fed PWM inverter
Active filter acts as	Current source: i_{af}	Voltage source: V_{af}
Harmonic producing	Cyclo-Converters & thyristor	Diode rectifiers with
Additional function	Reactive power compensation	AC voltage regulator
Present situation	Commercial stage	Laboratory level

2.5.1.1.3 Hybrid Active Filter:

(1) Hybrid active/passive filters:

Figure 2.5.1-3 to Figure 2.5.1-5 show three types of hybrid active/passive filters, the main purpose of which is to reduce initial costs and to improve efficiency. The shunt passive filter consists of one or more tuned LC filters and/or a high-pass filter. The difference among the three hybrid filters, in which, the active filters are different in function from the passive filters.

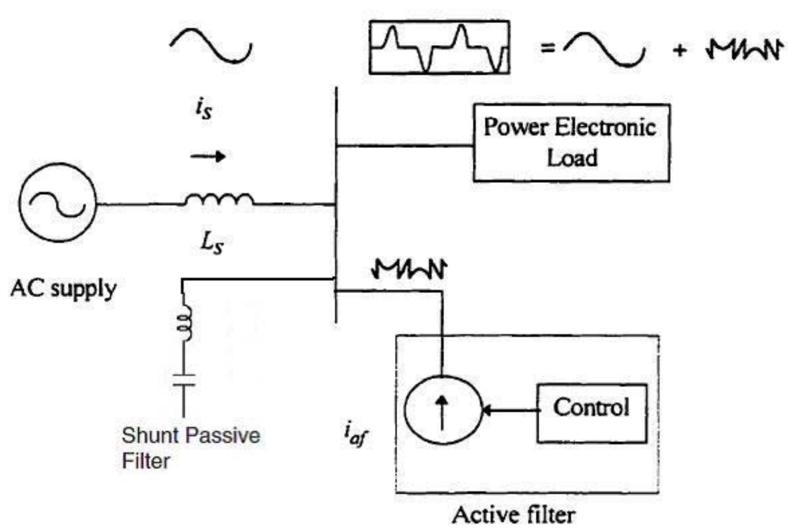


Figure 2.5.1-3: Combination of shunt active filter and shunt passive filter

(2) Hybrid shunt/series active filters:

The combination of shunt and series active filter will be applied in near future, not only for harmonic compensation but also for harmonic isolation between supply and load, and for voltage regulation and imbalance compensation. They are considered prospective alternatives to shunt or series active filters used alone.

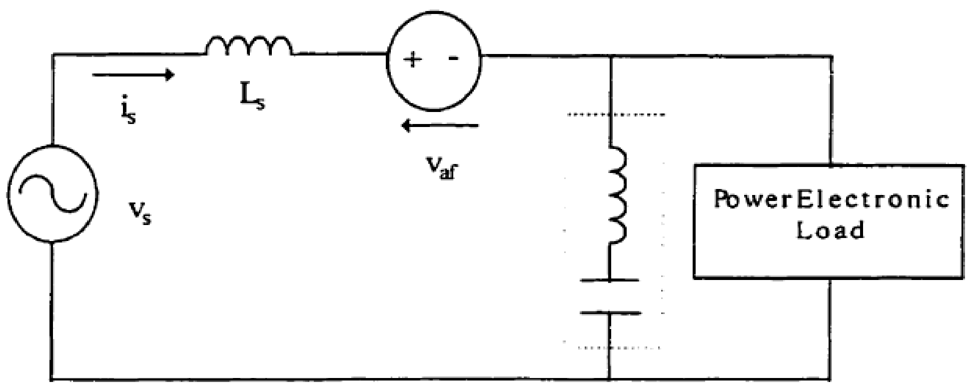


Figure 2.5.1-4: Combination of series active filter and shunt passive filter

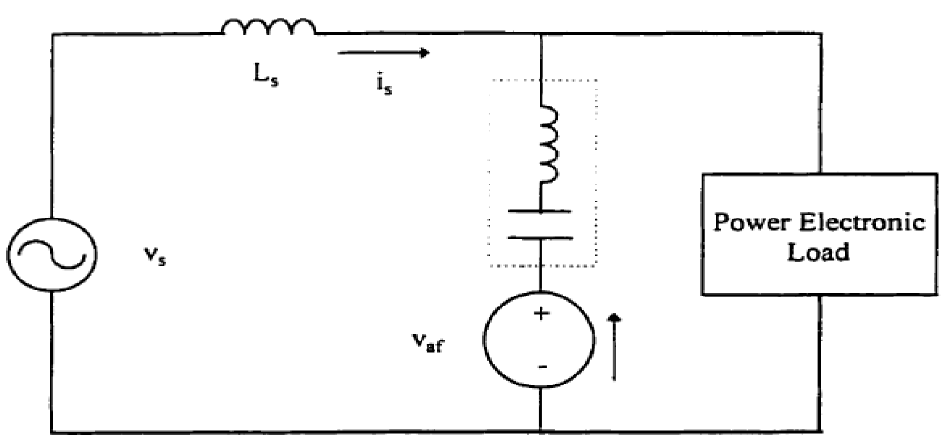


Figure 2.5.1-5: Combination of series active filter connected in series with and shunt passive filter

Comparison of hybrid active /passive filters

	Shunt active filter plus shunt passive filter	**Series active filter plus shunt passive filter**	**Series active filter connected in series with shunt passive filter**
System configuration	Figure 2.5.1-3	Figure 2.5.1-4	Figure 2.5.1-5
Power circuit of active filter	Voltage-fed PWM inverter with current minor loop	Voltage-fed PWM inverter without current minor loop	Voltage-fed PWM inverter with or without current minor loop
Function of active filter	Harmonic compensation or harmonic damping	Harmonic isolation and harmonic damping	Harmonic damping or harmonic compensation
Advantages	General shunt active filters applicable Reactive power controllable	Already existing shunt passive filters applicable No harmonic current flowing through active filter	Already existing shunt passive filters applicable Easy protection of active filter
Problem or issues	Share compensation in frequency domain between active filter and passive filter	Difficult to protect active filter against over current No reactive power control	No reactive power control
Present situation	Commercial stage	Field testing	Coming into market

2.5.1.2 CLASSIFICATION BY POWER CIRCUIT:

There are two type of power circuit used for active filters a voltage-fed PWM inverter and a current fed PWM inverter. These are similar to the power circuit used for AC motor drives. They are, however, different in their behavior because active filters act as non-sinusoidal current or voltage source. The voltage fed PWM inverter is preferred compare to the current fed PWM inverter because the voltage fed PWM inverter is higher in efficiency and lower in initial costs than the current fed PWM inverter. In fact, almost all active filters, which have been put into practical applications, have adopted the voltage fed PWM inverter as the power circuit.

2.5.1.3 CLASSIFICATION BY CONTROL STRATEGY:

The control strategy of active filters has great impact not only on the compensation objective and required kVA rating of active filters, but also on the filtering characteristics in transient state as well as in steady state.

2.5.1.4 FREQUENCY DOMAIN AND TIME DOMAIN:

There are mainly two kinds of control strategies for extracting current or voltage harmonics from the corresponding distorted current or voltage; one based on frequency domain i.e.“Fourier analysis” and the other is based on time domain i.e. “instantaneous reactive power theory” in three phase circuits, which is called " p-q theory",. “synchronous reference theory” and “sine multiplication theory”. The concept of the p-q theory in the time domain has applied to the control strategy of almost all the shunt active filters installed by individual high power consumer.

2.5.1.5 HARMONIC DETECTION METHODS:

Three kinds of harmonic detection methods in the time domain has been proposed for shunt active filters acting as a current source i_{AF}

load current detection $i_{AF} = i_{Lh}$

supply current detection $i_{AF} = Ks * i_{Sh}$

voltage detection $i_{AF} = Kv * v_h$

load current detection and supply current detection are suitable for shunt active filters installed in the vicinity of one or more harmonic producing loads by individual high-power consumer. Voltage detection is suitable for shunt active filter to be used as shunt device of the "unified power quality conditioner" which will be installed in primary distribution substations by utilities.

Supply current detection is the most basic harmonic detection method for series active filters acting as a voltage source v_{AF}

Supply current detection $v_{AF} = G * i_{SH}$

Series active filter are based on supply current detection. It is also suitable for a series active filter to be used as the series device of the unified power quality conditioner.

2.6 PULSE WIDTH MODULATION SCHEMES:

Pulse width modulation (PWM) techniques have been the subject of intensive research during the last few decades. A large variety of methods, different in concept and performance, have been newly developed and described. However, the main objective is to shape the output AC voltages to be as close to a sine wave as possible and hence to reduce current harmonics and to improve the harmonic spectrum. However while achieving the above, the switch utilization factor, losses and complexity of control is also kept in mind and is tried to be optimized.

The voltage source converter type of inverters can be divided into following two general categories.

1) Pulse-width modulated inverters. In these inverters, the DC voltage is essentially kept constant in magnitude, where a diode rectifier is used to rectify the line voltage. Therefore, the inverter must control the magnitude and the frequency of the AC output voltages. This is achieved by PWM of the inverter switches and hence such inverters are called PWM inverters. There are various schemes to pulse-width modulate the inverter switches in order to shape the output AC voltages to be as close to a sine wave as possible. Out of these various PWM schemes, a scheme called sinusoidal PWM and Space vector modulation (SVM) are the most popular.
2) Square-wave inverters. In these inverters, the input DC voltage is controlled in order to control the magnitude of the output AC voltage, and therefore the inverter has to control only the frequency of the output voltage. The output AC voltage has a waveform similar to a square wave, and hence these inverters are called square-wave inverters.

2.6.1. SQUARE WAVE OPERATION IN THREE-PHASE INVERTERS

If the input DC voltage V_d is controllable, the inverter in Figure 2.6.1-1a can be operated in a square-wave mode. Here, each switch is on for 180° (i.e., its duty ratio is 50%). Therefore, at any instant for time, three switches are on.

In the square-wave mode of operation, the inverter itself cannot control the magnitude of the output AC voltages. Therefore, the DC input voltage must be controlled in order to control the output in magnitude. Here, the fundamental-frequency line-to-line rms voltage component in the output can be obtained as,

$$V_{LL_{1(rms)}} = \frac{\sqrt{3}}{\sqrt{2}}\frac{4}{\pi}\frac{V_d}{2} \tag{2.6.1-1}$$

$$= \frac{\sqrt{6}}{\pi}V_d \tag{2.6.1-2}$$

$$= 0.78V_d \tag{2.6.1-3}$$

The line to line output voltage waveform does not depend on the load and contains harmonics ($6n \pm 1$; n = 1, 2 …), whose amplitudes decrease inversely proportional to their harmonic order, as shown in Figure 2.6.1.1c:

$$V_{LL_h} = \frac{0.78}{h}V_d \tag{2.6.1-4}$$

where $h = 6n \pm 1$ (n = 1, 2, 3 …)

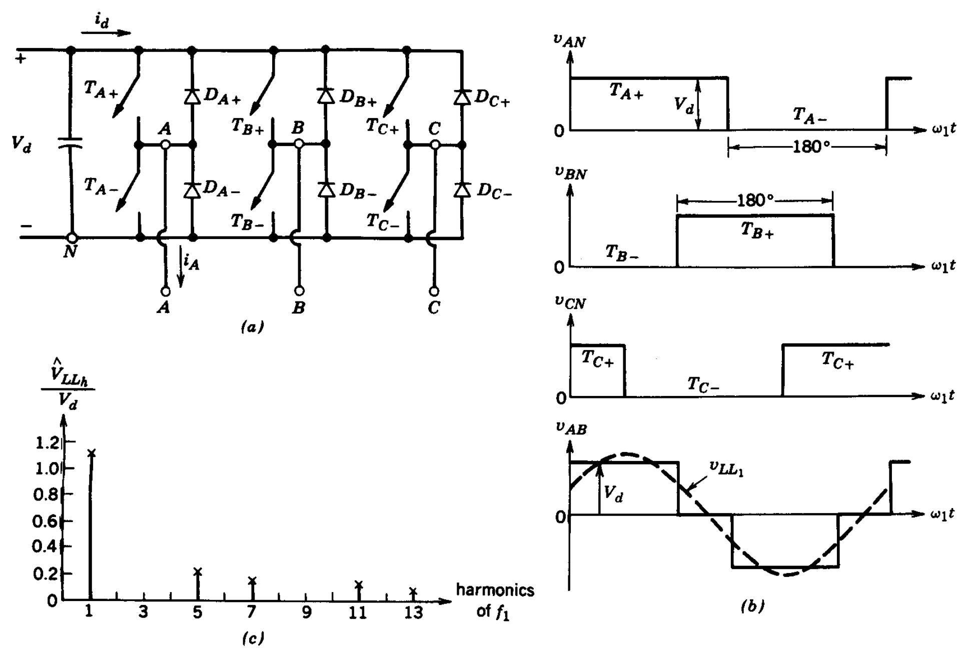


Figure 2.6.1-1: Square Wave inverter

2.6.2. SINUSOIDAL PULSE WIDTH MODULATION

To obtain balanced three-phase output voltages in a three-phase PWM inverter, triangular voltage waveform is compared with three sinusoidal control voltages that are 120° out of phase, as shown in Figure 2.6.2-1a.

It should be noted from Figure 2.6.2-1b that an identical amount of average DC component is present in the output voltages v_{AN} and v_{BN}, which are measured with respect to the negative DC bus. These DC components are cancelled out in the line-to-line voltages, for example in v_{AB} shown in Figure 2.6.2-1b. This is similar to what happens in a single-phase full-bridge inverter utilizing a PWM switching.

In general, the harmonics in the inverter (sinusoidal PWM control) output voltage waveform appears as sidebands, centered around the switching frequency and its multiples, that is, around harmonics m_f, $2m_f$, $3m_f$, and so on. Where m_f is the frequency modulation ratio. This general pattern holds true for all values of m_a (modulation index) in the range 0-1. Theoretically, the frequencies at which voltage harmonics occur can be indicated as,

$$f_h = (jm_f \pm k)\ f1$$

that is, the harmonic order *h* corresponds to the k^{th} sideband of *j* times the frequency modulation ratio m_f.

$$h = j\ (m_f)\ \pm k$$

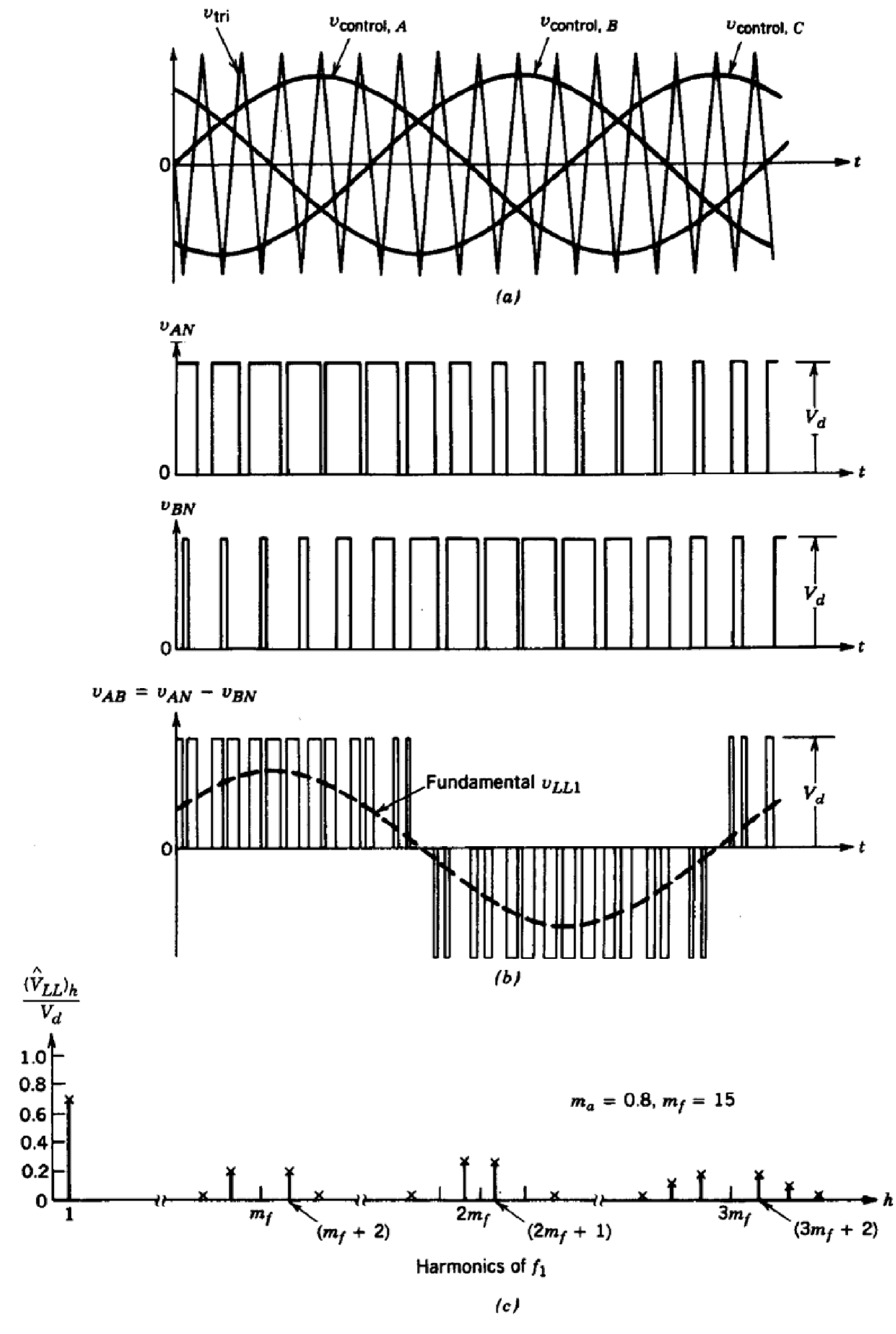


Figure 2.6.2-2: Three Phase PWM Waveforms and harmonics Spectrum

where the fundamental frequency corresponds to h=1. For odd values of *j*, the harmonics exist only for even values of *k*. For even values of *j*, the harmonics exist only for odd values of *k*.

However in three-phase inverters, only considering the harmonics at m_f (the same applies to its odd multiples), the phase difference will be equivalent to zero (a multiple of 360°) if m_f is odd and a multiple of 3. As a consequence, the harmonics at m_f is suppressed in the line-to-line voltage v_{AB}. The same argument applies in the suppression of harmonics at the odd multiples of m_f to be an odd multiple of 3 is to keep m_f odd and, hence, eliminate even harmonics). Figure 2.6.2-2d shows the variation of line-to-line output voltage (normalized with respect to DC link voltage v_d) of three sinusoidal pulse width modulated inverter for different values of modulation index. Figure also shows the linear and over modulation operating region.

If modulation index is defined as,

$m = \dfrac{u^*}{u_{1six-step}}$ where u^* = normalized fundamental reference vector

and $u_{1six-step}$ is the fundamental voltage of a six-step waveform.

Then, in case of sinusoidal pulse width modulation the maximum value of the modulation index, m = 0.785, is reached at a point where the amplitudes of the reference and the carrier signal become equal. However a distorted reference waveform, containing only triplen harmonics to an extent that its maximum assumes a flattop like shape as shown in Figure 2.6.2-2d increases the maximum modulation index, m=0.907. The added triplen harmonics modify the switched voltage waveform, increasing their fundamental content; they do not produce harmonic currents in a three-phase system.

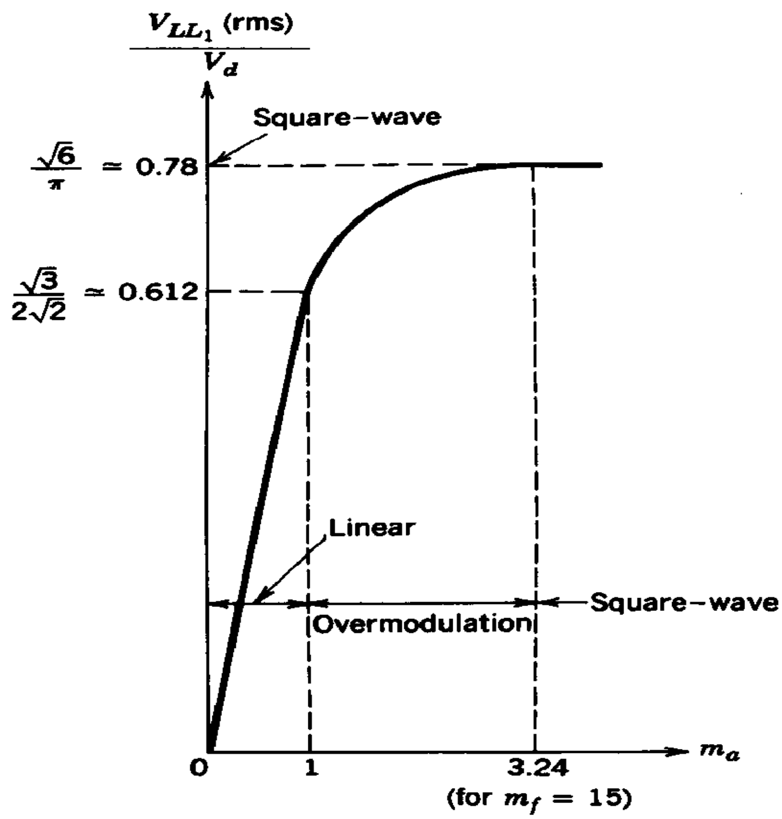


Figure 2.6.2-3: Three Phase Inverter V_{LL1}(rms)/V_d as a function of m_a

2.7 PWM CONVERTER FOR THREE PHASE FOUR WIRE ACTIVE FILTER SYSTEM:

In this section two configurations of voltage source inverters (VSI), which can be used in three-phase four wire systems are discussed. The fundamental difference between the VSI of Figure 2.7-1 and Figure 2.7-1 is the number of power semiconductor device.

A conventional three leg converter is used in Figure 2.7-1 and the AC neutral wire are connected directly to the mid-point of the DC bus, while in Figure 2.7-2 the AC neutral is provided through a fourth leg. Since the configuration has PWM current control, they behave as controlled current source. The AC current generated by the VSI has some high-order harmonics at the switching frequency, which can be easily filtered using a small passive filter (R and C in Figure 2.7-1. Ideally the currents track accurately their reference i_{ck}^* (k = a, b, c).

The controllability of the "four switch-leg" inverter topology Figure 2.7-2 is better than the "split-capacitor" inverter topology Figure 2.7-1. However the conventional three-leg converter is preferred because of its lower number of power semiconductor devices. The problems related to the DC capacitor voltage control by using the topology of "split-capacitor" will be discussed below.

Figure 2.7-3 shows a typical motion of the a-phase VSI current, controlled by hysteresis-based PWM current controller. If the current reference i_{ck}^* are assumed to be composed from zero sequence component, the line current i_{fk} (k = a, b, c) will return through the AC neutral wire. This forces, in the "split-capacitor" inverter topology, the current of each phase to flow either through C1 or through C2 and to return through the AC neutral wire.

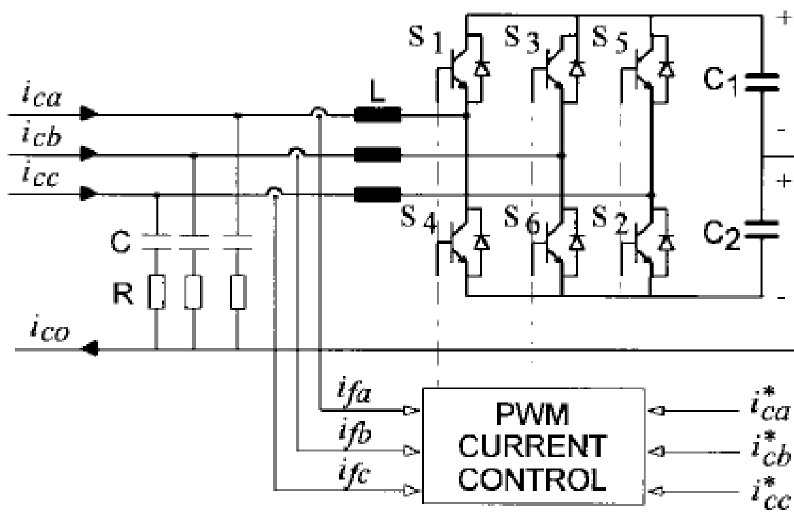


Figure 2.7-1: Three Phase Four Wire Three Leg Converter

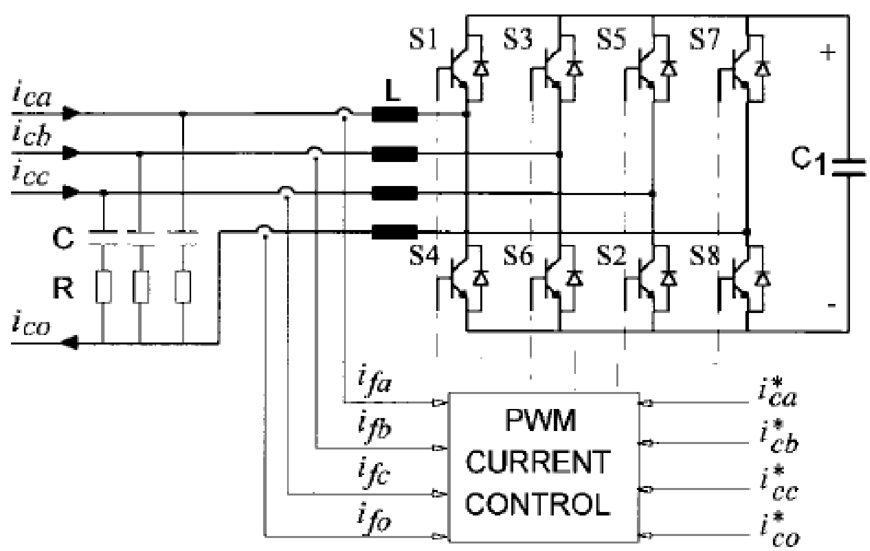


Figure 2.7-2: Three Phase Four Wire Four Leg Converters

The current can flow in both directions through the switches and capacitors. Table 2.7-1 summarizes the conditions that cause voltage variations in the capacitor C_1 and C_2 for a zero sequence current reference in the "split-capacitor" inverter topology.

Table 2.7-1: Conditions for the capacitor voltage Variation V_{c1} and V_{c2}

$i_{fk} > 0$ and $di_{fk} / dt < 0$	Increase the voltage in C_1
$i_{fk} < 0$ and $di_{fk} / dt < 0$	Decrease the voltage in C_1
$i_{fk} < 0$ and $di_{fk} / dt > 0$	Increase the voltage in C_2
$i_{fk} > 0$ and $di_{fk} / dt > 0$	Increase the voltage in C_2

When $i_{fk} > 0$, V_{c1} rises and V_{c2} decreases, but not with equal ratio because positive and negative values of di_{fk} / dt are different and depend on the instantaneous values of the AC phase voltages. The inverse occurs when $i_{fk} < 0$ the DC voltage variation depends also on the shape of the current reference and the hysteresis bandwidth.

Therefore, the total DC voltage as well as the voltage difference (V_{c2}-V_{c1}) will oscillates not only at the switching frequency, but also at the corresponding frequency of i_0 that is being generated by the VSI.

In the example given in Figure 2.7-3 the phase current i_{fa} causes voltage variations such that at the end of the period the voltage V_{c1} is higher and Vc2 lower. If a dynamic offset level is added to both limits of the hysteresis band, it is possible to control the capacitor voltage difference and to keep it within an acceptable tolerance margin. For instance, a negative offset in Figure 2.7-3 counteracts the above voltage variation

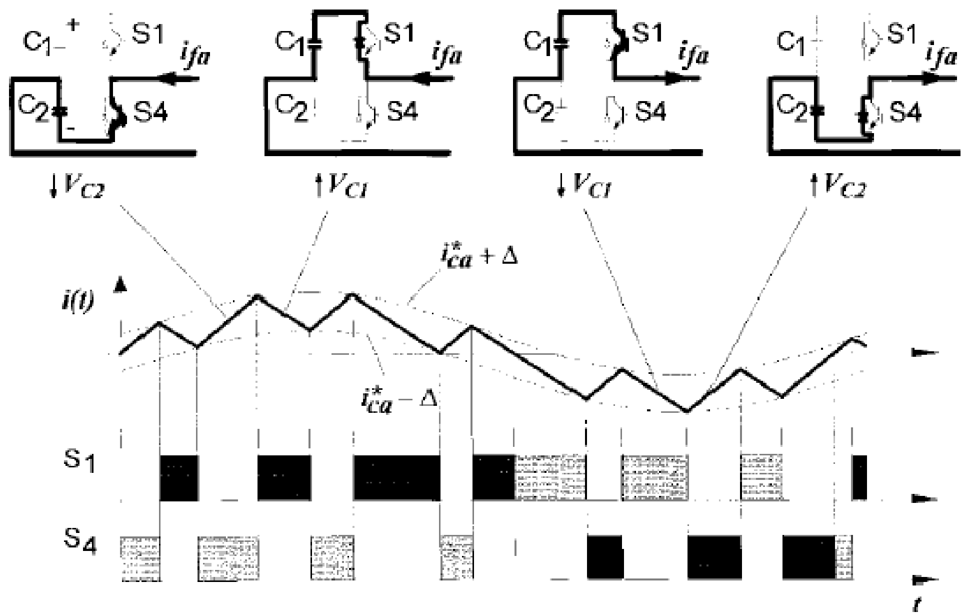


Figure 2.7-3: Hysteresis Band Current Control

2.8 SHUNT ACTIVE FILTER CONTROL APPROACH:

There are numerous approaches by which realization of shunt active filter control is possible. Instantaneous active reactive power theory was used to implement the shunt active filter. Instantaneous active reactive power theory based on which the active filter was designed is discussed below

2.8.1. INSTANTANEOUS ACTIVE REACTIVE POWER THEORY

The instantaneous power in three-phase is defined on the basis of instantaneous concept for arbitrary voltage and current waveforms including transient states. The detailed description of this theory is as follows.

To deal with instantaneous voltages and current in three phase circuits mathematically, it is adequate to express these quantities as the instantaneous space vectors. For simplicity, the three phase voltages and currents excluding zero sequence components will be considered in the following. The derivation with zero sequence is given in Appendix-II-B.

Figure 2.8.1-1 shows a three-phase three-wire system on the a-b-c coordinates, where no zero-sequence voltage is included in the three-phase three-wire system.

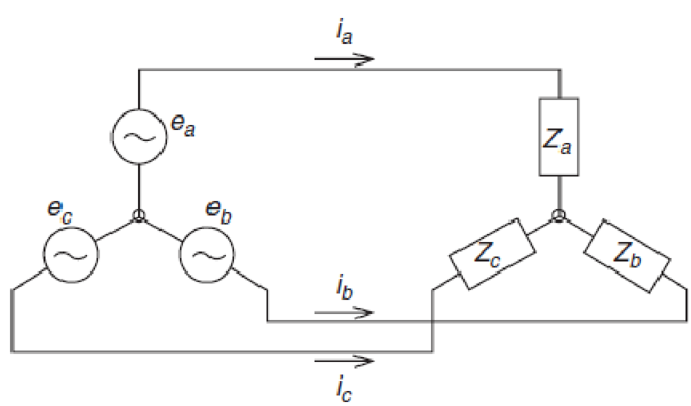


Figure 2.8.1-1: Three-phase three-wire system

In a-b-c coordinates, the a, b and c axes are fixed on the same plane, apart from each other by $2\pi/3$, as shown in Figure 2.8.1-2 The instantaneous space vectors, e_a and i_a are set on the a axis, and their amplitude and (+,-) direction vary with passage of time. In the same way, e_b and i_b are on the b axis, and e_c and i_c are on the c axis.

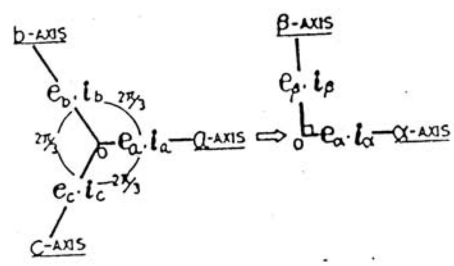


Figure 2.8.1-2: α-βCoordinates Transformer

These space vectors are easily transformed into α-β coordinates as follows:

$$\begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & \frac{-1}{2} & \frac{-1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} \tag{2.8.1-1}$$

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & \frac{-1}{2} & \frac{-1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{2.8.1-2}$$

Where the α and β axes are the orthogonal coordinates. Necessarily, e_α and i_α are on the α axis, e_β and i_β are on the β axis. Their amplitude and (+,-) direction vary with passage of the time.

Figure 2.8.1-2 shows the instantaneous space vectors on the α-β coordinated. As is well known, the instantaneous real power either on the a-b-c coordinates or on the α-βcoordinates is defined by

$$p = e_\alpha i_\alpha + e_\beta i_\beta = e_a i_a + e_b i_b + e_c i_c \qquad (2.8.1\text{-}3)$$

Where p is equal to the conventional power.

In order to define the instantaneous reactive power, the space vector is defined by

$$q = e_\alpha\, i_\beta - e_\beta\, i_\alpha \qquad (2.8.1\text{-}4)$$

As shown in Figure 2.8.1-3 this space vector is the imaginary axis vector and is perpendicular to the real plane on the α-β coordinates, to be in compliance with right hand rule.

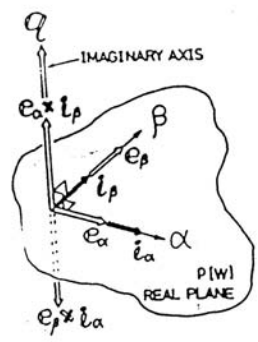


Figure 2.8.1-3: Instantaneous Space Vector

Taking into consideration that e_α is parallel to i_α, and that e_α is perpendicular to i_β and e_β to i_α, the conventional instantaneous imaginary power q, which of amplitude q, are expressed by

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \qquad (2.8.1\text{-}5)$$

In Equation (2.8.1-5) $e_\alpha i_\alpha$ and $i_\beta e_\beta$ obviously mean instantaneous power because they are defined by the product of the instantaneous voltage in one axis and the instantaneous current in the same axis. Therefore, "p" is the real power in the three-phase circuit, and its dimension is (W). Conversely, $e_\alpha i_\beta$

and $e_\beta i_\alpha$ are not instantaneous power because they are defined by the product of the instantaneous voltage in the in one axis and the instantaneous current not in the same axis but in the perpendicular axis.

Accordingly q cannot be dealt with conventional quantity. “q” is quite different from “p” in dimension and electric property although “q” looks similar in formulation to “p”. A common dimension for “q” should be introduced from both theoretical and practical points of view. A good candidate is [IW], that is, ‘‘imaginary watt.’’

Definition and physical meaning of instantaneous reactive power in Equation (2.8.1-5) is changed into the following equation:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} p \\ q \end{bmatrix} \tag{2.8.1-6}$$

Note that the determinant with respect to e_α and e_β in Equation (2.8.1-5) is not zero. The instantaneous currents on the α-β coordinates, i_α and i_β are divided into two kinds of instantaneous current components, respectively:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} p \\ 0 \end{bmatrix} + \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ q \end{bmatrix} \tag{2.8.1-7}$$

$$= \begin{bmatrix} i_{\alpha p} \\ i_{\beta p} \end{bmatrix} + \begin{bmatrix} i_{\alpha q} \\ i_{\beta q} \end{bmatrix} \tag{2.8.1-8}$$

α - axis instantaneous active current

$$i_{\alpha p} = \frac{e_\alpha}{{e_\alpha}^2 + {e_\beta}^2} p \tag{2.8.1-9}$$

α axis instantaneous reactive current

$$i_{\alpha q} = \frac{-e_\beta}{e_\alpha{}^2 + e_\beta{}^2} q \qquad (2.8.1\text{-}10)$$

β axis instantaneous active current

$$i_{\beta p} = \frac{e_\beta}{e_\alpha{}^2 + e_\beta{}^2} p \qquad (2.8.1\text{-}11)$$

β axis instantaneous reactive current

$$i_{\beta q} = \frac{e_\alpha}{e_\alpha{}^2 + e_\beta{}^2} q \qquad (2.8.1\text{-}12)$$

Both physical meaning and reason for the naming of the active and the reactive currents are clarified in the following.

Let the instantaneous powers in the α axis and the β axis is p_α and $p_\beta$respectively, they are given by the conventional definition as follows.

$$\begin{bmatrix} p_\alpha \\ p_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha i_\alpha \\ e_\beta i_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha i_{\alpha p} \\ e_\beta i_{\beta p} \end{bmatrix} + \begin{bmatrix} e_\alpha i_{\alpha q} \\ e_\beta i_{\beta q} \end{bmatrix} \qquad (7.1.13)$$

The instantaneous real power in the three phase circuit p is given as follows using Equation (2.8.1-7) and (2.8.1-13),

$$p = p_\alpha + p_\beta \quad = e_\alpha i_{\alpha p} + e_\beta i_{\beta p} + e_\alpha i_{\alpha q} + e_\beta i_{\beta q}$$

$$= \frac{e^2{}_\alpha}{e_\alpha{}^2 + e_\beta{}^2} p + \frac{e^2{}_\beta}{e_\alpha{}^2 + e_\beta{}^2} p + \frac{-e_\alpha e_\beta}{e_\alpha{}^2 + e_\beta{}^2} q + \frac{e_\alpha * e_\beta}{e_\alpha{}^2 + e_\beta{}^2} q \qquad (2.8.1\text{-}14)$$

Note that the sum of the third and fourth term of the right hand side is always zero from Equation (7.1.13) and (7.1.14) following equations are obtained

$$p = e_\alpha i_{\beta p} + e_\beta i_{\beta p} = p_{\alpha p} + p_{\beta p} \qquad (2.8.1\text{-}15)$$

$$0 = e_\alpha i_{\alpha q} + e_\beta i_{\beta q} = p_{\alpha q} + p_{\beta q} \qquad (2.8.1\text{-}16)$$

where,

α axis reactive power = $p_{\alpha p} = \dfrac{e^2{}_\alpha}{e_\alpha{}^2 + e_\beta{}^2} p$

α axis instantaneous active current = $p_{\alpha q} = \dfrac{-e_\alpha e_\beta}{e_\alpha{}^2 + e_\beta{}^2} q$

β axis instantaneous active current = $p_{\beta p} = \dfrac{e^2{}_\beta}{e_\alpha{}^2 + e_\beta{}^2} p$

β axis instantaneous reactive power = $p_{\beta q} = \dfrac{e_\alpha e_\beta}{e_\alpha{}^2 + e_\beta{}^2} q$

Equation (2.8.1-15) & (2.8.1-16) leads to the following conclusion

1. The sum of the power components, $p_{\alpha p}$ and $p_{\beta p}$, coincides with the three-phase instantaneous real power, "p", which is given by Equation (2.8.1-3).Therefore, $p_{\alpha p}$ and $p_{\beta p}$ are referred to as the α-phase and β-phase instantaneous active powers.
2. The other power components, $p_{\alpha q}$ and $p_{\beta q}$, cancel each other and make no contribution to the instantaneous power flow from the source to the load. Therefore, $p_{\alpha q}$ and $p_{\beta q}$ are referred to as the α-phase and β-phase instantaneous reactive powers.
3. Thus, a shunt active filter without energy storage can achieve instantaneous compensation of the current components, $i_{\alpha q}$ and $i_{\beta q}$ or the power components, $p_{\alpha q}$ and $p_{\beta q}$. In other words, the Akagi-Nabae theory based on Equation (2.8.1-5) exactly reveals what components the active filter without energy storage can eliminate from the α-phase and β-phase instantaneous currents, i_α and i_β or the α-phase and β-phase instantaneous real powers, $p_\alpha$and p_β as shown in Figure 2.8.1.4.

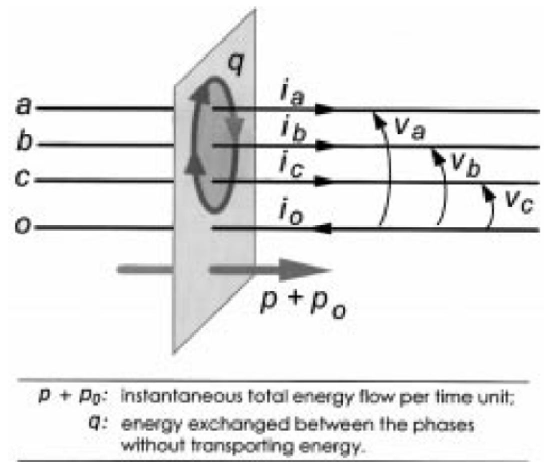


Figure 2.8.1-4 (a): physical meaning of the instantaneous power defined in the α-β-o reference frame

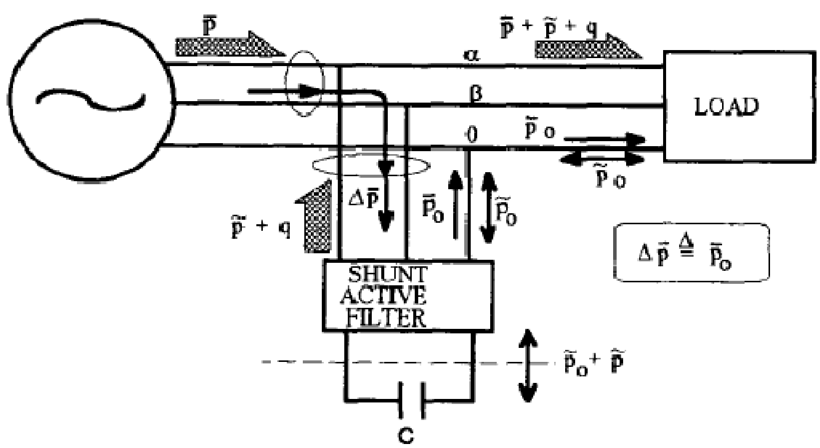


Figure 2.8.1-4 (b): Compensation of all undesirable power of the load by using a shunt active filter

2.8.1.1 CONTROL STRATEGY:

Figure 2.8.1.1-1 shows the basic compensation scheme of the instantaneous reactive power, where p_s and q_s are the instantaneous real and imaginary powers on the source side p_c and q_c are those on the active filter side.

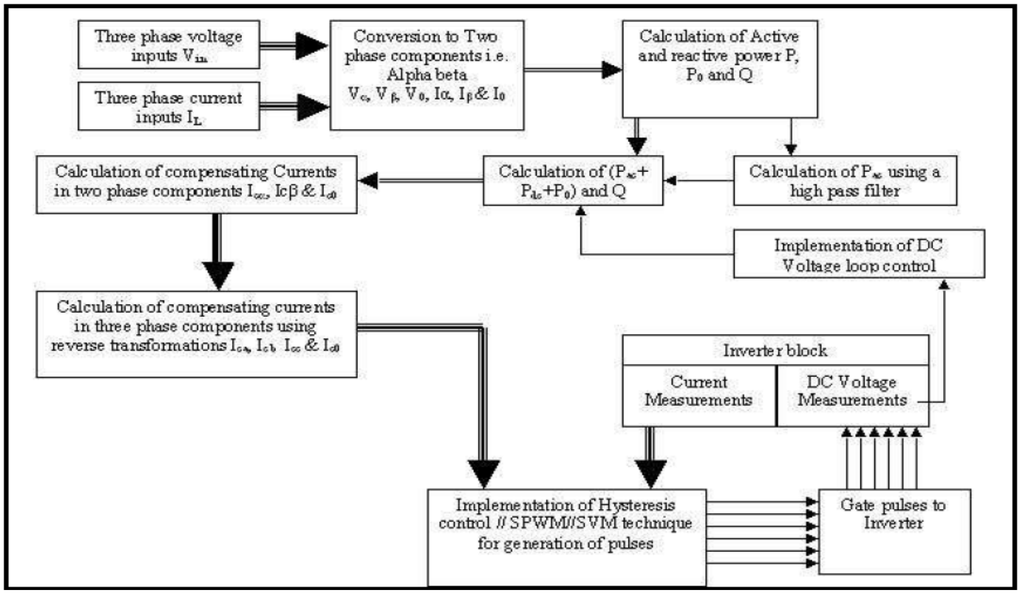


Figure 2.8.1.1-1: Shunt Active Power Filter control scheme using IARP theory

The active filter eliminates the instantaneous reactive power on the source side and harmonics, which are caused by the instantaneous power on the load side. The active power filter consists of only switching devices without energy storage components, because pc is always zero. The instantaneous compensating currents on the α-β coordinates ic_α and $i_{c\beta}$ are given by

$$\begin{bmatrix} i_{c\alpha} \\ i_{c\beta} \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} -p_{ac} \\ -q \end{bmatrix} \qquad (2.8.1)$$

The basic principle of the active filter will not be considered, concerning the α axis instantaneous current on the load side. The instantaneous active and the reactive currents are divided into the following two kinds of instantaneous currents respectively:

$$i_\alpha = \frac{e_\alpha}{{e_\alpha}^2 + {e_\beta}^2} p_{dc} + \frac{e_\alpha}{{e_\alpha}^2 + {e_\beta}^2} p_{ac} + \frac{-e_\beta}{{e_\alpha}^2 + {e_\beta}^2} q_{dc} + \frac{-e_\beta}{{e_\alpha}^2 + {e_\beta}^2} q_{ac}$$

(2.8.2)

where p_{dc} and p_{ac} are the DC and AC components of the instantaneous real power and q_{dc} and q_{ac} are the DC and AC components of the instantaneous imaginary power. The first term of the right hand side of Equation is the instantaneous value of the conventional fundamental active current. The third term is the instantaneous value of the conventional fundamental reactive current. The second term is the instantaneous value of the harmonic current which represents the AC component of the instantaneous real power. The fourth term is the instantaneous value of the harmonic currents, which represents the AC components of the instantaneous imaginary power. Accordingly, the sum of the second and fourth terms is the instantaneous value of the conventional harmonic currents.

Equation (7.2.2) leads to the following conclusions:

1). The active filter eliminates both the third and fourth terms. For this reason, the displacement factor is unity not only in steady states but also in transient states.

2). The harmonic currents represented by the second and fourth term can be eliminated by the active filter comprising switching devices without energy storage components.

2.8.1.2 SIMULATION RESULTS

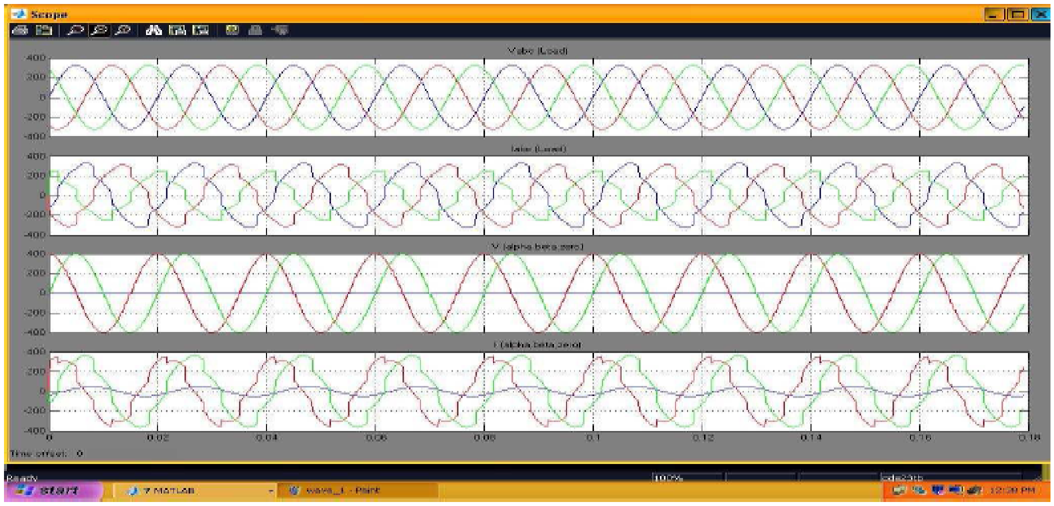


Figure 2.8.1.2-2: Source voltages and Load currents alpha beta components for voltage and current

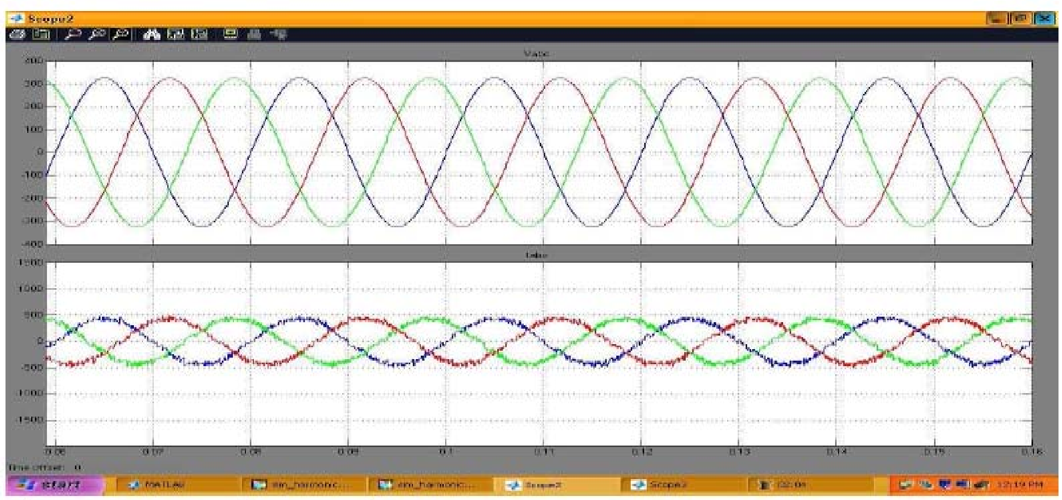


Figure 2.8.1.2-3: source voltages and currents

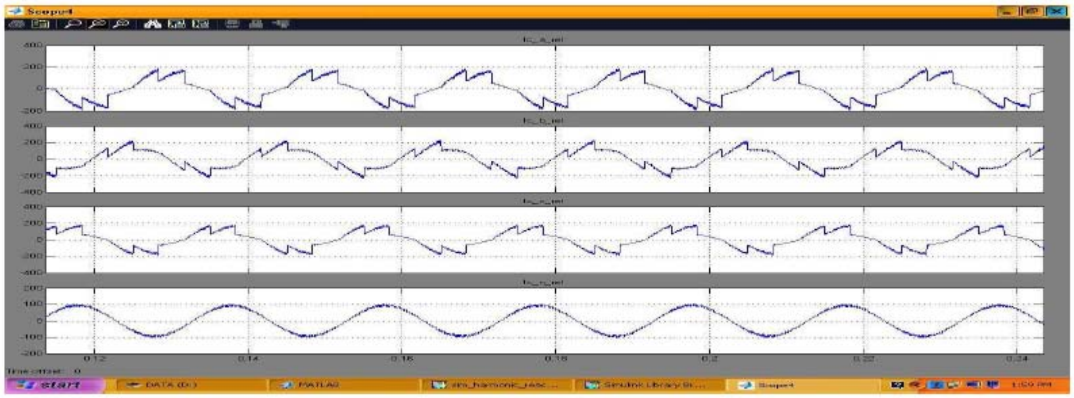


Figure 2.8.1.2-4: Filter currents

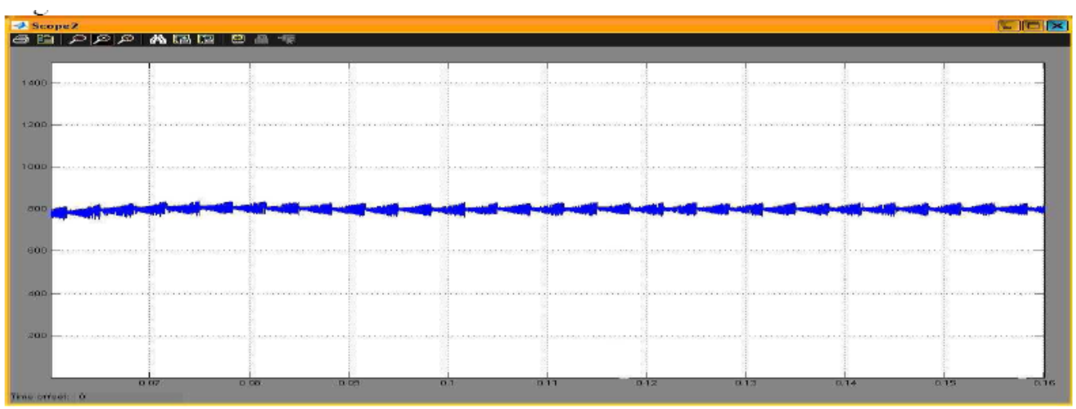


Figure 2.8.1.2-5: DC Bus Voltage

Figure 2.7.1 shows the schematic of three-phase four-wire active filter. A three-phase diode rectifier system with resistive load as harmonic producing loads for experimental purpose is used.

The corresponding waveforms of three phase compensated load currents and waveform at different places are shown at the end of this chapter

Detail simulation as well experimental results for shunt active power filter using instantaneous active reactive power theory are given in Annexure-II-C

2.8.2. SINE MULTIPLICATION THEORY

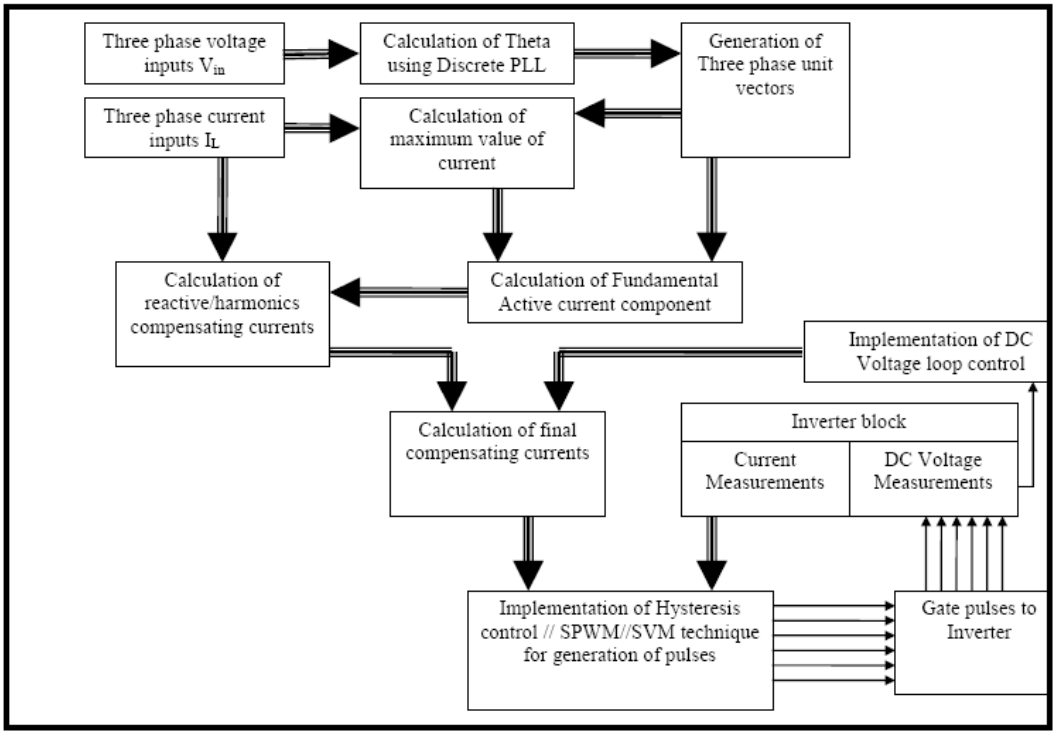


Figure 2.8.2-1: Shunt Active Power Filter control scheme using Sine Multiplication theory

Step I: Calculate sin(*t), cos(*t), *t using three phase lock loop

Step II: Calculate the Maximum value of the Fundamental active current using the sine multiplication method. Initially it is required to multiply phase load currents with corresponding sinusoidal unit vectors as follows:

$$\boxed{\begin{array}{l} I_{La} * \sin \omega t \\ I_{Lb} * \sin\left(\omega t - 2\pi/3\right) \\ I_{Lc} * \sin\left(\omega t + 2\pi/3\right) \end{array}}$$

Integrating the above set of equations over a cycle will generate maximum value of the current I_m

Multiplying the value of I_m with three phase unit vectors will generate three phase fundamental active currents supplied by the source as follows:

$$\boxed{\begin{aligned} I_{1a} &= I_m * \sin \omega t \\ I_{1b} &= I_m * \sin\left(\omega t - 2\pi/3\right) \\ I_{1c} &= I_m * \sin\left(\omega t + 2\pi/3\right) \end{aligned}}$$

Harmonics and reactive current components of the source currents can be calculated by subtracting the fundamental active current from the load currents. The resultant currents are the compensating reference currents for the active filter.

$$\boxed{\begin{aligned} \vec{I}_{ca} &= \vec{I}_{La} - \vec{I}_{af} \\ \vec{I}_{cb} &= \vec{I}_{Lb} - \vec{I}_{bf} \\ \vec{I}_{cc} &= \vec{I}_{Lc} - \vec{I}_{cf} \end{aligned}}$$

The unbalance in load currents in is compensated introducing forth leg in the three-leg inverter. The compensating reference current for this leg is calculated using three phase load currents.

$$\boxed{\vec{I}_{cn} = -\left(\vec{I}_{La} + \vec{I}_{Lb} + \vec{I}_{Lc}\right)}$$

The capacitor DC bus voltage is controlled using a proportional control. The output of the control is multiplied with the three phase voltage unit vectors and thus the resultant would be three current signals (one for each phase) required for maintaining DC bus voltage at desired levels.

Hysteresis current control technique is used for switching of the IGBTs of the inverter so as to make the inverter currents follow the reference compensating signals.

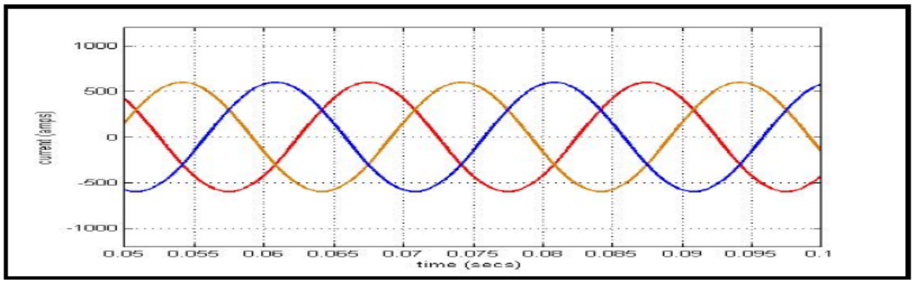


Figure 2.8.2-2: Input Source Voltages

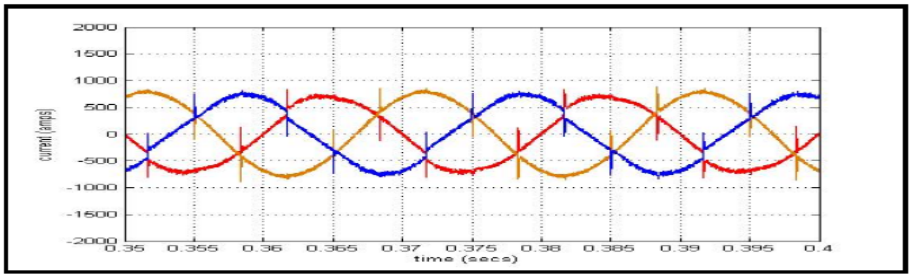


Figure 2.8.2-3: Input Source Currents

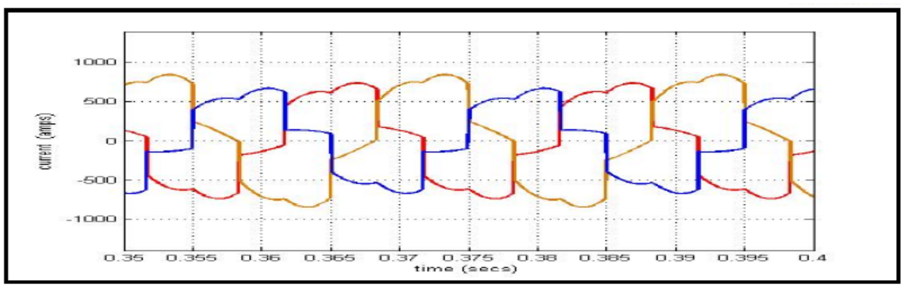


Figure 2.8.2-4: Load Currents

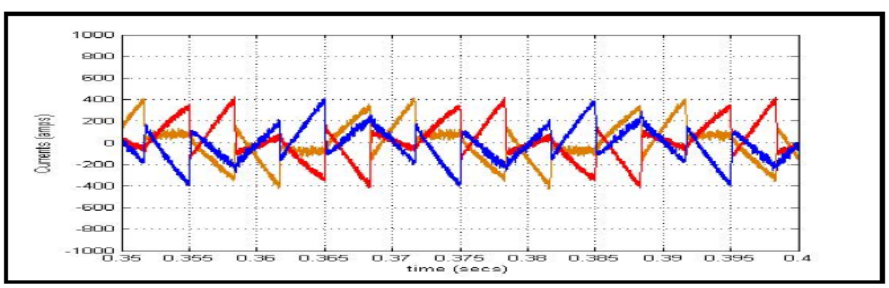


Figure 2.8.2-5: Filter Currents

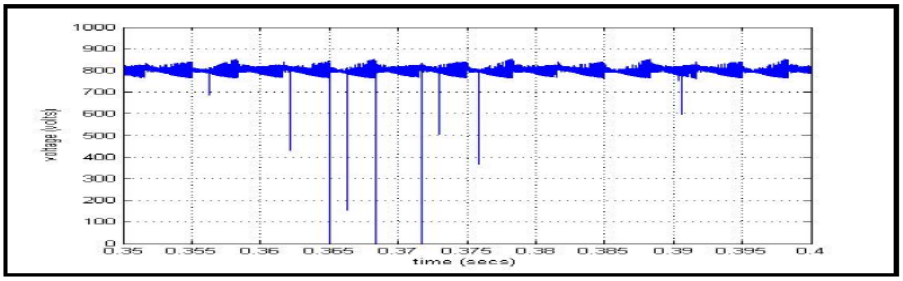


Figure 2.8.2-6: DC Bus voltage

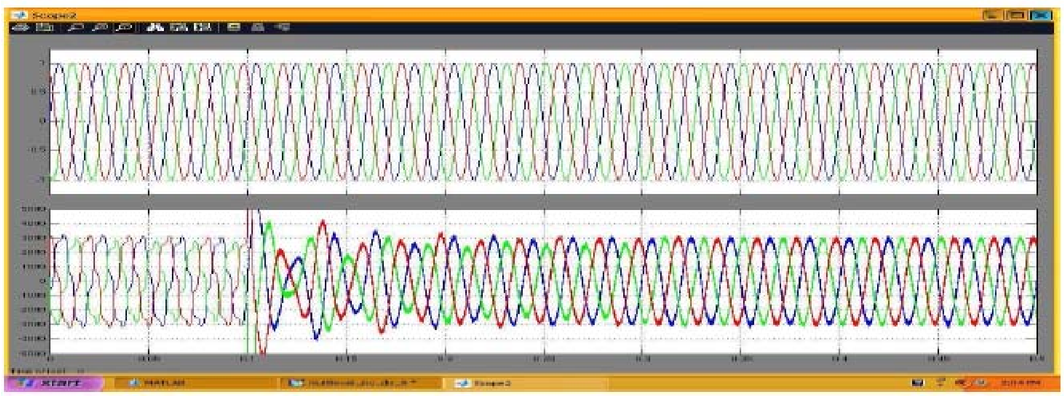


Figure 2.8.2-7 Input Source Voltages and Currents

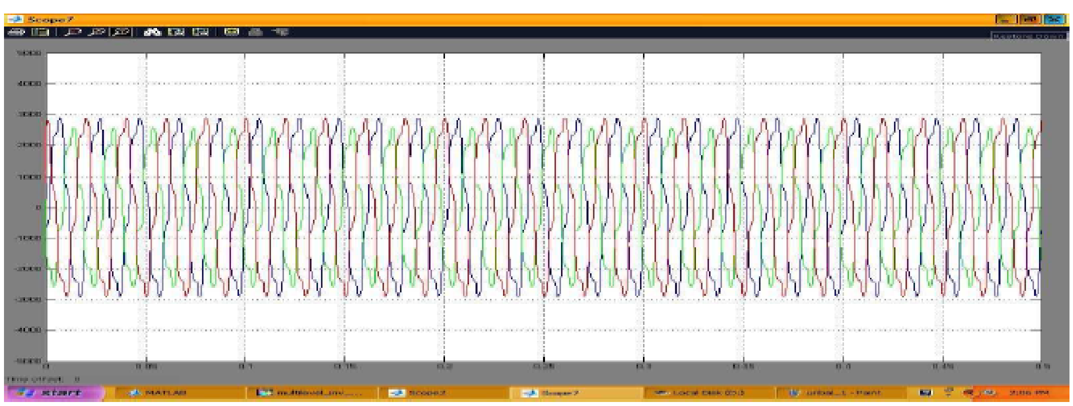


Figure 2.8.2-8: Load Currents

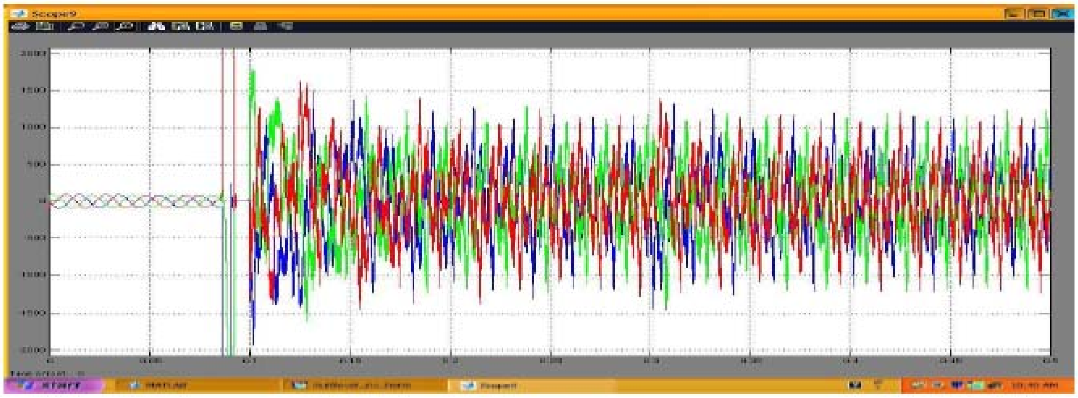


Figure 2.8.2-9: Filter Currents

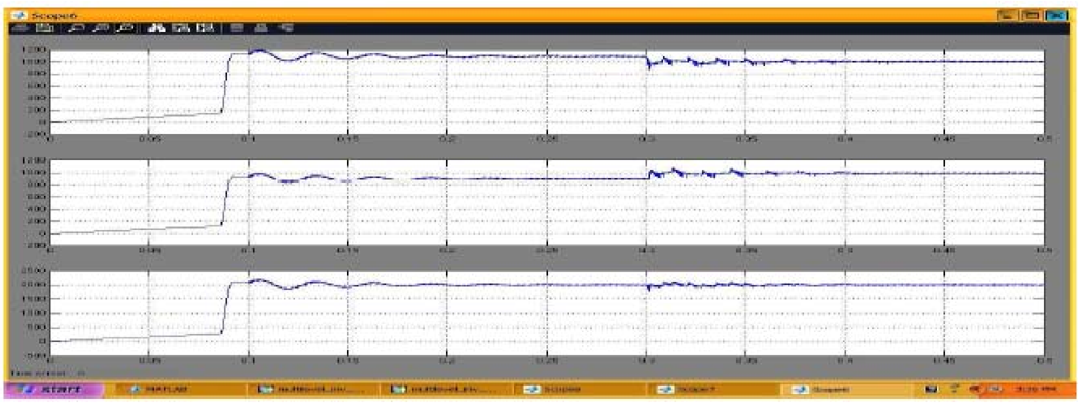


Figure 2.8.2-10: DC Bus Voltages

2.8.3. SYNCHRONOUS REFERENCE FRAME THEORY

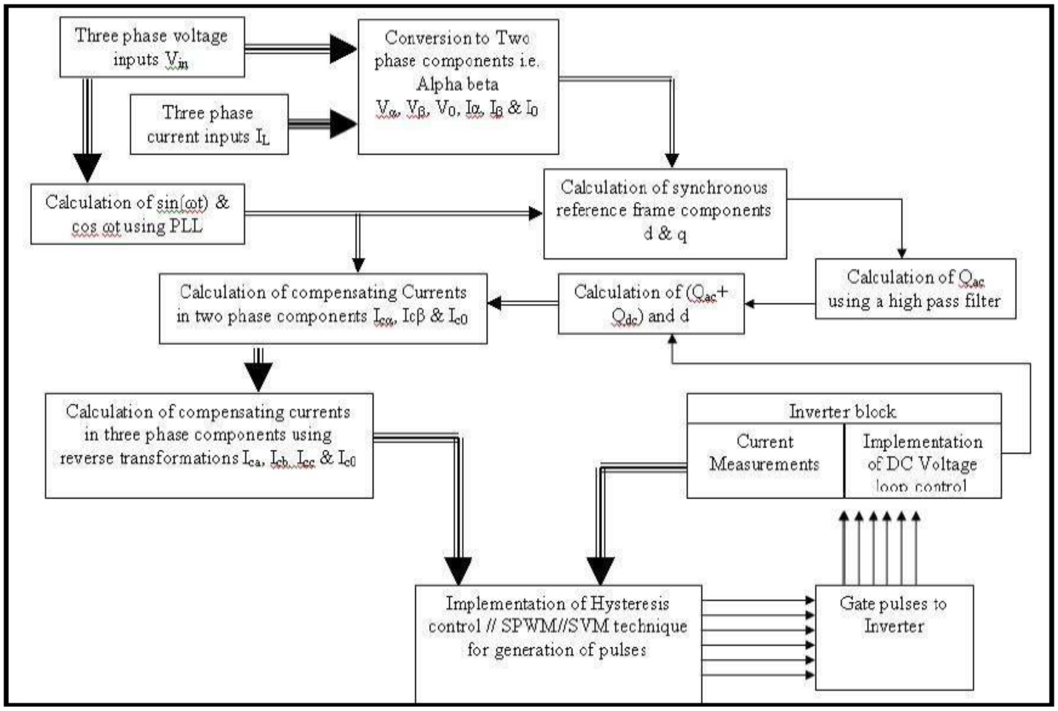


Figure 2.8.3-1: Shunt Active Power Filter control scheme using Synchronous Reference Frame Theory

Step 1. Conversion of three-phase coordinates to two-phase alpha beta coordinates

$$\begin{bmatrix} v_0 \\ v_\alpha \\ v_\beta \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}$$

$$\begin{bmatrix} i_0 \\ i_\alpha \\ i_\beta \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}$$

Step 2. Conversion of stationary reference frame components to rotating reference frame components

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix}$$

Step 3. Extraction of components required for compensation

$$\begin{aligned} i_d &= \overline{i_d} + \tilde{i}_d \\ i_q &= \overline{i_q} + \tilde{i}_q \end{aligned}$$

Out of these, id & i_{qac} are components of interest. Along with these components, the reference required to maintain the DC voltage across the capacitor is calculated using a PI control applied to the DC loop. The output of the PI loop is added to the iqac.

Step 4. Calculation of compensating reference signals in two-phase stationary frame from

rotating reference frame

$$\begin{bmatrix} i_{c\alpha} \\ i_{c\beta} \end{bmatrix} = \begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix} \begin{bmatrix} i_d \\ \tilde{i}_q + e_{dc} \end{bmatrix}$$

$$\boxed{i_{c0} = i_0}$$

Step 5. Calculation of compensating reference signals in three-phase a, b, c components

$$\begin{bmatrix} i_{ca} \\ i_{cb} \\ i_{cc} \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/\sqrt{2} & 1 & 0 \\ 1/\sqrt{2} & -1/2 & \sqrt{3}/2 \\ 1/\sqrt{2}\,0 & -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_{c0} \\ i_{c\alpha} \\ i_{c\beta} \end{bmatrix}$$

The reference-compensating signal for neutral currents is calculated as

$$\boxed{i_{cn} = i_{ca}^{*} + i_{cb}^{*} + i_{cc}^{*}}$$

A hysteresis controller with +/-5% band is used for generating the switching pulses for the inverter.

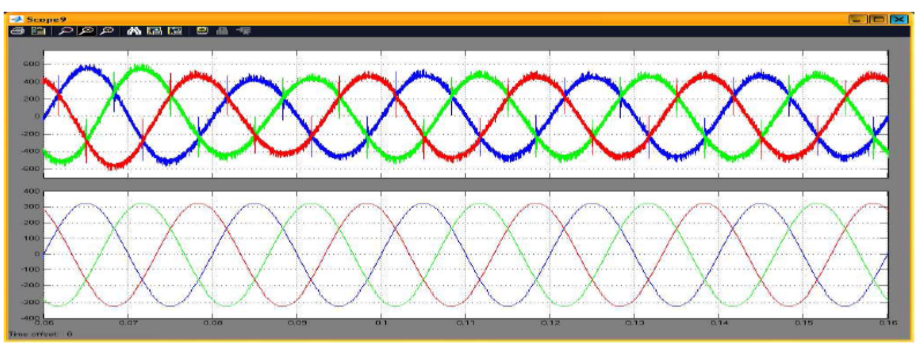


Figure 2.8.3-2: Source Currents and Voltages

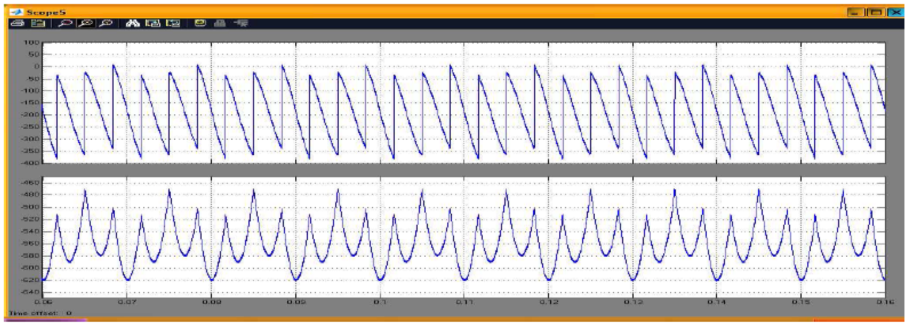


Figure 2.8.3-3: I_d-I_q components

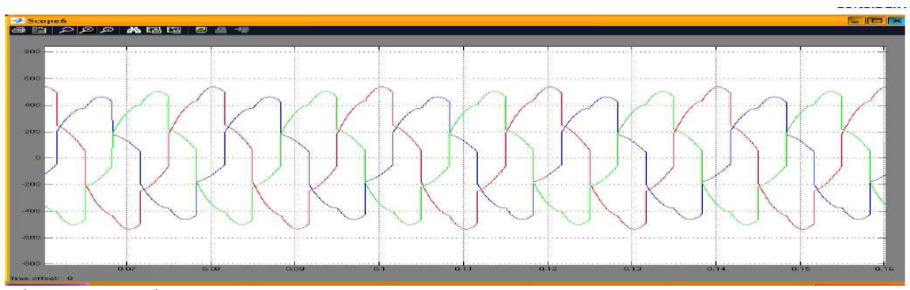


Figure 2.8.3-4: Load Currents

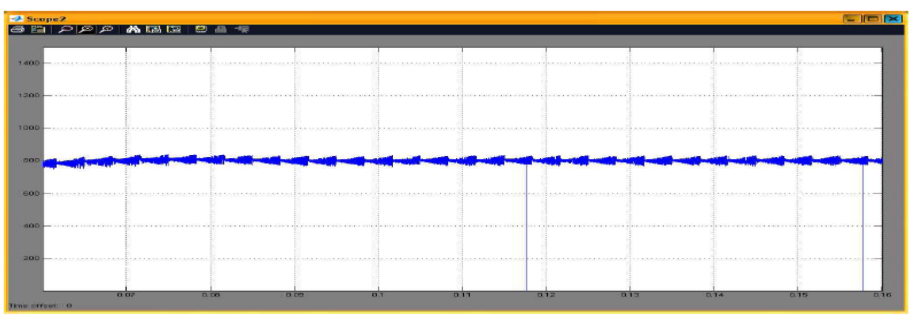


Figure 2.8.3-5: DC Bus Voltage

2.9 ACTIVE POWER FACTOR CORRECTION INTEGRATED TO THE INPUT STAGE OF THE EQUIPMENT

Power quality problems posed by single-phase rectifier load are well known. Such loads distort the supply voltage waveform, causing overheating of transformers & neutral conductors and ultimate equipment malfunction. There have been several solutions to the problems proposed in the recent years and various equipment and system standards [5]-[6] such as the IEEE 519:1999 have been published. Also with imposed of harmonic standards such as IEC 61000-3-2 by international communities, it would be much important aspect to consider in every design of the appliances.

Solutions such as (a) passive filter at input side (b) active filter for eliminating current harmonics, (c) active filter for eliminating voltage harmonics, negative sequence voltage, zero sequence voltage (d) DVR have been proposed to improve power quality [2]-[4].

Solution to improve power quality for single phase appliance include (a) passive filter at input side (b) Ferro resonant transformer, (c) active power factor correction integrated to the input stage of the rectifier, and (d) isolation transformer with third harmonic trap. These techniques improve the input current waveform and the power factor and hence power quality.

Due to advancement in power electronics technology, energy saving electronic ballasts is becoming more popular. Lighting load is more in domestic application and therefore power quality problems are much concern in distribution system. This paper presents the use of boost converter in HPSV electronic ballast to improve its performance so that input supply current becomes sinusoidal & power factor improves to unity and hence the power quality of the system will improve.

The operation of the proposed single-phase boost converter is described in 2.9.1. In section 2.9.2 analysis & simulation along with the design of current & voltage control loops are presented. In section IV experimental setup & comparison of current waveform with & without boost converter are explained which is followed by conclusion.

2.9.1. PRINCIPLE OF OPERATION

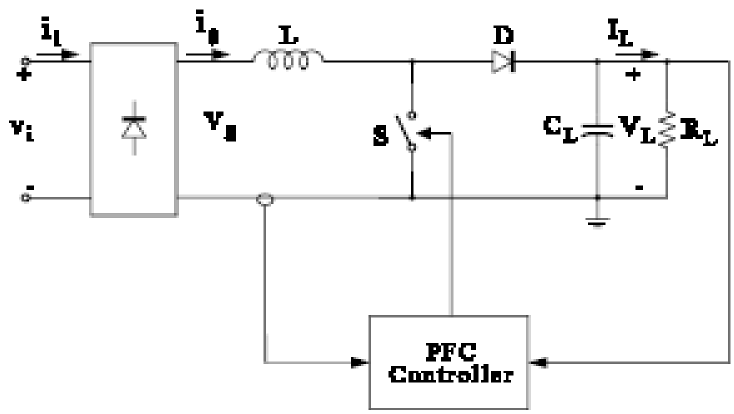


Figure 2.9.1-1: Principal scheme of a boost PFC converter

Figure 2.9.1-1shows the principle scheme of a boost PFC converter. The boost topology is very simple and allows low distorted input currents an almost unity power factor with different control techniques. Moreover, the switch is connected to ground, which simplifies the drive circuit. The main draw backs of this topology are: (1) startup over currents, due to the charge of the large output capacitor, (2) lack of current limitation during overload and short circuit condition due to direct connection between line and load, (3) lack of isolation and (4) higher output voltage always greater than peak input voltage.

In spite of these limitations many PFC’s based on the boost topology have been proposed in the literature. Various control strategies have also been implemented. In the following, borderline control scheme, which is implemented to improve harmonics & power factor, is discussed.

Borderline Control:

In this approach the switch on time is held constant during the line cycle & the switch is turned ON when the inductor current falls to zero, so that the converter operates at the boundary between continuous and discontinues inductor current mode (CCIM-DCIM) [7].

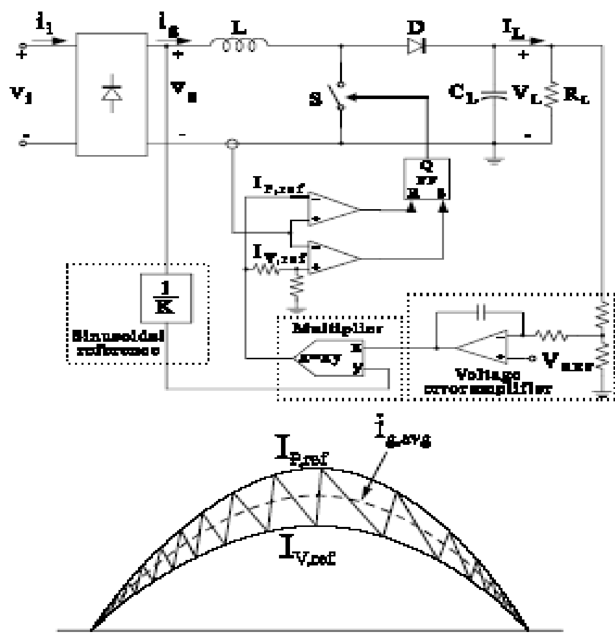


Figure 2.9.1-2: Hysteresis Control Scheme.

In this way, the freewheeling diode is turned off softly (no recovery losses) and the switch is turned ON at zero current, so the commutation losses are reduced. On the other hand, the higher current peaks increase device stresses and conduction losses and may call for heavier input filters (for some topologies).

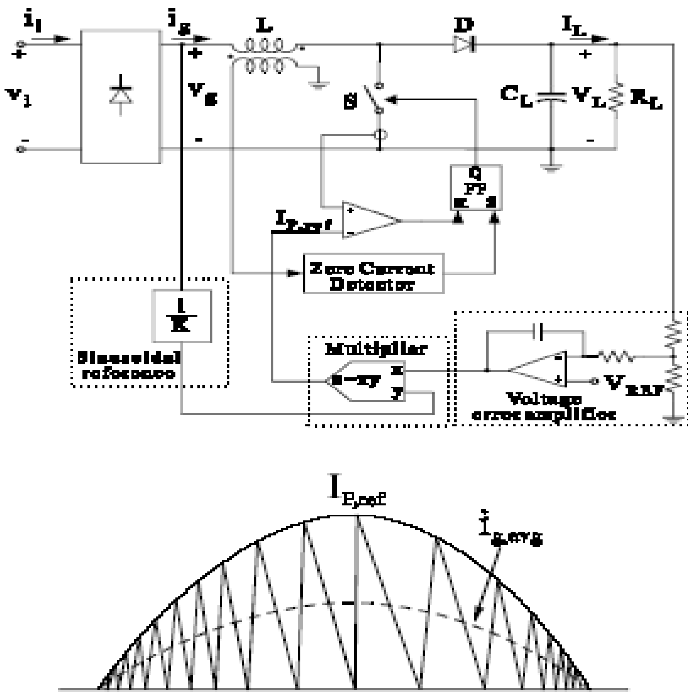


Figure 2.9.1-3: Borderline Control Scheme.

This type of control is a particular case of hysteresis control, in which the lower reference Iv, ref is zero anywhere, which is shown in Figure 2.9.1-2. The principle scheme of borderline control is shown in Figure 2.9.1-3.

The instantaneous input current is constituted by sequence of triangles whose peaks are proportional to the line voltage without duty-cycle modulation during the line cycle. This characterizes this control as an "automatic current shaper" technique.

2.9.2. DESIGN ANALYSIS & SIMULATION

The output of the boost converter [1] is given by,

$$V_o = V_{in}\left(\frac{1}{1-D}\right) \;\&\; I_o = I_d(1-D) \tag{2.9.2-1}$$

Average value of the inductor current I_{LB} at the boundary between continuous & discontinuous conduction is

$$I_{LB} = \frac{T_s V_o}{2L} D(1-D) \tag{2.9.2-2}$$

Average output current I_{oB} at the edge of continuous current is

$$I_{oB} = \frac{T_s V_o}{2L} D(1-D)^2 \tag{2.9.2-3}$$

$$I_{LB,\max} = \frac{T_s V_o}{8L} \quad when \quad D = 0.5$$

$$I_{oB,\max} = \frac{2}{27}\frac{T_s V_o}{L} \qquad \text{when} \quad D = 0.333 \tag{2.9.2-4}$$

To design the boost converter

The maximum power required is calculated using

$$p_o = V_o I_o \tag{2.9.2-5}$$

Similarly, the Peak inductor current at minimum required ac line voltage for output regulation.

$$I_{l(pk)} = \frac{2\sqrt{2}P_o}{\eta V_{ac(LL)}} \tag{2.9.2-6}$$

If the switching frequency for input voltage of 85 to 265 Vac is 25 kHz, the inductance L can be calculated as

$$L_p = \frac{t\left(\frac{V_o}{\sqrt{2}} - V_{ac(LL)}\right)\eta\left(V_{ac(LL)}\right)^2}{\sqrt{2}V_o Po} \tag{2.9.2-7}$$

Switching ON time t_{on} for this scheme is nearly constant, therefore switch OFF time t_{off} which varies through the cycle can calculated as

$$T_{off} = \frac{T_{on}}{\left(\frac{V_o}{\sqrt{2}V_{ac}\left|Sin\theta\right|} - 1\right)} \qquad (2.9.2\text{-}8)$$

Where θ represents the angle of the ac line voltage, Ton represents switch on time, T_{off} represents switch off time and L represents boost inductor.

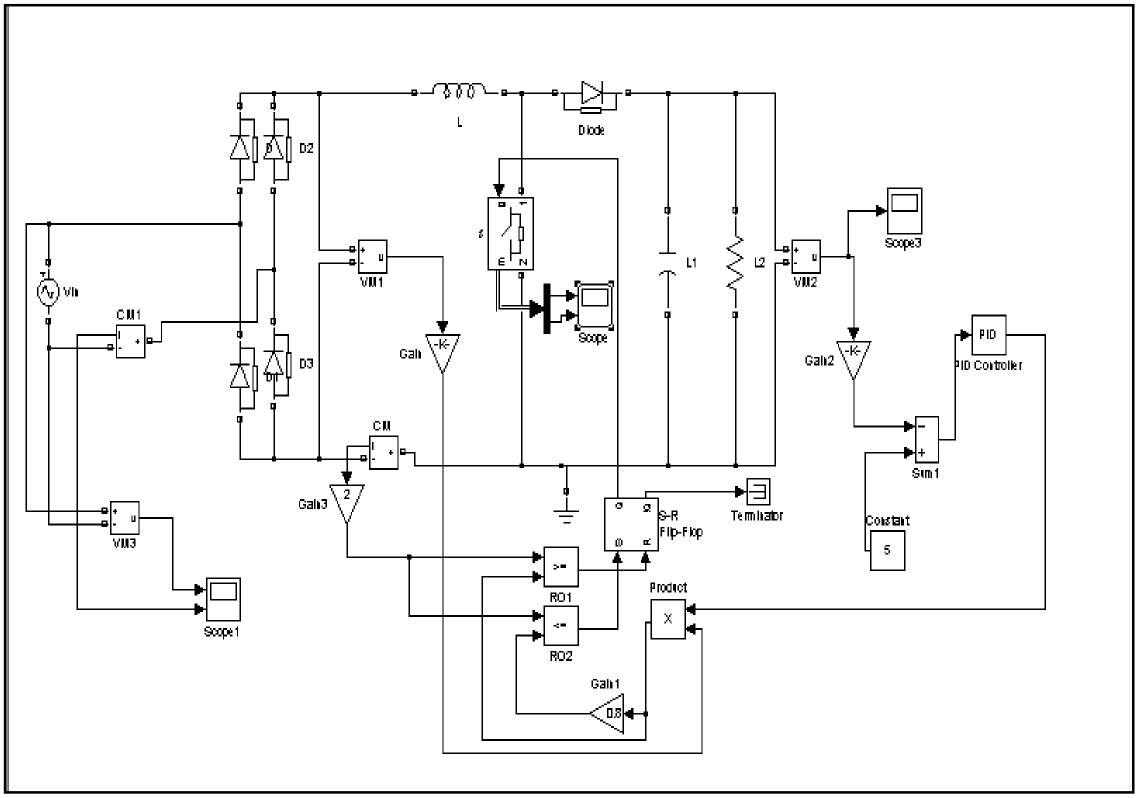


Figure 2.9.2-4: Simulated system using MATLAB for UPF boost converter

The mains AC supply voltage is rectified and supplied to the boost converter that mainly consists of inductor, a power MOSFET, a high frequency power diode and a Bulk Capacitor.

The simulation of the single-phase boost converter with low THD & unity power factor was done using MATLAB-SIMULINK software for its performance. The Block diagram of UPF boost converter simulated is shown in Figure 2.9.2-4.

The input voltage, input current, DC output voltage & voltage across inductor respectively are shown in Figure 2.9.2-5. The waveforms show that the input current & voltage waveform are in phase with minimum distortion.

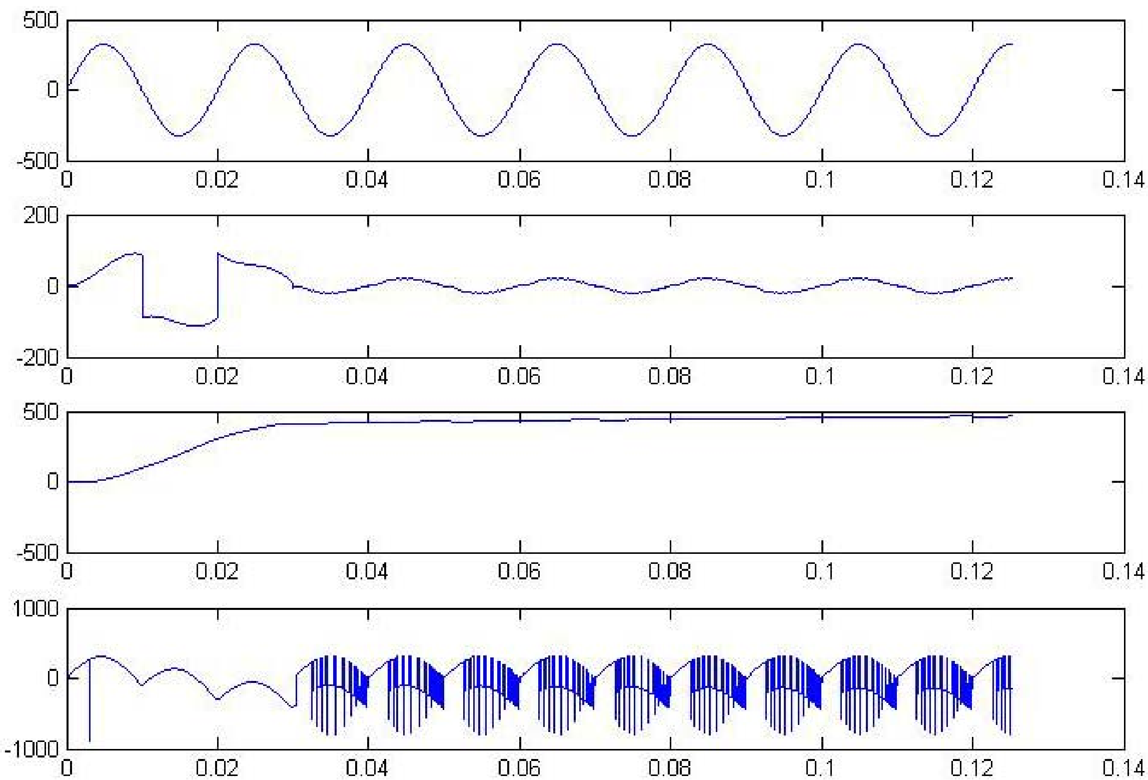


Figure 2.9.2-5: Simulated waveform of input voltage, input current, output DC voltage & voltage across inductor respectively.

2.9.3. EXPERIMENTATION & RESULTS

Input current show heavy inrush while starting, due to initially bulk capacitor is charged. Inductor was fabricated using two EE ferrite core of standard size EE 42 x 21 x 20 having the value of inductance 1.86 mH.

The number of Turns is 66 with air gap of 1.8 mm. To detect the zero current, separate winding was inserted in the inductor. The polarity of the second winding is such that when the main winding voltage is positive the second winding voltage is of reverse polarity. The numbers of Turns in the second winding is 6.

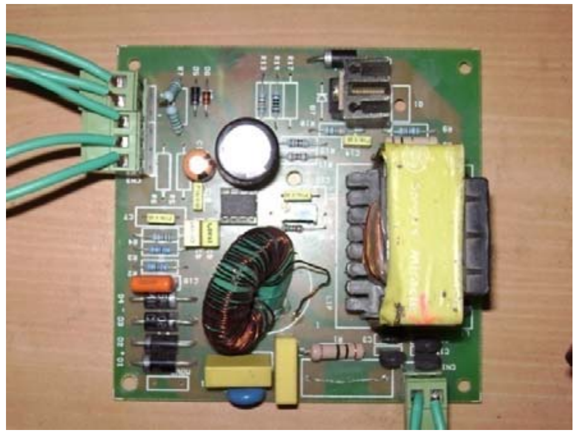


Figure 2.9.3-6: Rectifier with PFC assembled on PCB

IRF 840 MOSFET is used as switch to operate at 25 kHz. Bulk capacitor of the value 220 μF, 450 volts is used to filter out the output ripple. Ultra High frequency diode MUR 460 with very low t_{rr} is used in the boost converter.

To limit the initial inrush current to charge the bulk capacitor NTC-15 are used as shown in the Figure 2.9.3-7. A small π filter is used to filter out high frequency ripple in input current which is due to switch of MOSFET.

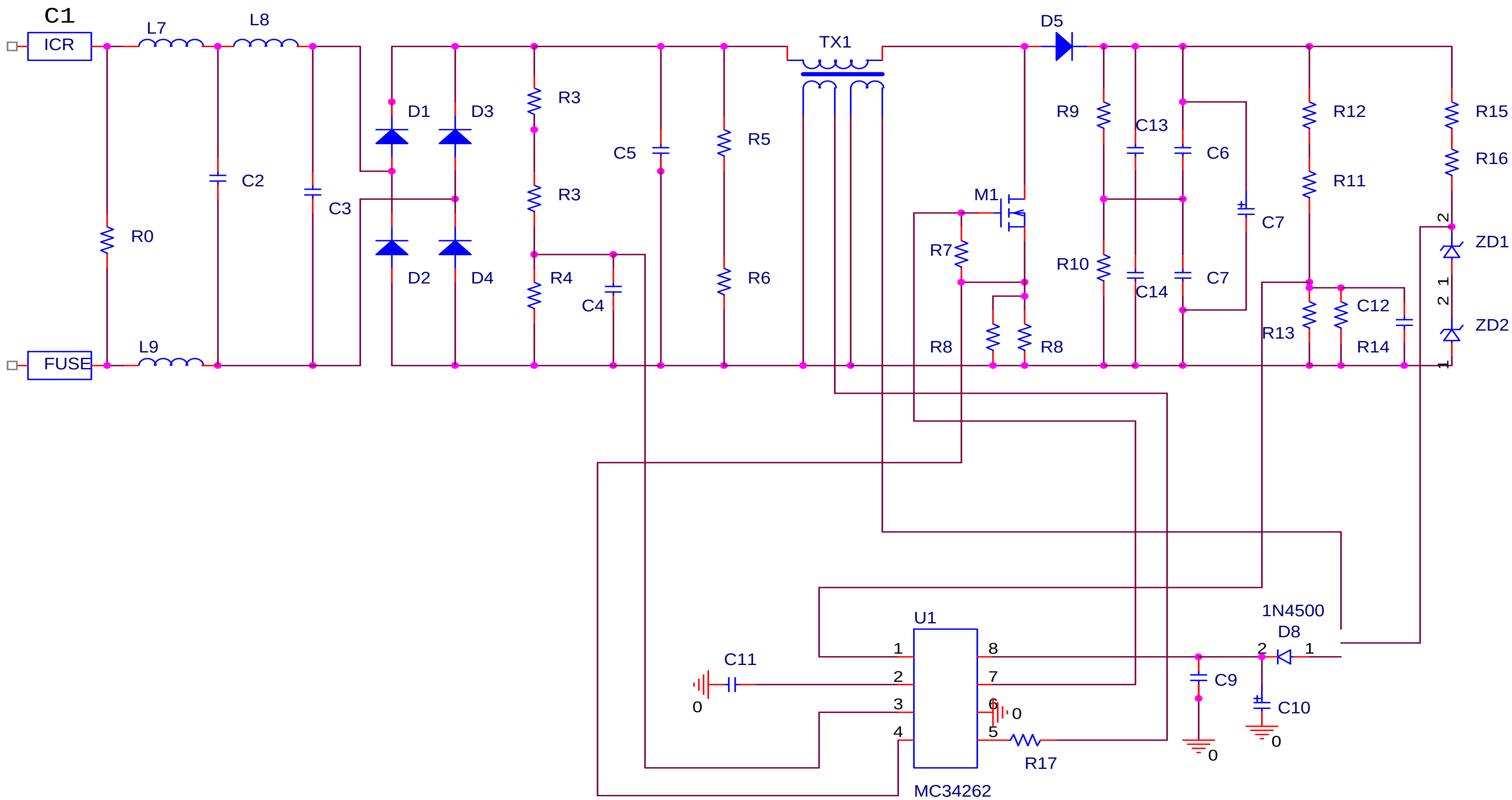


Figure 2.9.3-7: Schematic of PFC circuit

The actual waveform of the PFC correction is shown is Figure 2.9.3-8 to Figure 2.9.3-10.

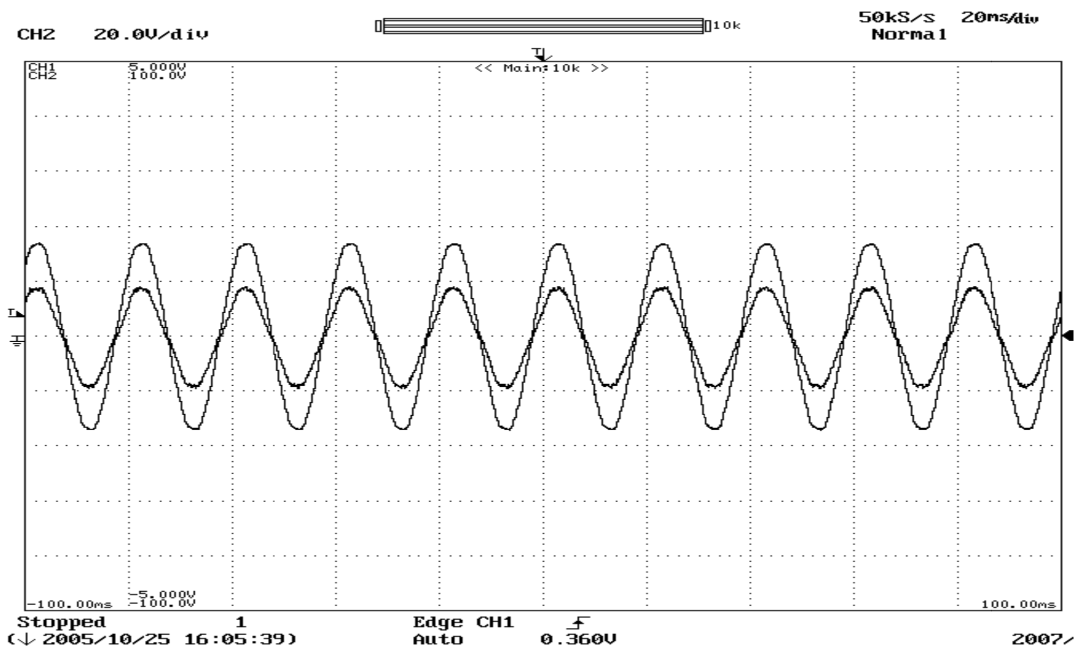


Figure 2.9.3-8: Input voltage and current

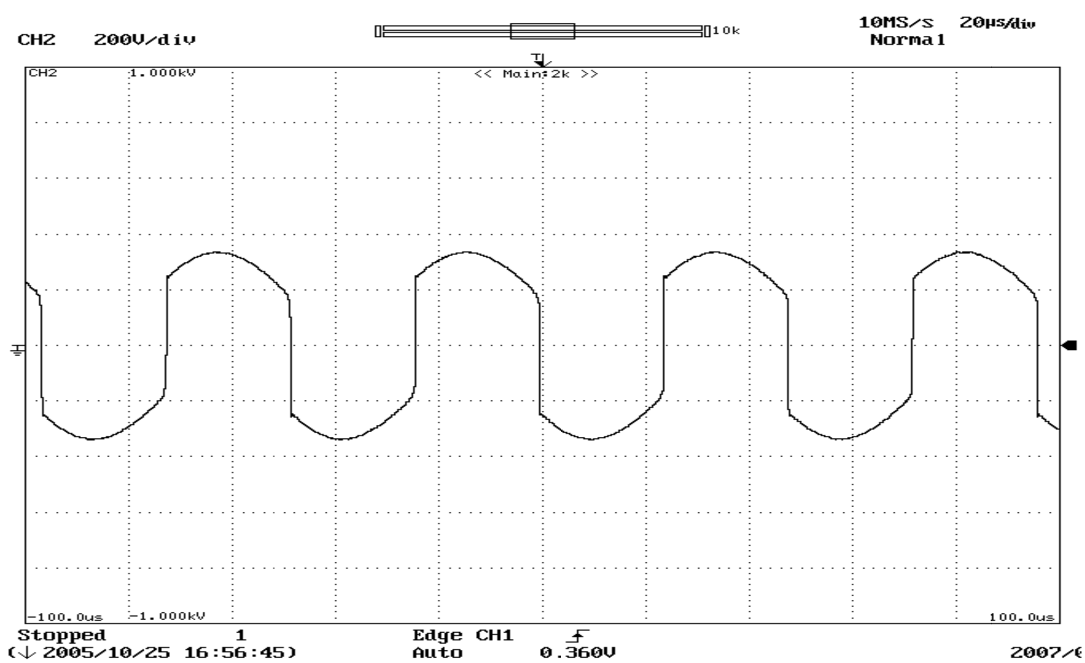


Figure 2.9.3-9: Voltage across device

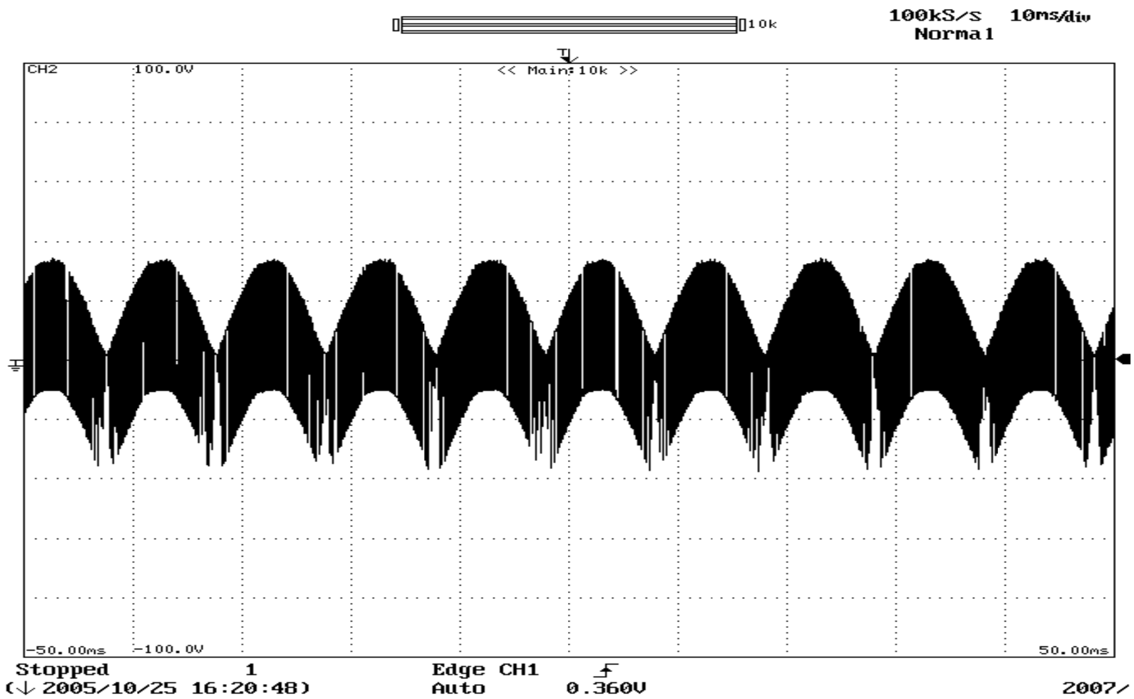


Figure 2.9.3-10: Voltage across inductor

The waveform of the rectifier without PFC circuit is shown in Figure 1.9.3-10.

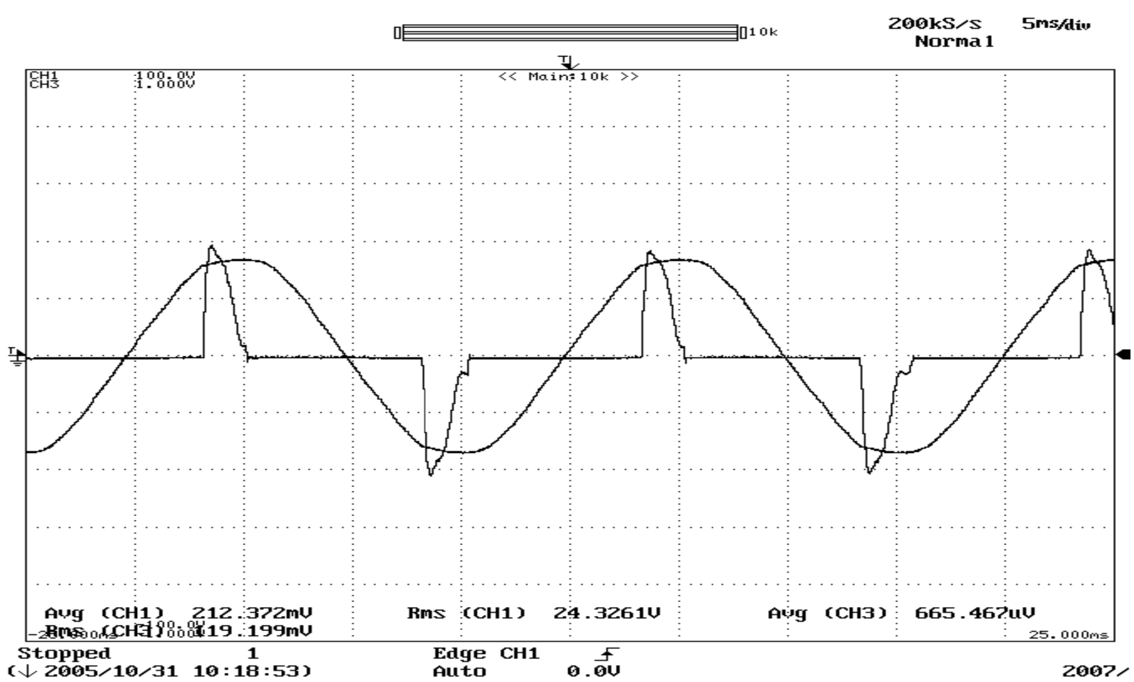


Figure 2.9.3-11: Voltage & current waveform of rectifier without PFC

The PFC was checked with the load of 150 watt HPSV electronic ballast. The performance results of 150 watt HPSV ballast without & with PFC circuit are as follows:

Table 2.9.3-1: 150 watt HPSV Ballast Without PFC Circuit

Input Volt	**Input Amp**	**Input Watt**	**Input PF**	**% THD**	**Lumens**
[V]	[A]	[W]	[λ]	[%]	Lumens
240.0	1.187	152.1	0.533 lead	150.4%	12449

Table 2.9.3-2: 150 watt HPSV Ballast With PFC Circuit

Input Volt	**Input Amp.**	**Input Watt**	**Input PF**	**% THD**	**Lumens**
[V]	[A]	[W]	[λ]	[%]	Lumens
240.0	0.784	187.7	0.998	6.02%	15724

The above results show improvement in PF from 0.533 lead to 0.998 which is closed to unity & % Total Harmonics Distortion reduces from 150% to 6.02%, which is even much better than the limits specified limits in IEC 61000-3-2 standard.

Single-phase boost converter circuit reduces the THD to 6% and improves the PF close to unity. This is much better than the limit specified by international standard IEC 61000-3-2. By improving the THD & power factor, boost converter helps to improve the power quality of the distribution system by reducing the generation of harmonics.

2.10 CONCLUSION:

In this chapter different techniques for elimination of current harmonics are discussed. It started with Passive filter design having different configuration of tuned filter. Its advantage and disadvantage along with its limitation was presented.

Different type of Shunt active filter based on application was discussed along with its control schemes. Shunt active filter to eliminate current harmonics in the system using IARP theory was presented along with simulation as well as experimental result. For detection of harmonic in the power system, mainly three techniques 1) IARP theory 2) Sine multiplication theory 3) Synchronous Reference Frame theory are used. These theories in brief was discussed but IARP theory was discussed in detail as this theory is used for other novel power quality improvement tool which are presented in subsequent chapter.

Also other techniques to improve power quality by incorporating in the power electronics tool, which are the major source of harmonic, so that this limits the generation of harmonics was presented along with simulation as well as experimental results.

Chapter-III

Series Active Filter – Solution To Voltage Harmonics And Distortion

TABLE OF CONTENTS

3.1 INTRODUCTION

This chapter introduces the technique to develop a novel three-phase series active filter to eliminate voltage harmonics and compensate reactive power based on instantaneous active reactive power theory (p-q theory). A 3-phase voltage source inverter with DC bus capacitor (center tapped) is used as a 3-phase active filter. A SPWM based control technique is employed to generate the gating signal for devices of the 3-phase active filter. Three distorted voltages are generated by putting the reactor at input side in series with Diode Bridge. Because of the current drawn by the diode bridge, the voltage drops across the series reactor. This drop in series reactor, distort the voltage at input of the diode bridge. If any linear load is connected in parallel to Diode Bridge, this load also experiences the harmonic; hence performance of the same is deteriorated. The 3-phase active filter generates the compensating voltages in the close vicinity of the desired reference voltages. The performance characteristics of the active filter is simulated and tested experimentally. The experimental result establishes the validity.

New advancement in power electronics has driven the development of ever more flexible, cost effective, high performance electrical equipment. However, such increasingly sophisticated equipment draws non-sinusoidal current from the source. Every source has an inherent impedance characteristic. There is a non linear voltage drop due to the harmonic current flowing from the source which in turn distorts the network power supply voltage. The currents harmonics, which cause the voltage distortion, lead to malfunctions, abnormal heating and vibrations in equipment, connected to network.

Power quality and reliability are essential for proper operation of industrial processes, which involve critical and sensitive loads. With the growing usage of solid-state converters in application such as adjustable speed drives and computer supplies, there has been a considerable increase in the voltage/current harmonic congestion and distortion on the power network. Short-duration power disturbances, such as voltage sags, swells and short interruptions, are major concern for industrial customer. Due to wide usage of sensitive electronic equipment in process automation e.g. semiconductor industry, traction signaling etc, even voltage sags which last for only few tenths of second may cause serious problems such as Production stoppages with considerable associated costs; Equipment restarting, Damaged or lower-quality product and customer satisfaction Malfunctioning of relays and sensors The high costs associated with these disturbances explain the increasing interest towards power quality & reliability problems mitigation techniques. The cost of mitigation intervention has to be compared with the loss of revenue and takes into account all economic factors involved. Also the proliferation of Non-linear loads such as static power converter and arc furnaces results in a variety of undesirable phenomena in the operation of power system. The most important among these are harmonic contamination, increased reactive power demand and power system fluctuations. Harmonic contamination has become a major concern of power system specialist due to its effects on sensitive loads and on power distribution system. Harmonic current components increase power system losses, causes excessive heating in rotating machinery, can create significant interference with the commutation circuits that shared common right of ways with ac power lines can generate noise on regulating and control circuits causing erroneous operation of such equipments.

Electric power generated by the utilities is distributed to the consumer in the form of 50/60 Hz ac voltage. The utilities have a tight control on the design and operation of the equipment used for transmission and distribution, and can therefore keep frequency and voltage delivered to their customers within close limits. Unfortunately, an increasing portion of loads connected to the power system are comprised of power electronic converters. These loads are non linear and inject distorted currents in the network and consequently, through line drops, they generate harmonic voltage waveforms. Power converters such as rectifiers, power supplies and at a higher power level, arc furnaces are ideal sources of distortion. According to the survey conducted by ERDA [A16] in 2000 and 2007, 35-40% of ideal electric power flows through electronic converters. This is expected to increase to 60% by the year 2020. The distortion, whether it is produced by a large single source or by the cumulative effect of many small loads, often propagates for miles along distribution feeders.

As the use of non linear power equipment is spreading, the degradation of the power quality in the utility networks is increasing and is becoming a major problem. Limiting the voltage distortion is therefore a concern for both utilities and consumers. For these reasons international agencies like IEEE and IEC (worldwide) and CEA/BIS (India) are proposing or enforcing distortion limits [G26] [G27]. The simple block diagram of Figure 3.1-1 illustrates the distortion problem due to harmonic at low and medium power levels.

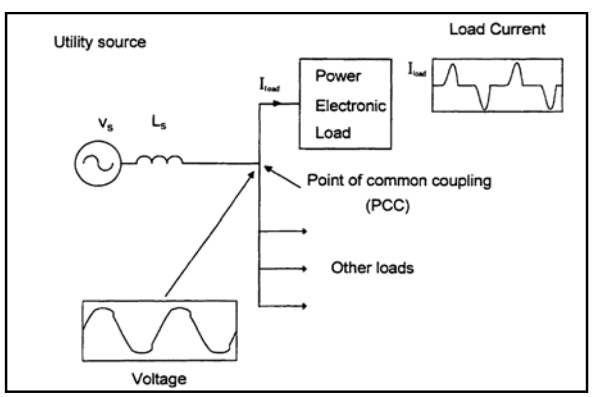


Figure 3.1-1: Harmonic distortion at PCC

The utility is represented by an ideal ac voltage source in series with lumped impedance representing lines and transformers. The voltage waveform at the point of common coupling is distorted due to harmonic current generated by the power electronic load or the load. This results in the following effects on the power system components:

(i) Malfunction of harmonic sensitive loads;

(ii) Increased losses in parallel connected capacitor, transformers, and motors;

(iii) Improper operation of protection relays and circuit breakers.

3.2 EFFECT OF HARMONICS ON SYSTEM VOLTAGES

Every source has a definite impedance characteristic and this impedance plays important role. Consider the Figure 3.2-1, which shows the impedance (lumped) of 5% with the source voltage Vac. Two different cases are explained wherein one is having linear load and other is having load. The case in which the linear load is connected, the current drawn from the source is sinusoidal in nature. This leads to sinusoidal voltage drop across the source impedance. Hence at the point of common coupling of source and load, the voltage is in sinusoidal form having no distortion.

In the second case, the Non-linear load draws non sinusoidal harmonic current (distorted current) from the source. As a result, there is a Non-linear distorted voltage drop [C17] across the source impedance. Hence the net voltage developed at the point of common coupling gets distorted. The percentage of source voltage distortion depends in the percentage of source impedance.

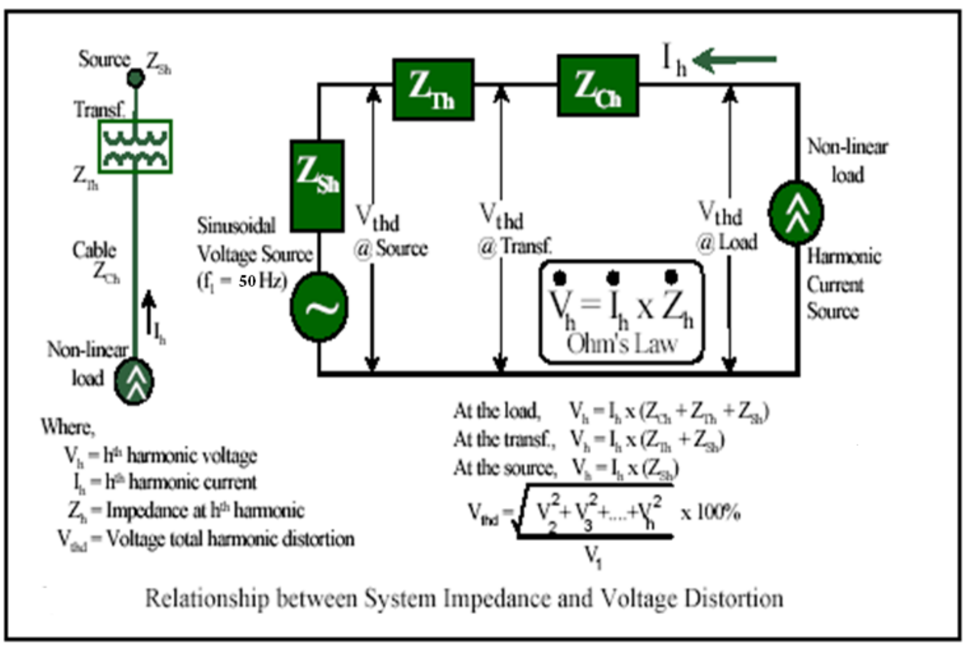


Figure 3.2-1: Relationship between system impedance and voltage distortion

3.3 SOLUTION TO SYSTEM VOLTAGE HARMONIC PROBLEMS

3.3.1 PASSIVE FILTER

Conventionally, series passive L inductor can be used to eliminate the voltage harmonics to be propagated to the load. Being connected in series in the system, the filters are to be designed at the line voltage and current ratings. However, in practical applications this also drops the fundamental voltage hence effective rms voltage to the load reduces. These filters present following disadvantages:

1) The source impedance strongly affects filtering characteristics.

2) As both the harmonics and the fundamental current components flow into the filter, the capacity of the filter must be rated by taking into account both currents.

3) When the harmonic current components increase, the capacity of the filter will be overloaded

4) Resonance between the power system and the passive filter causes amplification of the harmonic contents on the source side at a specific frequency

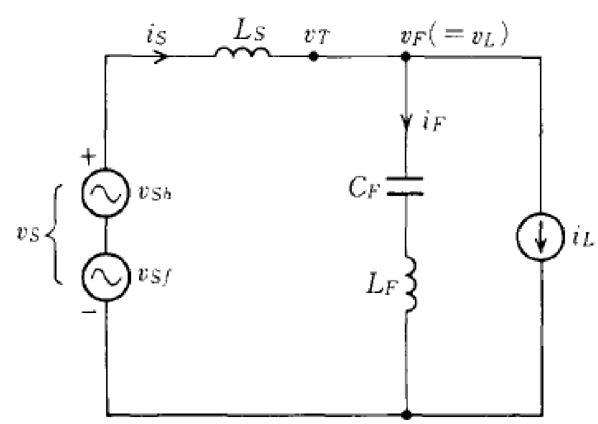


Figure 3.3-1: Basic principle of shunt passive filter

3.3.2 SERIES ACTIVE POWER FILTER

To solve the preceding problems of the shunt passive filter, shunt active filters using PWM inverters have been studied and developed in recent years. The basic principle of shunt active filters was originally presented by H. Sasaki and T. Machida in 1971 [l]. As shown in Figure 3.3-2, a shunt active filter is controlled in such a way as to actively shape the source current, *Is,* into sinusoid by injecting the compensating current, I_c. This is considered the arche type of shunt active filters. Since a linear amplifier was used to generate the compensating current, its realization is unreasonable due to low efficiency. In 1976, L. Gyugyi and E. C. Strycula [G26] presented a family of shunt and series active filters, and established the concept of the active filters consisting of PWM inverters using power transistors [G26]. However, no attention has been paid to series active filters and no experimental result has been shown in any papers, because there is no available way to shape the source current into sinusoid.

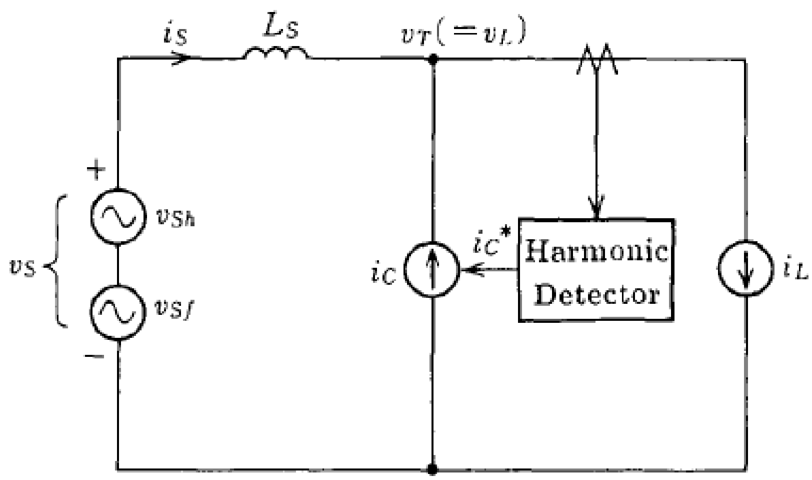


Figure 3.3-2: Basic principle of shunt active filter

In the beginning, shunt active filters were proposed to suppress the harmonics generated by large rated thyristor converters and inverters used in HVDC transmission systems. However, they could not be realized in real power systems because high-power high-speed switching devices were unavailable in the 1970's. Then N. Mohan et al. [G26] presented a practical means for injecting the compensating current, which was implemented by using naturally commutated thyristor inverters with a specially tuned passive circuit to reduce the fundamental voltage rating of the shunt active filter. However, the thyristor inverters generate undesirable high-order harmonics, which thus discourages their practicability. With remarkable development and advances in switching speed and capacity of power semiconductor devices in the 1980's, shunt active filters using PWM inverters have been studied, with a focus on their practical applications in real power systems [G27]-[G35]. At the same time, the following problems of shunt active filters have been pointed out, delaying their practical uses.

- It is difficult to realize a large rated PWM inverter with rapid current response and low loss for use as a main circuit of shunt active filters.
- The initial cost is high as compared with that of shunt passive filters, and shunt active filters are inferior in efficiency to shunt passive filters.
- Injected currents by shunt active filters may flow into shunt passive filters and capacitors connected on the power system [G28].

As is known, filtering characteristics of a shunt passive filter partially depends on the source impedance, which is not accurately known and is predominantly inductive. The impedance of the shunt passive filter should be lower than the source impedance at a turned frequency to provide the attenuation required. Hence the higher the source impedance, the better the filtering characteristics. However, the source impedance should exhibit a negligible amount of impedance at the fundamental frequency so that it does not cause any appreciable fundamental voltage drop. These two requirements, which contradict each other, can be satisfied only by inserting active impedance in series with the ac source. Also, series and parallel resonances in the shunt passive filter, which are partially caused by the inductive source impedance, can be eliminated by inserting active impedance. The active impedance can be implemented by a series active filter using voltage-source PWM inverters. Hence a new approach, which combines the use of a shunt passive filter and a small rated series active filter, is the answer to the question.

Series active filters work as isolators, instead of generators of harmonics and, hence, they use different control strategies.

3.4 A SERIES ACTIVE POWER FILTER COMBINED WITH SHUNT PASSIVE FILTER BASED ON A SINUSOIDAL CURRENT-CONTROLLED VOLTAGE-SOURCE INVERTER

The circuits of Figure 3.4-1 (a) and (b) show the block diagram and the main components, respectively, of the proposed system: the shunt passive filter, the series active filter, the current transformers (CT's), a low-power pulse width modulation (PWM) converter, and the control block to generate the *sinusoidal template* I_{ref} for the series active filter. The shunt passive filter, connected in parallel with the load, is tuned to eliminate the fifth and seventh harmonics and presents a low-impedance path for the other load current harmonics. It also helps to partially correct the power factor. The series active filter, working as a sinusoidal current source in phase with the line voltage supply V_L, keeps "unity power factor," and presents very high impedance for current harmonics.

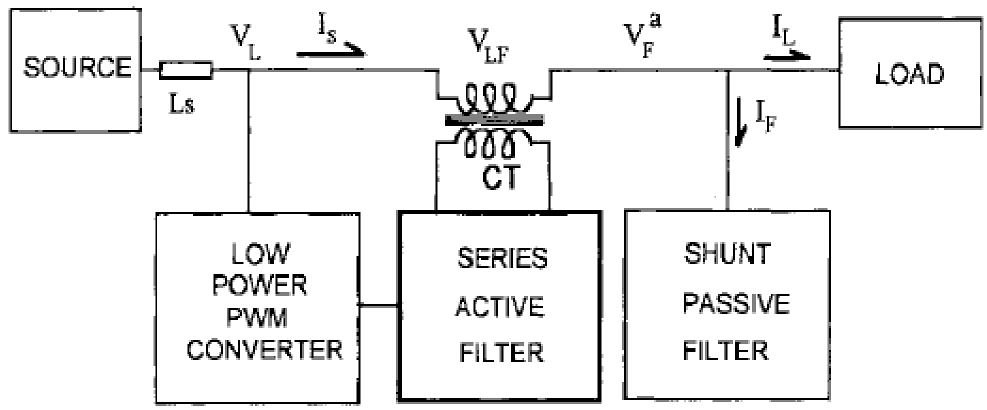


Figure 3.4-1: Block diagram showing main components of the series active filter.

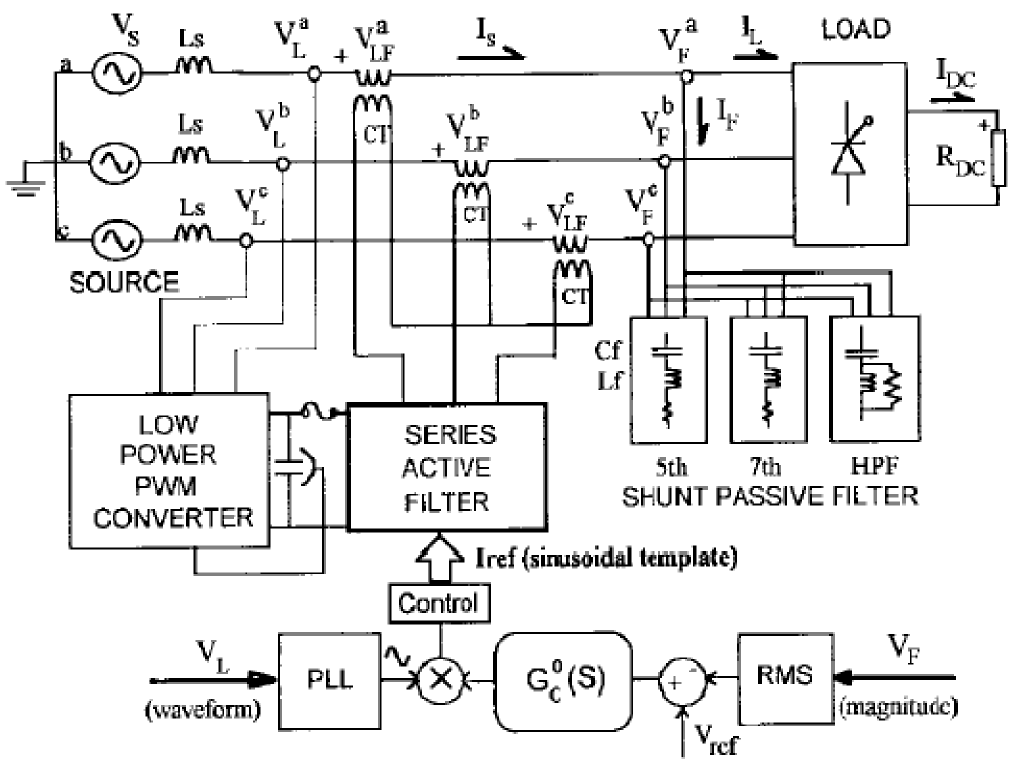


Figure 3.4-2: Components diagram of the series active filter

The CT's allow for the isolation of the series filter from the mains and the matching of the voltage and current rating of the filter with that of the power system. In Figure 3.4-2, I_L represents the *load current,* I_F the current passing through the *shunt passive filter,* and the I_S *source current*. The source current I_S is forced to be sinusoidal because of the PWM of the series active filter, which is controlled by I_{ref}. The *sinusoidal waveform* of I_{ref} comes from the line voltage V_L, which is filtered and kept in phase with the help of the PLL block Figure 3.4-1 (b).

By keeping the load voltage V_F constant, and with the same magnitude of the nominal line voltage V_L, a "zero regulation" characteristic at the load node is obtained. This is accomplished by controlling the magnitude of I_{ref} through the error signal between the load voltage V_F and a reference voltage V_{ref}. This error signal goes through a PI controller, represented by the block $G_c^0(S)$. V_{ref} is adjusted to be equal to the nominal line voltage $V_{L.}$

The two aforementioned characteristics of operation ("unity power factor" and "zero regulation"), produce an automatic phase shift between V_F and V_L, without changing their magnitudes.

3.4.1 POWER-FACTOR COMPENSATION

To have adequate power-factor compensation in the power system, the series active filter must be able to generate a voltage V_{LF} the magnitude of which is calculated through the circle diagram of Figure 3.4.1-1 according to

$$V_{LF} = 2 \,.\, V_L \,.\, \sin\frac{\text{Ø}}{2} \qquad (3.4.1\text{-}1)$$

Assuming, for example, a series filter able to generate a voltage V_{LF}, the magnitude of which is 50% of the fundamental amplitude V_L, the maximum phase shift should be approximately Φ=29°, which poses a limit in the ability to maintain unity power factor. The larger the value of V_{LF}, the larger the rating of the series active filter (kVAr). From Figure 3.4.1-1:

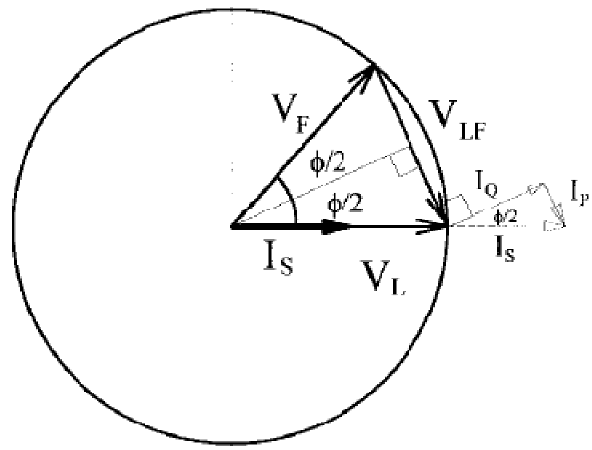


Figure 3.4.1-1: Circle diagram of the series filter

$$Q_{FILTER} = V_{LF}.I_S\,.\cos\frac{\emptyset}{2} \tag{3.4.1-2}$$

Replacing (3.4.1-1) into (3.4.1-2)

$$Q_{FILTER} \;\; = \; 2V_F.I_S\,.\sin\frac{\emptyset}{2}\,\cos\frac{\emptyset}{2}$$

$$= \; V_F\,.I_S\,.\sin\emptyset \tag{3.4.1-3}$$

Then, (3.4.1-2) corresponds to the total reactive power required by the load to keep unity-power-factor operation from the mains point of view.

It can be observed from the circle diagram of Figure 3.4.1-1 that, in order to obtain unity power factor at the line terminals (V_L), a little amount of active power has to go through the series filter.

$$P_{FILTER} = V_{LF}\,.I_P$$

$$= V_{LF}.I_S\,.\sin\frac{\emptyset}{2} \tag{3.4.1-4}$$

However, most of this active power is returned to the system through the low-power PWM converter shown in Figure 3.4-1. The amount of active power that has to go through the series active filter, according to Figure 3.4.1-1, is given by

P_{FILTER} can also be obtained through

$$P_{FILTER} = P_{LINE} - P_{LOAD}$$

$$= V_F \,.\, I_S \,(1 - \cos \text{Ø}) \qquad (3.4.1\text{-}5)$$

Equations (3.4.1-4) and (3.4.1-5) are equivalent. They are related through (3.4.1-1) and the trigonometric identity $2\, sin^2(\text{Ø}/2) = 1 - cos(\text{Ø})$

For cost considerations, it is important to keep as low as possible. Otherwise, the power ratings of both the series filter and the small PWM rectifier shown in Figure 3.4-1 become large. This means that the capability to compensate power factor of the series filter has to be restricted. The theoretical kilovolt ampere ratings of the series filter and the low-power PWM converter can be related to the kilovoltampere rating of the load (S_{LOAD}). The kilovoltampere rating of the series filter, from Figure 3.4.1-1 or from Equation (3.4.1-2) and (3.4.1-4), is

$$S_{FILTER} = V_{LF} \,.\, I_S$$

$$= 2 \,.\, V_F \,.\, I_S \,.\, \sin\frac{\text{Ø}}{2} \qquad (3.4.1\text{-}6)$$

As

$S_{LOAD} = V_F \,.\, I_S$, it yields

$$\frac{S_{FILTER}}{S_{LOAD}} = 2 \,\sin\frac{\text{Ø}}{2}$$

$$= \frac{V_{LF}}{V_S} \qquad (3.4.1\text{-}7)$$

On the other hand, the relative kilovoltampere rating of the low-power PWM converter comes from (3.4.1-5) and is

$$\frac{S_{CONV}}{S_{LOAD}} = \frac{P_{CONV}}{S_{LOAD}}$$

$$= \frac{P_{FILTER}}{S_{LOAD}}$$

$$= 1 - \cos\emptyset \qquad (3.4.1\text{-}8)$$

If it is again consider Φ=29°, it yields S_{CONV} = 12.5% of that of the power load. It can be noticed that when no power factor compensation is required, both the series filter and the small PWM converter become theoretically null. However, the small converter has to supply the power losses of the series filter (which are very small), and the series filter needs to compensate the harmonic reactive power. The low-power PWM converter is a six-pack insulated-gate-bipolar-transistor (IGBT) module, inserted into the box of the series filter.

3.4.2 HARMONIC COMPENSATION

The *kVAr* requirements of the series filter for harmonic compensation are given by

$$Q^h_{FILTER} = V^h_{LF} \,.\, I_S \qquad (3.4.2\text{-}1)$$

where V^h_{LF} is the rms harmonic voltage at the series filter terminals and I_S is the fundamental current passing through the filter. As the series filter is a fundamental current source, harmonic currents through this filter do not exist.

The harmonic compensation is achieved by blocking the harmonic currents from the load to the mains. As the series filter works as a fundamental sinusoidal current source, it automatically generates a harmonic voltage V^h_{LF} equal to the harmonic voltage drop V^h_F at the shunt passive filter. In this way, harmonics cannot go through the mains. Then, the rms value of V_{LF}^{h} can be evaluated through the harmonic voltage drop at the shunt passive filter:

$$V_{LF}^{h} = V_{F}^{h}$$

$$= \sqrt{\sum \left(V_{F}^{j}\right)^{2}} \qquad (3.4.2\text{-}2)$$

where V_{F}^{j} represents the rms value of the voltage drop produced by the j^{th} harmonic in the shunt passive filter. This voltage drop is related with the j^{th} harmonic impedance of the filter and the j^{th} harmonic current:

$$V_{F}^{j} = I_{F}^{j} \, . Z_{F}^{j} \qquad (3.4.2\text{-}3)$$

Assuming a *six-pulse thyristor rectifier load,* with a shunt passive filter like the one shown in Figure 3.4-1, the j^{th} harmonic current can be evaluated in terms of the fundamental I_S:

$$I_{F}^{j} = \frac{I_S}{j} \text{ (j = 6n ± 1, with n = 1, 2, 3 …)} \qquad (3.4.2\text{-}4)$$

Replacing (3.4.2-2)–(3.4.2-4) into (3.4.2-1) yields

$$Q_{FILTER}^{h} = (I_S)^2 \, . \sqrt{\sum \left(\frac{Z_F^j}{j}\right)^2} \qquad (3.4.2\text{-}5)$$

The impedanceZ_{F}^{j}, will depend on the parameters of the filter (C_f , L_f, R_f), and is very small for the fifth and seventh harmonics. On the other hand Z_{F}^{j}, takes a constant value for high-order harmonics (high-pass filter) and, for this reason, when j is large, the terms $\left(Z_{F}^{j}/j\right)^{2}$ in the summation in (3.4.2-5) can be neglected (j>60). With these assumptions, the term represented by the square root in (3.4.2-5), can be as small as 3%–10% of the load base impedance. Then,

$$\frac{Q_{FILTER}^{h}}{S_{LOAD}} = 3\% - 10\% \tag{3.4.2-6}$$

The small size of series filters, compared with the shunt active filters (30%-60% of S_{LOAD}), is one of the main advantages of this kind of solution. The small size of series filters also helps to keep the power losses at low values.

3.4.3 POWER LOSSES

The power losses of the series active filter depend on the inverter design. In this paper, the series filter was implemented using a three-phase PWM modulator, based on IGBT switches. With this type of power switches, efficiencies over 96% are easily reached. Then, 4% power losses can be considered for the series filter, based on its nominal kilovoltampere. Now, if the filter works only for harmonic compensation, its rating power will be between 3%–10% of the nominal load rating. Then, power losses of the series filter represent only 0.12%–0.4% (less than 1%) of that of the kilovoltampere rating of the load [4]. However, if the series filter is also designed for power-factor compensation ($\cos\Phi_{MAX}$ = 0.875 or S_{FILTER} = 0.5 S_{LOAD}), the relative power losses can be as high as 2%.

3.5 A SERIES ACTIVE POWER FILTER COMBINED WITH SHUNT PASSIVE FILTER BASED ON VOLTAGE-SOURCE INVERTER

Figure 3.5-1 shows a system configuration of the above approach to harmonic compensation.

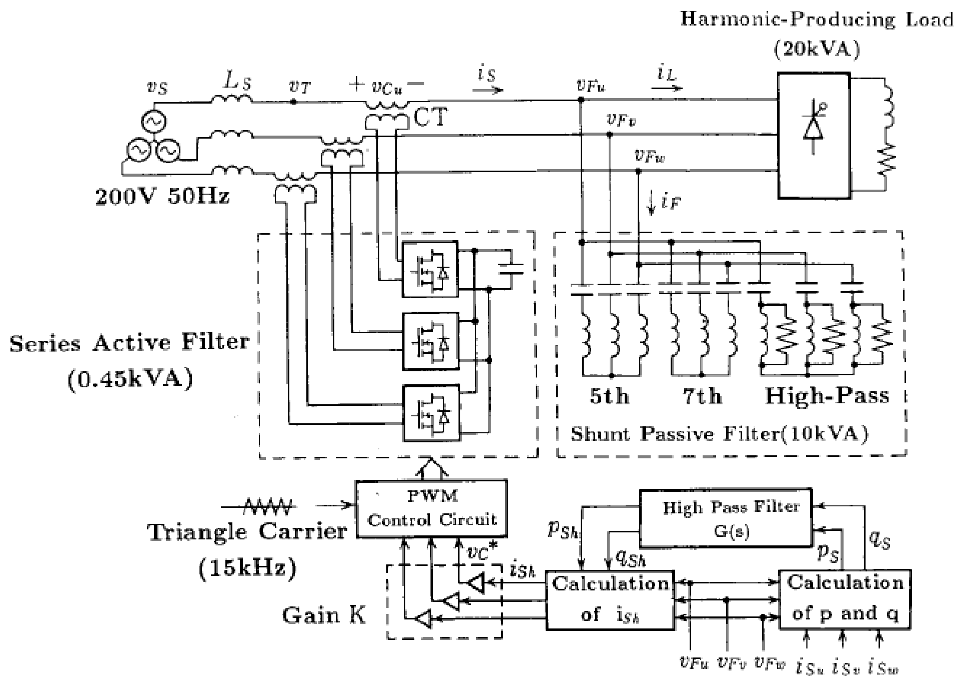


Figure 3.5-1: Circuit configuration of combined system.

Figure 3.5-2 shows a detailed circuit of a series active filter on a per-phase base. A passive filter consisting of a 5th and 7th-tuned ***LC*** filter and a high-pass filter is shunted with a three-phase six-pulse thyristor converter, which is considered a typical harmonic-producing load. The circuit constants of the shunt passive filter of rating 10 kVA are shown in Table 3.5-1. Three single-phase voltage-source PWM inverters are inserted in series with the source impedance through three single-phase current transformers (CT's; turns ratio = 1:20), thus forming a series active filter. A single-phase diode rectifier is connected on the dc side of the inverters, supplying the energy corresponding to the switching and conducting losses in the inverters.

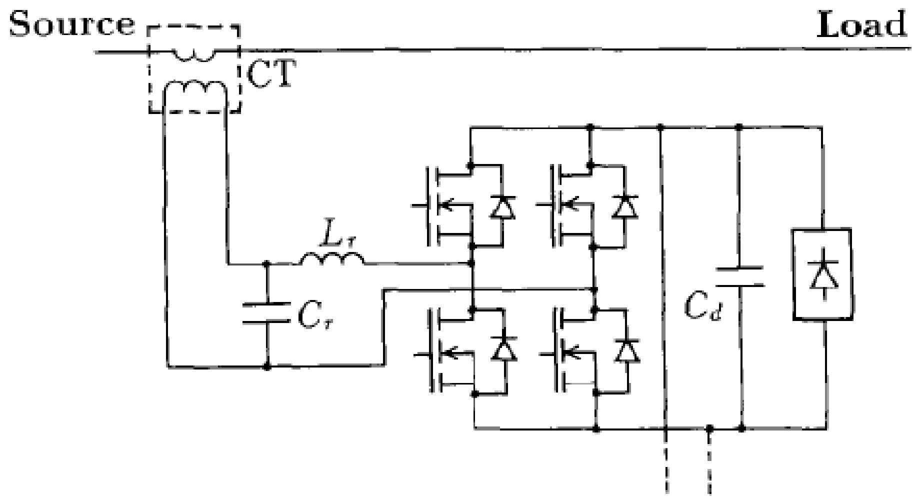


Figure 3.5-2: Detailed circuit configuration of series active filter on per-phase base

Table 3.5-1: CIRCUIT CONSTANTS OF THE SHUNT PASSIVE FILTER

Order of Harmonics Filter	Inductance	Capacitance	Quality Factor
5th	L=1.20mH	C=340 pF	14
7th	L=1.20mH	C=170 pF	14
HPF	L=0.26mH	C=300pF	R = 3Ω

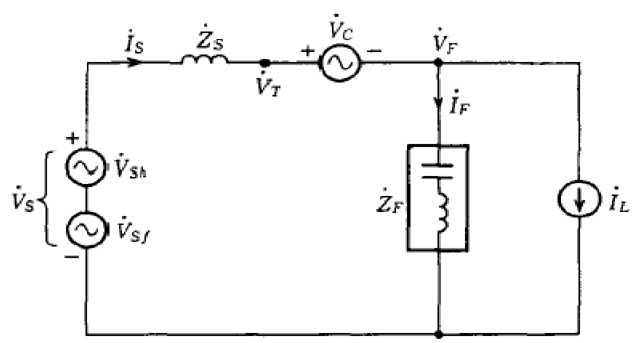


Figure 3.5-3: Equivalent circuit on per-phase base for Figure 3.5-1

The purpose of the CT's is not only to isolate the **PWM** inverters from the power system, but also to match the voltage and current rating of the **PWM** inverters with that of the power system. The function of the series active filter is not to directly compensate for the harmonics of the rectifier, but to improve the filtering characteristics of the shunt passive filter and to solve the problems of the shunt passive filter used alone. In other words, the series active filter acts not as a harmonic compensator but as a harmonic isolator. Hence the required rating of the series active filter is much smaller than that of a conventional shunt active filter.

3.5.1 COMPENSATION PRINCIPLE

Assuming for simplicity's sake that the voltage-source **PWM** inverter is an ideal controllable voltage source v_c Figure 3.5-1 is represented on a per-phase base by Figure 3.5-3**.** The three phase thyristor converter is also assumed to be a current source i_L due to the presence of sufficient inductance on the dc side. Here, $\dot{\boldsymbol{Z}}_{\boldsymbol{F}}$ is the equivalent impedance of the shunt passive filters, the constants of which are shown in Table 3.5-1, and $\dot{\boldsymbol{Z}}_{\boldsymbol{S}}$ is the source impedance.

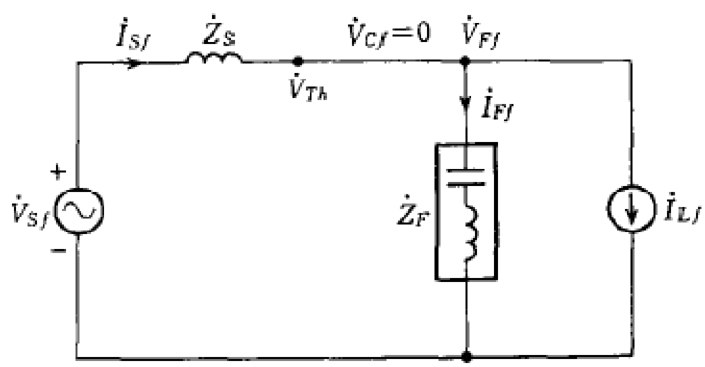


Figure 3.5-4: Equivalent circuit for fundamental frequency.

The series active filter is controlled in such a way as to present zero impedance to the external circuit at the fundamental frequency and a high resistance ***K*** to source or load harmonics.

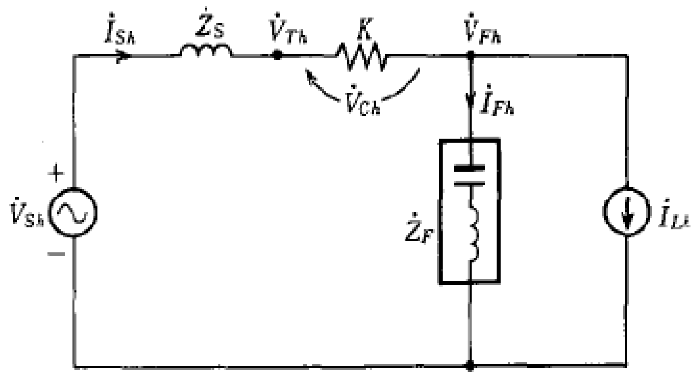


Figure 3.5-5: Equivalent circuit for harmonic frequencies.

For the fundamental and harmonics, application of the law of superposition to Figure 3.5-3 gives us two equivalent circuits, as shown in Figure 3.5-4 and Figure 3.5-5, respectively. Here, $\dot{V}_{Sf}$ is the source fundamental voltage and $\dot{\boldsymbol{I}}_{\boldsymbol{Lf}}$ is the load fundamental current in Figure 3.5-4, while $\dot{V}_{Sh}$ is the source harmonic voltage and $\dot{I}_{Lh}$ is the load harmonic current in Figure 3.5-5. As is seen in Figure 3.5-4, the shunt passive filter behaves as a capacitor for power factor improvement, but the series active filter does not play any role. Figure 3.5-5 shows that the series active filter acts as a harmonic isolator between the source and load.

To demonstrate the compensation principle of the combined system, some important equations are derived on the basis of Figure 3.5-5.

3.5.2 SOURCE HARMONIC CURRENT: $\dot{\mathbf{I}}_{\mathbf{Sh}}$

The harmonic current flowing in the source, which is produced by both the load harmonic current $\dot{\boldsymbol{I}}_{\boldsymbol{Lh}}$ and the source harmonic voltage $\dot{\boldsymbol{V}}_{\boldsymbol{Sh}}$,is given as follows:

$$\boldsymbol{I_{Sh}} = \frac{\boldsymbol{Z_F}}{\boldsymbol{Z_S + Z_F + K}} \cdot \boldsymbol{I_{Lh}} + \frac{\boldsymbol{V_{Sh}}}{\boldsymbol{Z_S + Z_F + K}} \qquad (3.5\text{-}1)$$

$$\boldsymbol{I_{Sh}} \simeq \mathbf{0} \quad \text{if K} \gg \text{Z}_S, \text{Z}_F \qquad (3.5\text{-}2)$$

Here, Z_S is the amplitude of $\dot{Z}_S$ and Z_F is that of $\dot{Z}_F$. The first term on the right side of (3.5-1) means that the series active filter acts as a "damping resistance," which can eliminate the parallel resonance between the shunt passive filter and the source impedance, while the second term means that the series active filter acts as a "blocking resistance," which can prevent the harmonic current produced by the source harmonic voltage from flowing into the shunt passive filter. If the resistance K is much larger than the source impedance, variations in the source impedance have no effect on the filtering characteristics of the shunt passive filter, thus reducing the source harmonic current to zero, as shown in (3.5-2).

3.5.3 OUTPUT VOLTAGE OF SERIES ACTIVE FILTER: $\dot{V}_C$

The output voltage of the series active filter, which is equal to the harmonic voltage appearing across resistance K in Figure 3.5-5, is given by

$$V_C = KI_{Sh} = K \cdot \frac{Z_F I_{Lh} + V_{Sh}}{Z_S + Z_F + K} \quad (3.5\text{-}3)$$

$$V_C \simeq Z_F I_{Lh} + V_{Sh} \quad \text{if } K \gg Z_S, Z_F \quad (3.5\text{-}4)$$

Equation (3.5-4) implies that the voltage rating of the series active filter is given as a vector sum of the first term on the right side, which is inversely proportional to the quality factor of the shunt passive filter, and the second term, which is equal to the source harmonic voltage.

3.5.4 FILTER HARMONIC VOLTAGE: $\dot{V}_{Fh}$

The filter harmonic voltage, which is equal to the harmonic voltage appearing across the shunt passive filter, is given by

$$\dot{V}_{Fh} = -\frac{\dot{Z}_S + K}{\dot{Z}_S + \dot{Z}_F + K} \cdot \dot{Z}_F \dot{I}_L + \frac{\dot{Z}_F}{\dot{Z}_S + \dot{Z}_F + K} \cdot \dot{V}_{Sh} \quad (3.5\text{-}5)$$

$$\dot{V}_{Fh} \simeq -\dot{Z}_F \dot{I}_{Lh} \quad \text{if } K \gg Z_S, Z_F \quad (3.5\text{-}6)$$

Equation (3.5-6) tells us that the source harmonic voltage does not appear on the load side because it applies across the series active filter.

3.5.5 CONTROL CIRCUIT

To control the series active filter in such a way as to present zero impedance for the fundamental and pure resistance for the harmonics, the reference output voltage of the series active filter is given by

$$v_c^* = K i_{Sh} \tag{3.5-7}$$

where is, is the source harmonic current, which can be calculated by applying the instantaneous real and imaginary power theory, the so-called "p-q theory" developed by H. Akagi et al. [G24].

Transformation of the phase voltages u_{Lu} , u_{Lv} , and u_{Lw} on the load side and source currents i_{su} , i_{sv}, and i_{sw}, into α-β orthogonal coordinates gives the following expression:

$$\begin{bmatrix} v_{L\alpha} \\ v_{L\beta} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{Lu} \\ v_{Lv} \\ v_{Lw} \end{bmatrix} \tag{3.5-8}$$

$$\begin{bmatrix} i_{S\alpha} \\ i_{S\beta} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_{Su} \\ i_{Sv} \\ i_{Sw} \end{bmatrix} \tag{3.5-9}$$

According to [G24], the instantaneous real power *p* and the instantaneous imaginary power *q* can be defined as

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{bmatrix} v_{L\alpha} & v_{L\beta} \\ -v_{L\beta} & v_{L\alpha} \end{bmatrix} \begin{bmatrix} i_{S\alpha} \\ i_{S\beta} \end{bmatrix} \tag{3.5-10}$$

Note that the dimension of *q* is not watt, volt-ampere, or VAr because $v_{L\alpha}$ $i_{S\beta}$ and $v_{L\beta}$ $i_{S\alpha}$ are defined by the product of the instantaneous voltage in one phase and the instantaneous current in the other phase.

The harmonic components p_h and q_h are extracted from p and q by using a high pass filter. A first-order high pass filter, the cutoff frequency of which is 35 Hz, is used. In the calculation circuit of i_{sh}, the following calculations are performed:

$$\begin{bmatrix} i_{Shu} \\ i_{Shv} \\ i_{Shw} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 \\ -1/2 & \sqrt{3}/2 \\ -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{L\alpha} & v_{L\beta} \\ -v_{L\beta} & v_{L\alpha} \end{bmatrix}^{-1} \begin{bmatrix} p_h \\ q_h \end{bmatrix} \qquad (3.5\text{-}11)$$

The reference output voltage given by (3.5-7) is compared with a triangle carrier, producing the PWM switching patterns. Here the frequency of the triangle carrier is 6 kHz. Therefore the series active filter operates as a controllable voltage source, while a conventional active filter operates as a controllable current source. Hence a voltage-source PWM inverter is suitable for the series active filter rather than a current-source PWM inverter. The voltage-source PWM inverter used in the series active filter can be protected against overvoltage and over current by the following means. All the power IGBT/MOSFET's of the upper legs are turned off to release the dc capacitor from the secondary of the CT's, while those of the lower legs are turned on to short the secondary of the CT's through the on-state power IGBT/MOSFET's and diodes.

3.6 SERIES ACTIVE POWER FILTER COMBINED WITH SHUNT PASSIVE FILTER BASED ON VOLTAGE-SOURCE INVERTER

As per the section 3.2.1 the series active filter references for compensating the harmonics are generated through indirect method. More over assumption is made that source voltage is pure sine wave which is not the case in practical application, where as the series active filter explained in section 3.2.2 is aim to reduce the rating of the active filter in combination with shunt passive filter.

However there are the case as shown in Figure 3.6-1, where substation a small feeder having linear but important load are affected due to voltage harmonic the above method is not cost effective. Normally, Feeder having linear load is drawing sinusoidal current but due the bus voltage are distorted it draws nonsinusoidal current. In such case a series active filter which can isolated linear load from voltage harmonics is the only solution which is presented in this thesis.

To simplify the control strategy for series active filters, a simple approach is presented, i.e., the series filter is controlled as a voltage source whose reference is get through *" p-q theory". This approach presents the following advantages.*

1) The control system is simpler & instantaneous compensation is possible for elimination of the voltage harmonics.

2) Rating of the active filter is small as it restricted for particular feeder only.

3) It controls the voltage at the load node, to compensate the effect of blocking the voltage harmonics, allowing excellent regulation characteristics.

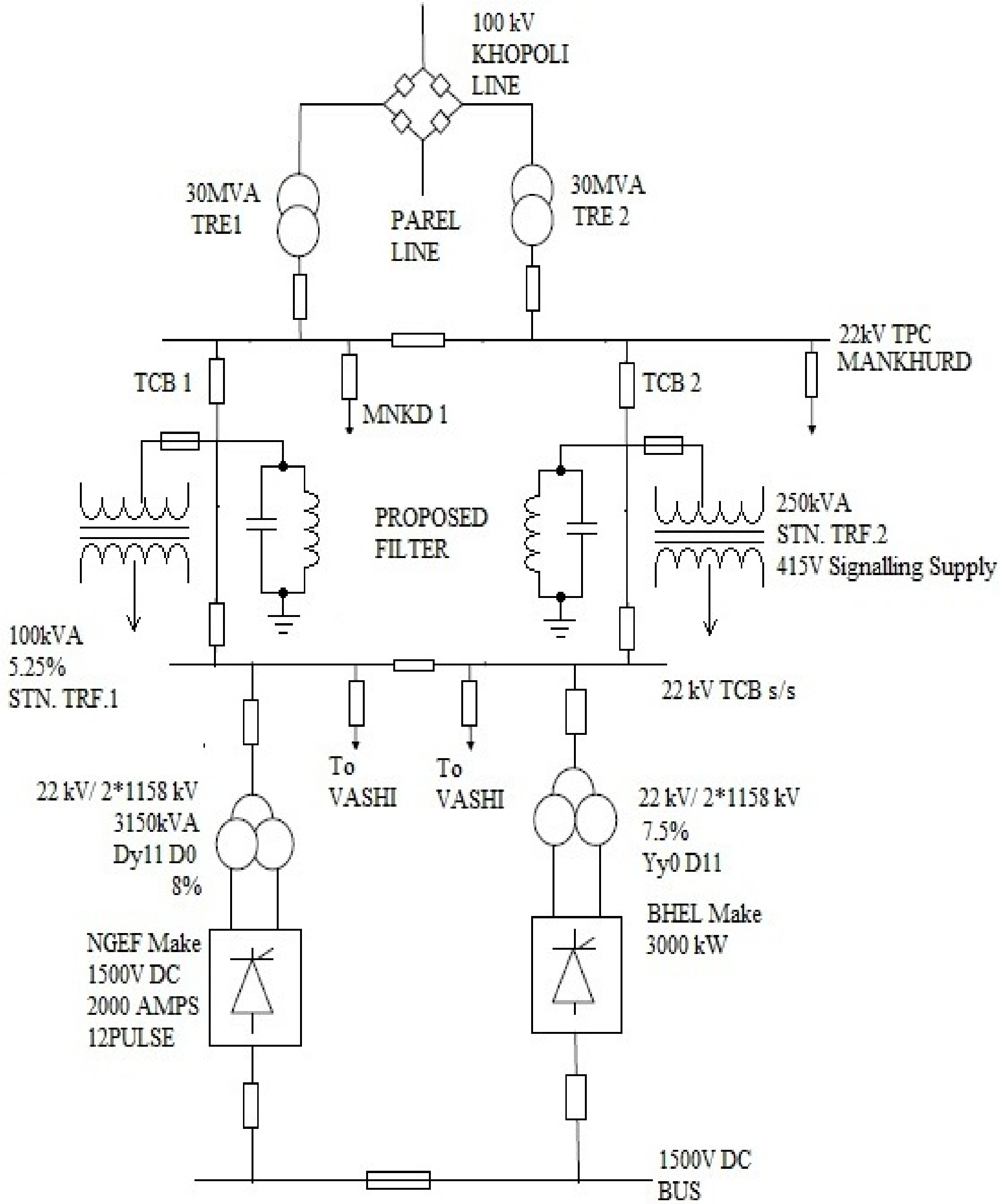


Figure 3.6-1: Traction substation having small linear load

3.7 NOVEL SERIES ACTIVE POWER FILTER COMBINED WITH SHUNT PASSIVE FILTER BASED ON VOLTAGE-SOURCE INVERTER USING ANALOG APPROACH

In this section, detailed description of control and power circuit realized using analog approach is described. Each block describes working principle and actual circuit realization. As previously described the system is based on instantaneous reactive power theory.

According to instantaneous reactive power theory six signals (three reference voltages and three distorted voltages) are required for the calculation of the three compensation voltages. These voltages signals are derived from the phase voltage using potential transformers.

3.7.1 CONTROL CIRCUIT APPROACH & IT'S REALIZATION

3.7.1.1 SIGNAL CONDITIONING CIRCUIT

Three-phase distorted source voltages are fed as input to the control cards through PT's. These signals are required to be conditioned to make them compactible for the calculation of compensating reference voltages. The reference voltages are generated by filtering all other components except the fundamental component at the control level. Due to the filtering of the distorted signal, a little phase shift is generated between the distorted input and the output sinusoidal signal. This phase difference is eliminated using phase shifter circuit.

3.7.1.2 TRANSFORMATION OF SIGNALS INTO ORTHOGONAL COORDINATES:

By p-q theory these voltages (v_a , v_b , & v_c) and (v_{ar} , v_{br} , & v_{cr}) are transformed into α-β coordinates (orthogonal coordinates) by following relations.

$$\begin{bmatrix} v_{0r} \\ v_{\alpha r} \\ v_{\beta r} \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{ar} \\ v_{br} \\ v_{cr} \end{bmatrix} \tag{3.7.1-1}$$

$$\begin{bmatrix} v_{0} \\ v_{\alpha} \\ v_{\beta} \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{a} \\ v_{b} \\ v_{c} \end{bmatrix} \tag{3.7.1-2}$$

After solving matrices following was obtained

$$v_{0r} = 0.5773\, v_{ar} + 0.5773\, v_{br} + 0.5773\, v_{cr}$$

$$v_{\alpha r} = 0.8165\, v_{ar} - 0.4082\, v_{br} - 0.4082\, v_{cr}$$

$$v_{\beta r} = 0.0\, v_{ar} - 0.7071\, v_{br} - 0.7071\, v_{cr} \tag{3.7. 1-3}$$

Further same approach can be used to determine v_α and v_β Therefore, following was obtained

$$v_{0} = 0.5773\, v_{a} + 0.5773\, v_{b} + 0.5773\, v_{c}$$

$$v_{\alpha} = 0.8165\, v_{a} - 0.4082\, v_{b} - 0.4082\, v_{c}$$

$$v_{\beta} = 0.0\, v_{a} - 0.7071\, v_{b} - 0.7071\, v_{c} \tag{3.7.1-4}$$

These transformation equations are used for the practical circuit realization. The corresponding circuit using operational amplifiers for each Equation is shown in Figure 3.7.1-1

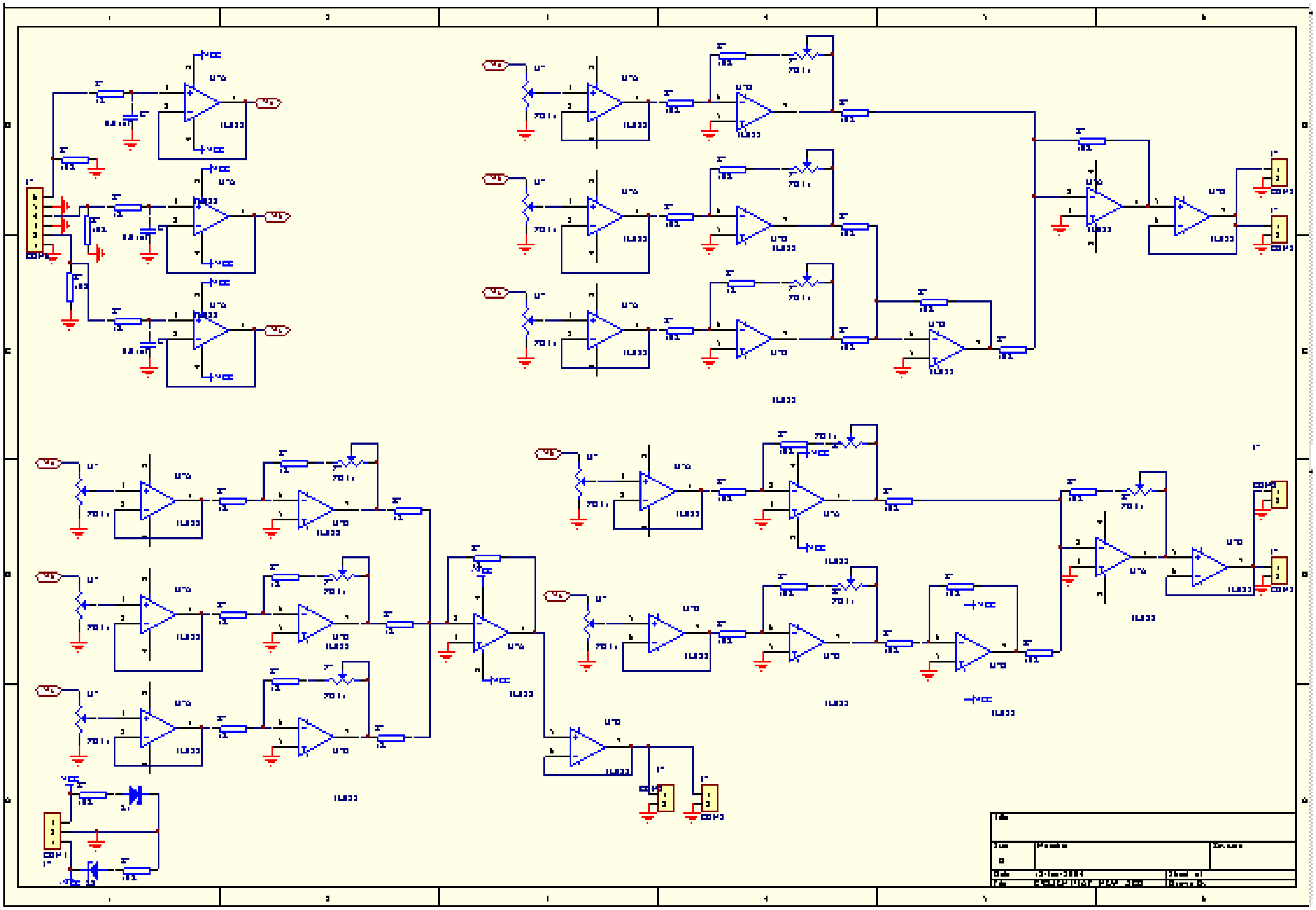


Figure 3.7.1-1: Transformation of voltage Signals into Orthogonal Coordinates

3.7.1.3 INSTANTANEOUS REAL AND IMAGINARY POWER CALCULATIONS

The resultant of these equations i.e. orthogonal reference voltages and distorted voltages are used to define instantaneous real and power.

By p-q theory instantaneous real and imaginary power are defined as

$$\begin{bmatrix} p_0 \\ p \\ q \end{bmatrix} = \begin{bmatrix} v_{0r} & 0 & 0 \\ 0 & v_{\alpha r} & v_{\beta r} \\ 0 & -v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} v_0 \\ v_\alpha \\ v_\beta \end{bmatrix} \tag{3.7.1-5}$$

Solving following was obtained,

$$p_0 = v_{0r}\, v_0$$

$$p = v_{\alpha r}\, v_\alpha + v_{\beta r}\, v_\beta$$

$$q = v_{\alpha r}\, v_\beta - v_{\beta r}\, v_\alpha \tag{3.7.1-6}$$

In these equations distorted voltage is multiplied with reference voltage so for practical multiplication purpose AD633JN multiplier is used and remaining addition and subtraction is done by Op-Amps. The diagram of practical circuit is shown in Figure 3.7.1-2.

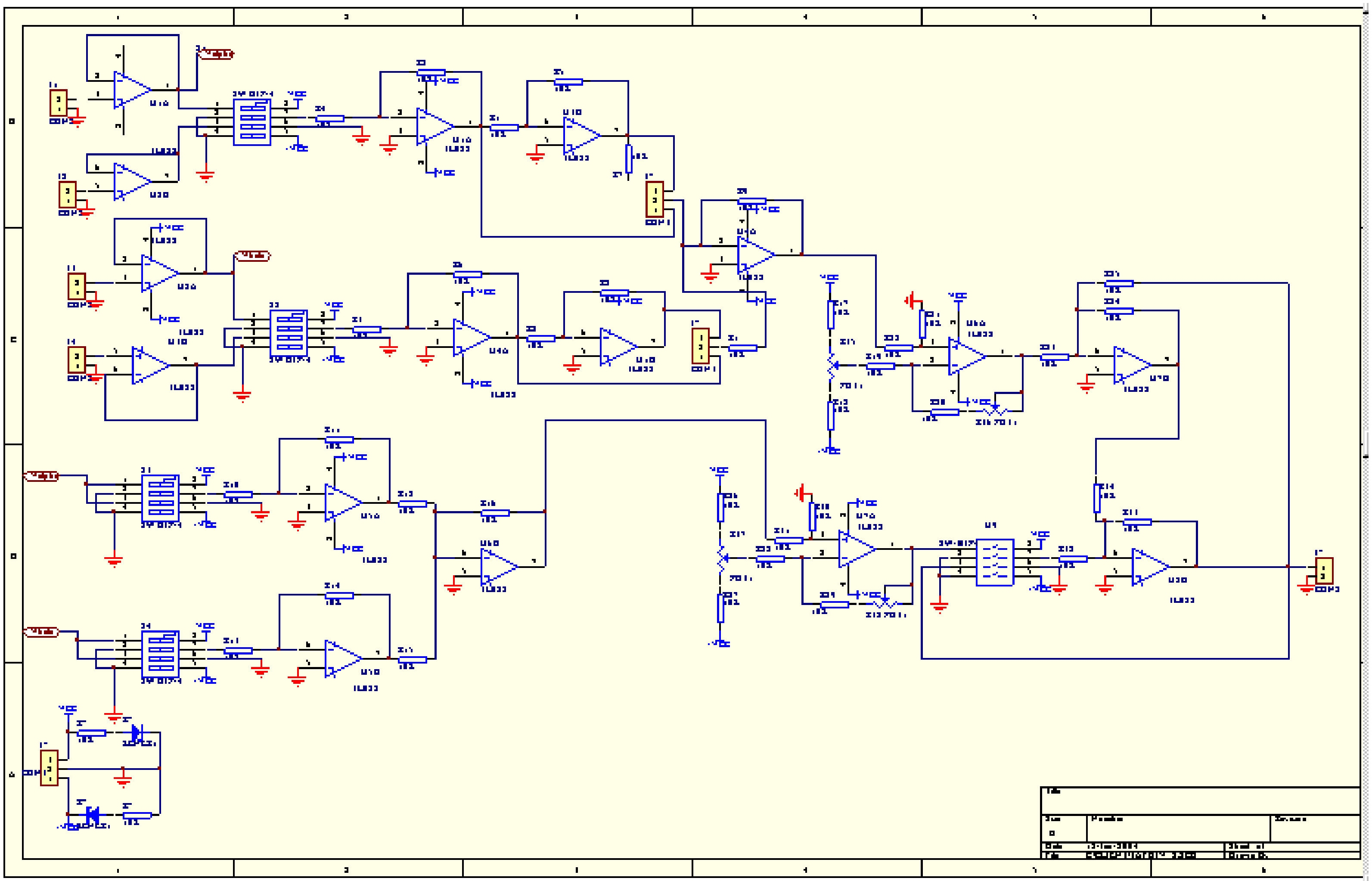


Figure 3.7.1-2: Instantaneous Real And Imaginary Power Calculations

3.7.1.4 FILTER CIRCUIT CONSIDERATION AND ORTHOGONAL COMPENSATION VOLTAGES

To eliminate harmonics and the calculated real power p is passed through high pass second order Butterworth filter. The cutoff frequency for the filter is 45Hz. In this system cascading two first order filters makes this second order filter. The corresponding circuit is shown in Figure 3.7.1-3. By using the filter the AC component is eliminated giving,

$$p_{ac} = p - p_{dc}$$

$$q_{ac} = q - q_{dc} \tag{3.7.1-7}$$

The filter used introduces phase in the input and output signal. To compensate this phase shift a phase shifter is used in the system. Further, to compensate reactive power must be made negative in matrix equations,

$$\begin{bmatrix} v_{\alpha c} \\ v_{\beta c} \end{bmatrix} = \begin{bmatrix} v_{\alpha r} & v_{\beta r} \\ -v_{\beta r} & v_{\alpha r} \end{bmatrix}^{-1} \begin{bmatrix} -p_{ac} \\ -q_{ac} \end{bmatrix}$$

$$\begin{bmatrix} v_{\alpha c} \\ v_{\beta c} \end{bmatrix} = \frac{1}{(v_{\alpha r}{}^2 + v_{\beta r}{}^2)} \begin{bmatrix} v_{\alpha r} & -v_{\beta r} \\ v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} -p_{ac} \\ -q_{ac} \end{bmatrix} \tag{3.7.1-8}$$

Thus, to cancel harmonics in the power system and reactive power the p_{ac} and q_{ac} is made negative. Further solving following was obtained,

$$v_{\alpha c} = \frac{1}{(v_{\alpha r}{}^2 + v_{\beta r}{}^2)} \left(-v_{\alpha r}\, p_{ac} + v_{\beta r} q_{ac}\right)$$

$$v_{\beta c} = \frac{1}{(v_{\alpha r}{}^2 + v_{\beta r}{}^2)} \left(-v_{\beta r} p_{ac} - v_{\alpha r} q_{ac}\right) \tag{3.7.1-9}$$

Where $v_{c\alpha}$ and $v_{c\beta}$ are compensating voltages in the orthogonal coordinates. The corresponding realization is shown in Figure 3.7.1-3.

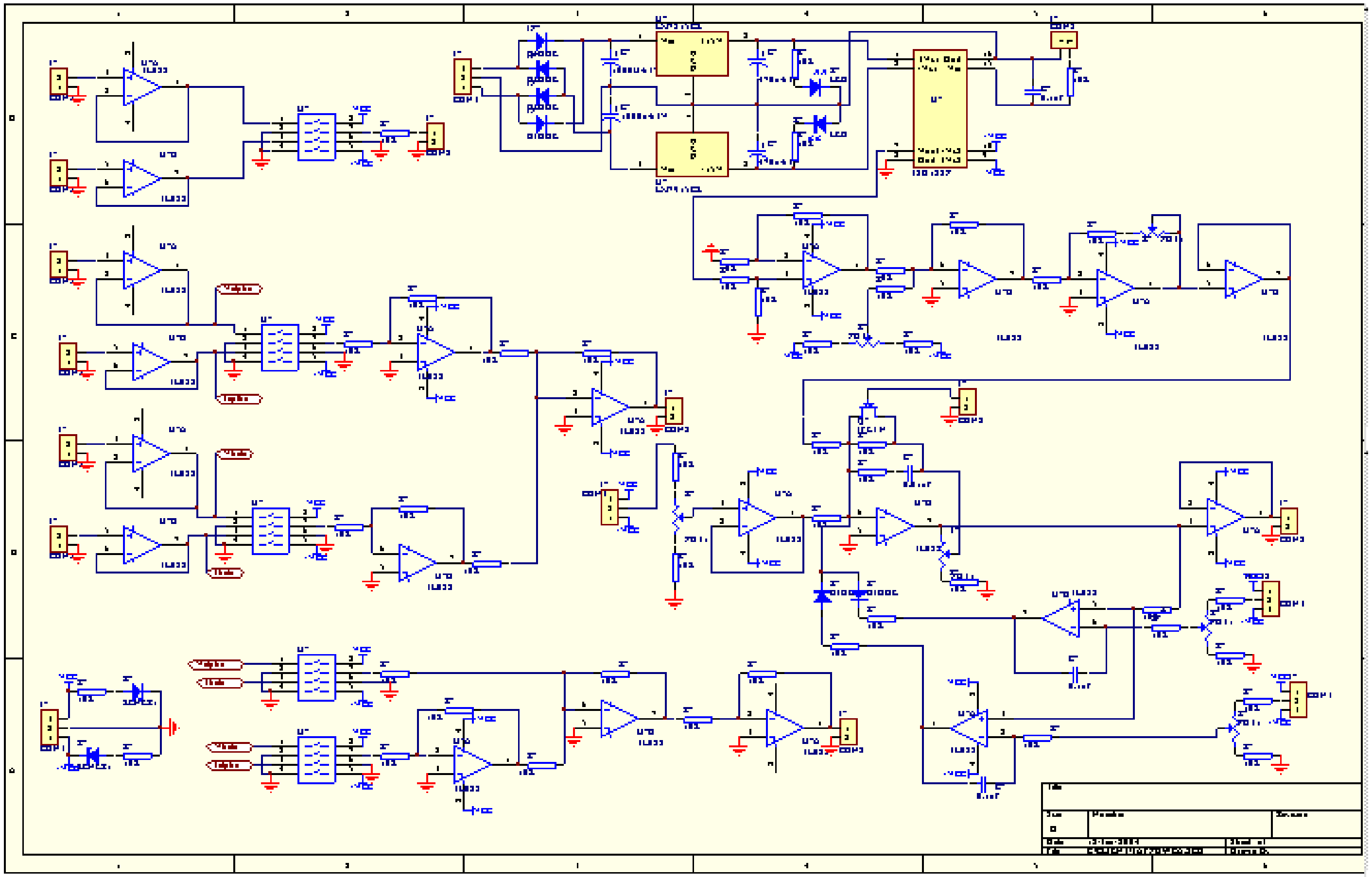


Figure 3.7.1-3: $v_{c\alpha}$ and $v_{c\beta}$ calcultation

3.7.1.5 TRANSFORMATION OF ORTHOGONAL VOLTAGES BACK INTO THREE PHASE VOLTAGES

This $e_{c\alpha}$ and $e_{c\beta}$ compensated orthogonal voltages are transformed back into three phase voltages by reverse transformation of matrix as

$$\begin{bmatrix} v_{ac} \\ v_{bc} \\ v_{cc} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1 & 0 \\ 1/\sqrt{2} & -1/2 & \sqrt{3}/2 \\ 1/\sqrt{2} & -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_0 \\ v_{\alpha c} \\ v_{\beta c} \end{bmatrix} \tag{3.7.1-10}$$

By solving these Equation following was obtained

$$v_{ac} = 0.5773\, v_{0c} + 0.8165\, v_{\alpha c} + 0.0000\, v_{\beta c}$$

$$v_{bc} = 0.5773\, v_{0c} - 0.4082\, v_{\alpha c} + 0.7071\, v_{\beta c}$$

$$v_{cc} = 0.5773\, v_{0c} - 0.4082\, v_{\alpha c} - 0.7071\, v_{\beta c} \tag{3.7.1-11}$$

These e_{ac}, e_{bc}, e_{cc} are compensated three phase reference compensating voltages. The corresponding realization is shown in Figure 3.7.1-4

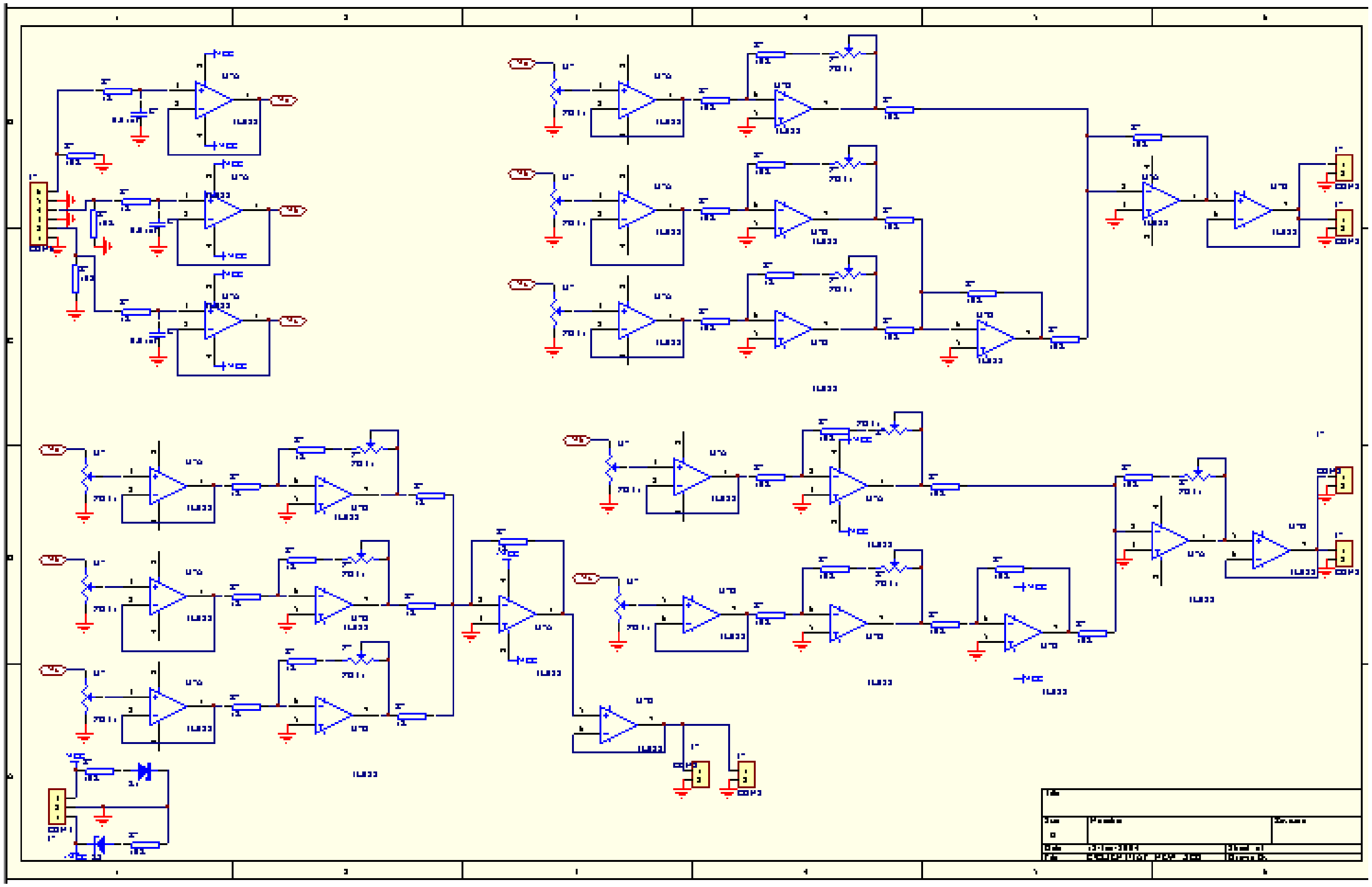


Figure 3.7.1-4: Transformation of Orthogonal voltage Back into Three Phase voltage

3.7.1.6 COMPENSATING VOLTAGE GENERATION

The compensation voltages derived from previous block is used to generate gate signals to eliminate harmonics.

The implementation of series type active filter is simpler if the type of inverter used is "SPWM Voltage Source" type. The SPWM voltage source inverter employs a SPWM switching technique in which the output voltage of the inverter is pulse width modulated type and is due to switching pattern formed due to the comparison of the reference compensating voltages and triangular wave of 4.8 kHz.

3.8 ISOLATION AND DRIVER CIRCUIT

The gate pulses, which are to be given to, the power device goes through two stages: isolation and driver stage. The isolation is must when gate pulses are given to power devices when connected in inverter configuration. Otherwise there is possibility of short circuit in low power circuitry by entering high power from power circuit when device gets short-circuited. The isolation is normally provided to gate pulses through opto-couplers, which isolates power circuit and low power circuit optically. Further these optically coupled signals are given to the driver circuit. The gate pulse, which is given to the IGBT, is with respect to the source terminal .So to avoid short circuit of drivers four separate power supplies are designed for each driver.

Since the power circuit consists of six devices and lower DC bus is common to all devices in lower limbs of inverter so only single power supply is necessary for lower limb drivers but it must have current rating thrice than the upper limb driver power supply.

The driver circuit consists hybrid IC, opto-coupler and isolated power supply. The main advantage of using hybrid IC driver is that it is made as per IGBT specification and all protection is provided in the same IC. IC M57962L is used as the hybrid driver IC and has the provision to sense the V_{ce} of the IGBT and monitor continuously voltage across the collector and emitter. When gate pulse applied to the hybrid IC and within 10 μ second if the device is not turn ON then this gives a signal which is useful for instantaneous over current protection. The Vce rises in the IGBT when access current is flow through it.

3.9 POWER CIRCUIT

The schematic diagram of series converter is shown in Figure 3.10-1 with load circuitry. This converter consists of six power IGBT, capacitor (C1). The filter capacitor has the function of suppressing the harmonic currents generated by the switching operation of IGBT. The capacitors (C1) maintain the DC bus voltage constant. These passive components size depends on the switching frequency and rating of the active filter. Diode bridge rectifier with resistive load is considered as Non-linear load (harmonic generating source). The RC snubber circuit is used for dv/dt protection of the IGBT. The value of the same is 10 ohm and 0.01 micro -farad capacitor. These snubbers are connected across each device. This snubber absorbs the peak voltage generated due to stray inductance and limits the dv/dt of the IGBT during turn off. Same way RC snubber is connected across the diode bridge module which is used to charge the DC bus capacitor and supply the active energy absorbs by the IGBT of voltage source inverter, reactor and snubber. Bleeder resistor is connected across the DC bus capacitor to balance the voltage within the voltage rating. The bleeder resistor are necessary as two capacitor of 6900μF, 450V DC, with ripple current of 10 amp are connected in series to increase the voltage rating up to 900V DC. The series active filter consists of three dual modules each module having two IGBT hence total six IGBTs in three-phase three-wire inverter configuration.

The values of components used along with active filter are :

C1, C2 = 6900 μF (4700 μF, 450 VDC Capacitor two in series and such three branch connected in parallel ALCON make with center point of each branch common)

IGBT= SKM50 GB123 B (50 amp, 1200 V, dual Module, with anti-parallel diode)

Driver IC = M57962L

The series active power filter is tested at 5 amp, 415Vac, and 50Hz three-phase supply and DC voltage of 350 V_{dc}.

The power circuit mainly consists of inverter, a high frequency transformer, LC filter. The high frequency transformer plays important role in the series compensation. The secondary of this transformer is connected in series between the source and the linear load. Generally it is found that the THD content in the source cannot be more than 20%. Actual measurements show that the harmonic content is not more than 10%. Considering this as a base for deciding the voltage of harmonic content, the rating of the high frequency transformer are fixed as follows:

Voltage ratings	:0-110-150//0-110-150volts per phase
Current rating	:15 amps per phase
Configuration	:three phase
VA ratings	:2250 VA per phase

Insulation Level	:3 kV
Leakage reactance	:10%

The frequency of switching is kept 4.8 kHz. The PWM signals (generated as a result of comparison of triangular wave with reference compensating voltages) are fed as gate signals to the IGBTs through the driver circuit. Thus the output of the inverter contains the components of harmonic contents plus the switching frequency. As the switching frequency component is not required, it is filtered out using the tuned LC filter.

3.10 EXPERIMENTAL SET-UP FOR TESTING

For the testing of the series active filter, artificial voltage harmonics are generated in the laboratory using a inductor and a diode bridge rectifier. The point of coupling of inductor and the diode bridge is the source of generating voltage harmonics. These distorted waveforms are utilized as input for the verification of p q theory. The PTs with 440/3.0 Volt ratio and burden of 0.2 VA are used for sensing the distorted voltages. The control circuit and power circuit are tested under the harmonic conditions. The THD was 20% approximately. The results are given in section 3.12 of this chapter

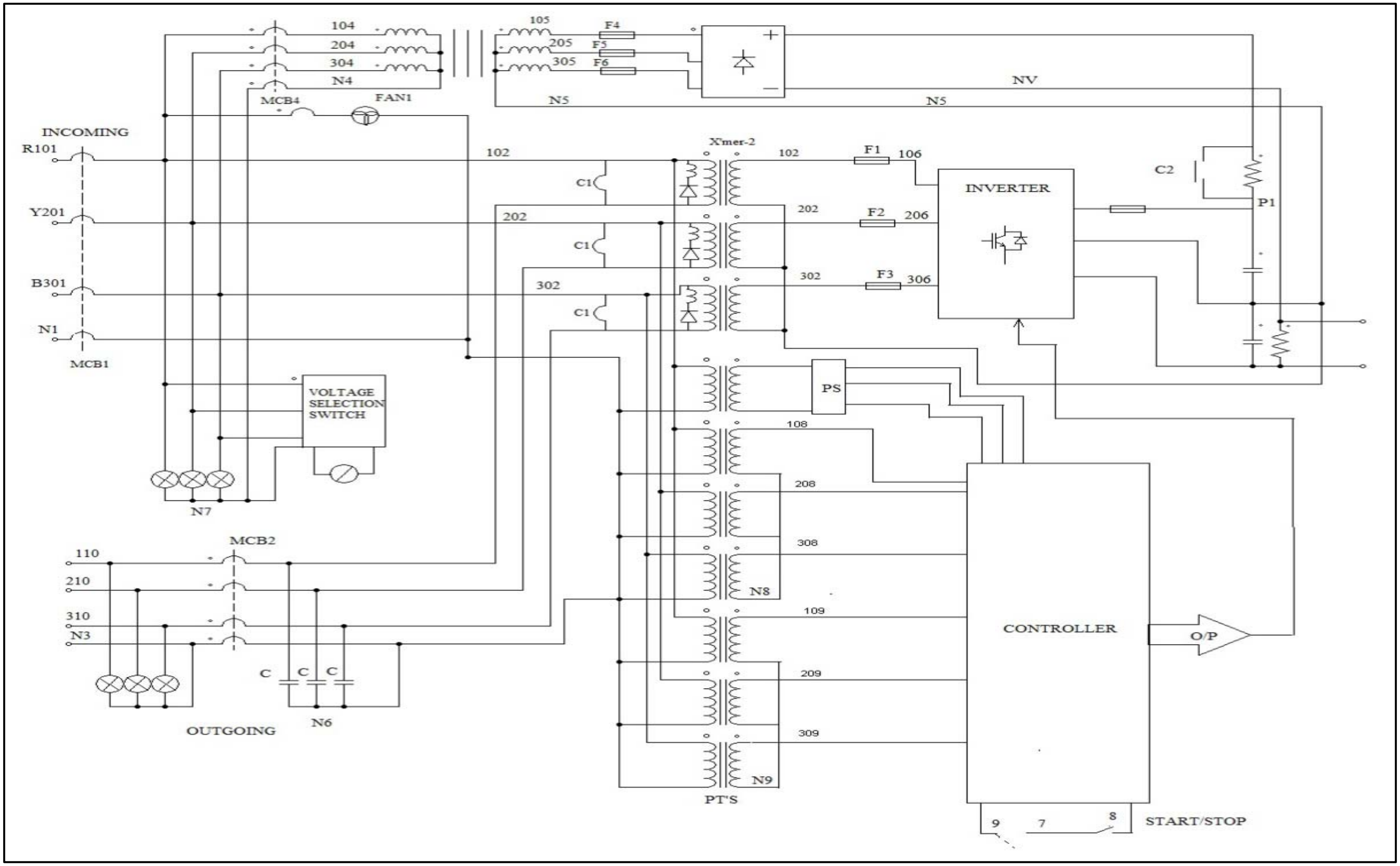


Fig 3.10-1: Schematic diagram of Series Active Filter developed

3.11 BRIEF DESCRIPTION OF WORKING OF SERIES ACTIVE FILTER

Initially the series active filter was checked for compensation of voltage harmonics at low current. First the artificial voltage harmonics are generated using the inductor and diode combination. The THD content of the source voltage is measured using the wave analyzer and these signals are fed to the control cards for generation of compensating voltages for elimination of harmonics. Bipolar triangular wave of 4.8 kHz frequency is generated using IC 8038 to imply the SPWM technique for switching of the inverter. The reference compensating voltages are compared with the triangular wave of 4.8 kHz frequency and accordingly the switching pulses are generated. Thee pulses are then fed to the driver card.

A separate three-phase transformer is used for generating DC for charging the capacitor with the help of a three phase diode bridge. A toggle switch is kept for giving and removing the gate pulse for the inverter. The output voltage of the inverter is dependent on the magnitude of the reference compensating voltages. This output of the inverter serves as the input to the transformer. The internal impedence of the transformer is kept 10%. This impedence in addition to the tuned LC filter serves the purpose of filter for eliminating the switching frequency component from the output of the transformer. The secondary side of this transformer is connected in series between the source and the linear load. The output of this transformer compensates the voltage harmonics in the source by injecting the negative of the harmonic signals.

The sequence of switching on and off and the protection scheme for inrush of currents during turning on of the active filter is explained henceforth.

Timer-relay logic has been developed to switch on the active filter. The inrush of currents during the sudden application of the DC voltages to the inverter damages the inbuilt freewheeling diodes of the IGBTs modules. To avoid these inrush currents, ERDA has implemented a soft charging circuit using resistors and timer-contactor combination. The timer is set to operate after six seconds and the charging resistor (R) is so selected that the charging time is equal to RC time constant. After six seconds the resistor is bypassed through a contractor.

The mode of operation of start/ stop of the active filter is as follows:

1) Manual mode
2) Auto mode

When SW1, SW2, and SW3 (please refer the schematic diagram and diagram of logic for switching on of active filter) is in ON position, the manual start / stop logic is bypassed. Whenever mains supply fails, whole system is switched off. At that time, position of all MCB's is in ON condition as the system is in operating condition. When the mains supply resumes, this energizes transformer-1 and transformer-2 and control logic gets ac supply. The DC capacitor of the inverter is energized through Resistor. After 6 sec, which is the time set in the timer; the contactor C2 gets energized and by passes the charging resistor. The contactor C2A is used to separate the DC and AC arc interruption. The contactors C2 and C2A both energizes simultaneously. The C2A contactor enables the start logic and as the SW1 is in ON position the RL1 gets energized which gives start command to Series Active filter and then energizes the C1 contactor. The C1 contactor is normally closed contactor and hence when C1 contactor energizes, all the contact becomes open this enables the Series active filter, which is already started, to compensate for voltage harmonic.

This sequence can be bypassed to manual mode depending upon the position of the SW1, SW2 and SW3. The detail control and power logic is given in the attached drawing. Figure 3.11-1 shows the control logic diagram of switching on series active filter. Figure 3.11-2 shows the front view of the panel developed at ERDA. Figure 3.11-3 & 3.11-4 shows the general allocation of components within the developed panel.

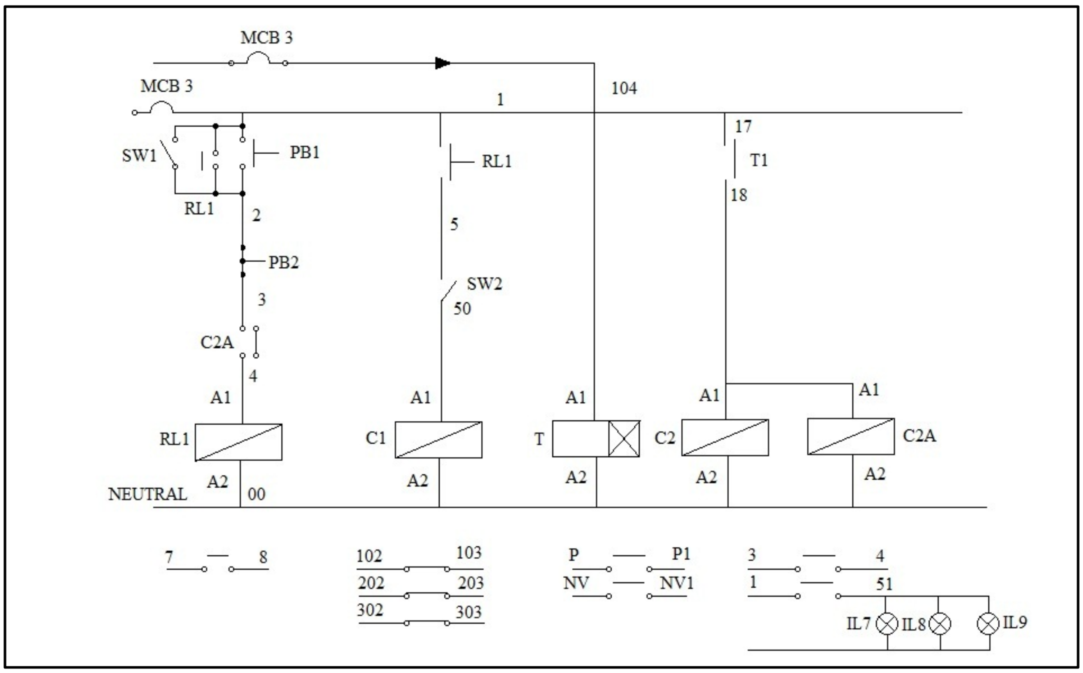


Figure 3.11-1: Control logic diagram of switching sequence of series active filter

Figure 3.11-2: Photograph of front view of the panel

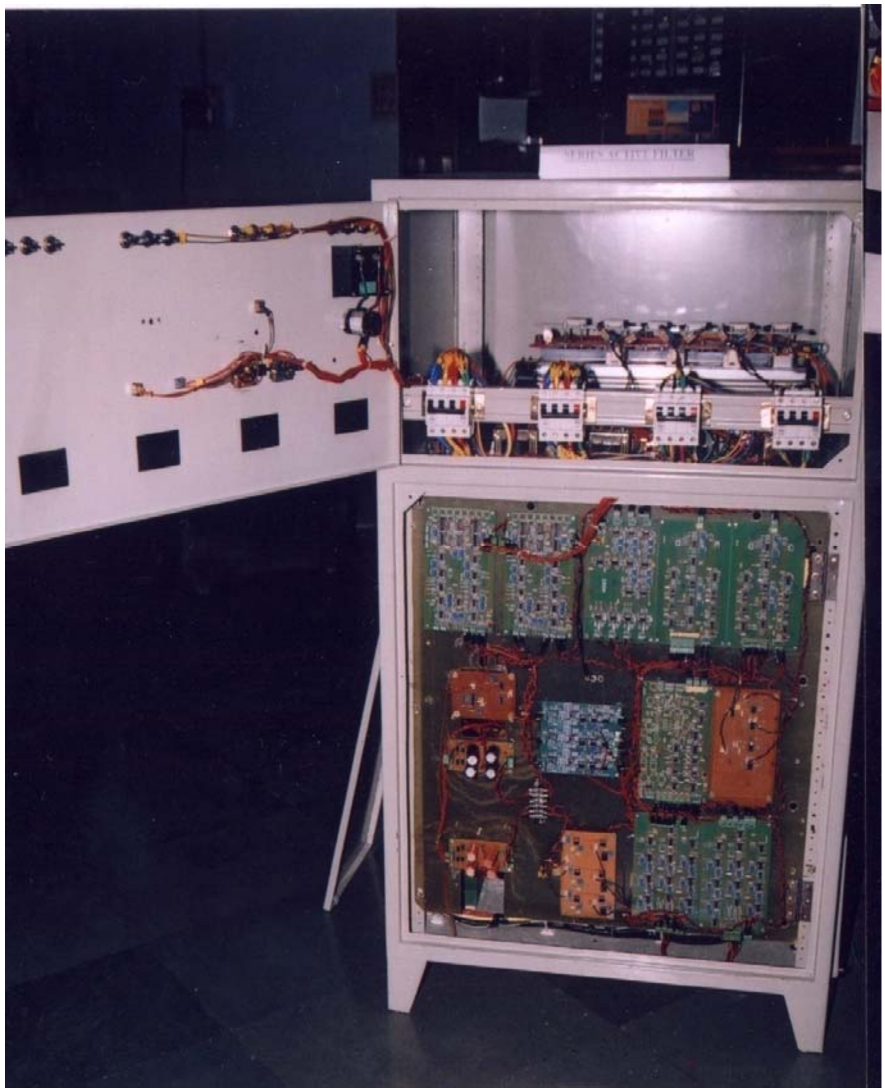


Figure 3.11-3: Photograph of components on the front side of the panel

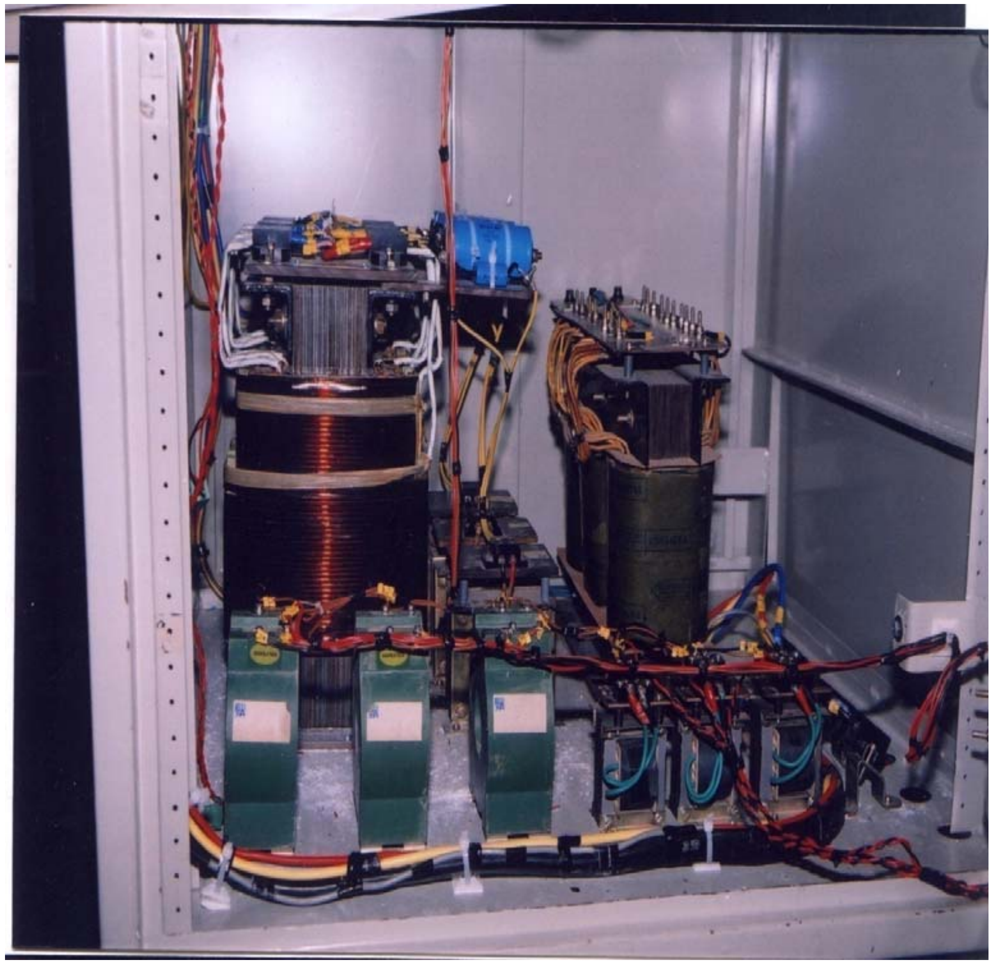


Figure 3.11-4: Photograph of the components mounted on bottom flange

3.12 RESULT

The simulations as well as experimental results are given here.

3.12.1 SIMULATION RESULTS

A complete model of the series active filter for 5 amps, 350V DC with input voltage of 440V and 5-amp load was implemented using MATLAB (SIMULINK) and the most important results will be presented to compare actual and simulated results. The fundamental frequency of the system is 50 Hz. The wave-form of source voltage, source voltage, load current and compensating voltage is observed for a-phase, b-phase, & c-phase using scope available in the simulation tools. These results shows that series active filter inject compensating voltage in series with load such that total harmonic voltage of the load becomes zero.

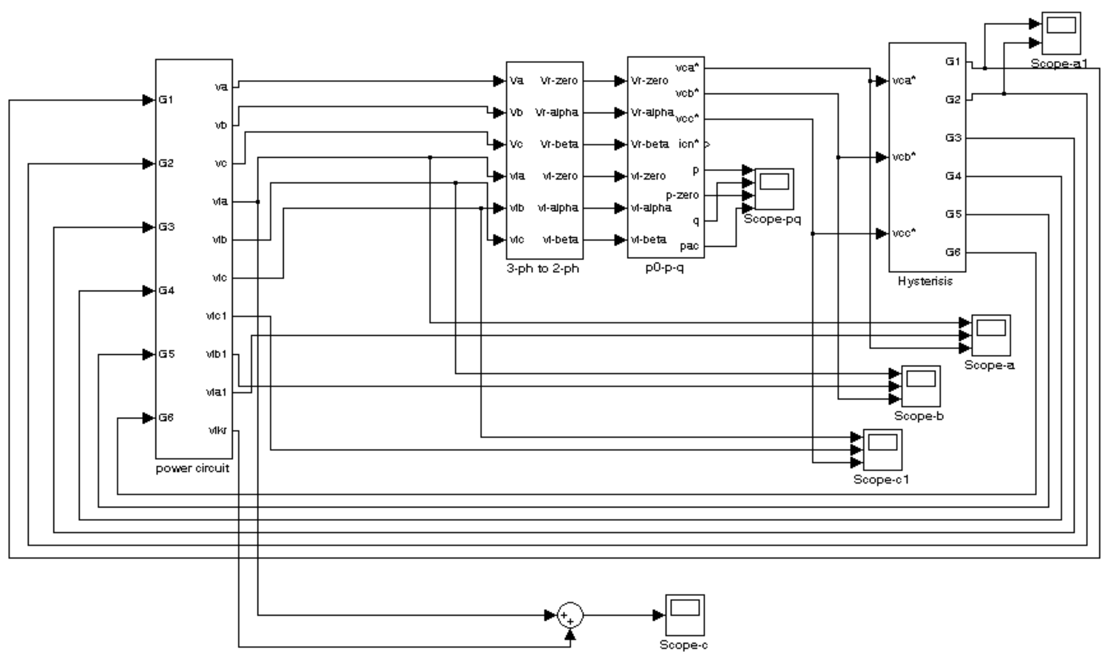


Figure 3.12.1-1: Simulation Block Diagram Used For Series Active Filter Simulation

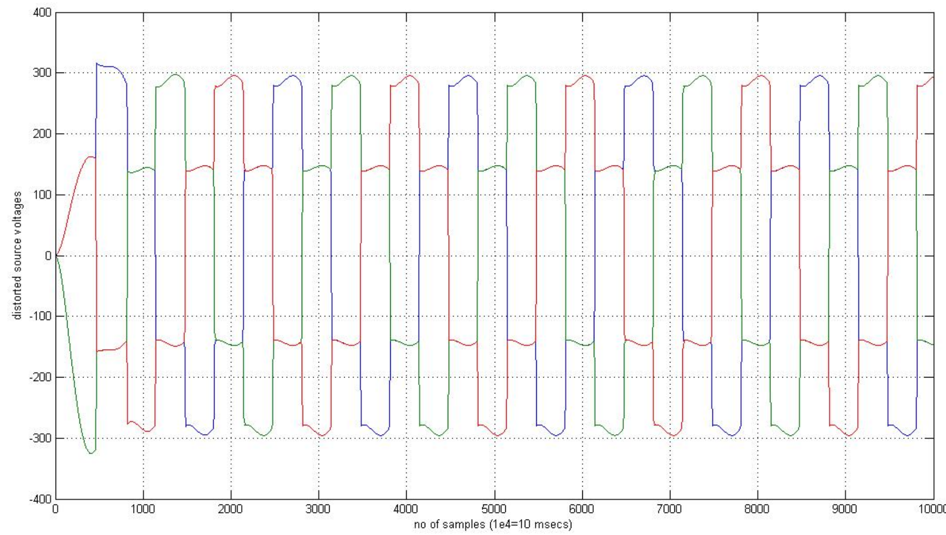


Figure 3.12.1-2: Distorted waveform at load bus before compensation

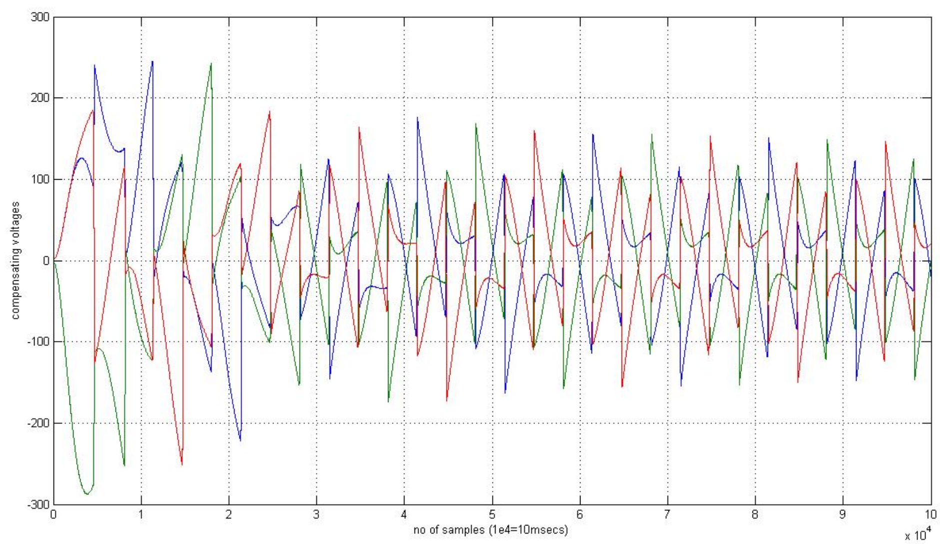


Figure 3.12.1-3: compensation waveform of series active filter to compensate the distortion in voltage

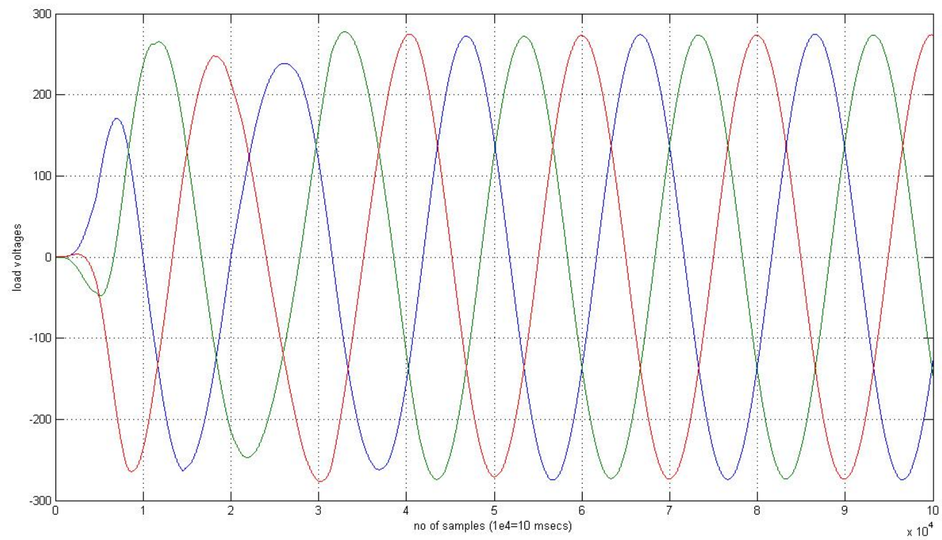


Figure 3.12.1-4: waveform at load bus after compensation through series active filter

In ideal case the capacitor voltage should remain constant but in practice every components have some losses hence DC capacitor voltage reduces. So DC Bus voltage needs to be controlled. In this scheme DC bus voltage are not controlled through inverter. To control the DC bus voltage a separate isolated rectifier having output DC voltage of 800 volts is connected in parallel to DC bus capacitor. Input of rectifier is having step up transformer with step up ratio of 1.3. The input of rectifier transformer is connected at input of series active filter as shown in Figure 3.10-1

3.12.2 EXPERIMENTAL RESULTS

Figure 3.10-1 shows the complete experimental setup of three-phase series active filter module. Three inductors of 25 mH each in series with three-phase diode rectifier system with resistive load are used as harmonic affected source for experimental purpose.

The initial testing results indicate that there is decrease in RMS value of the load voltage due to harmonic elimination from the load voltage after filtration. To avoid this decrement in the output load voltage, some percentage of the fundamental voltage is added along with harmonic compensation voltage at the control level.

The corresponding waveforms of three phase compensated load voltages and waveform at different places are shown at the end of this chapter.

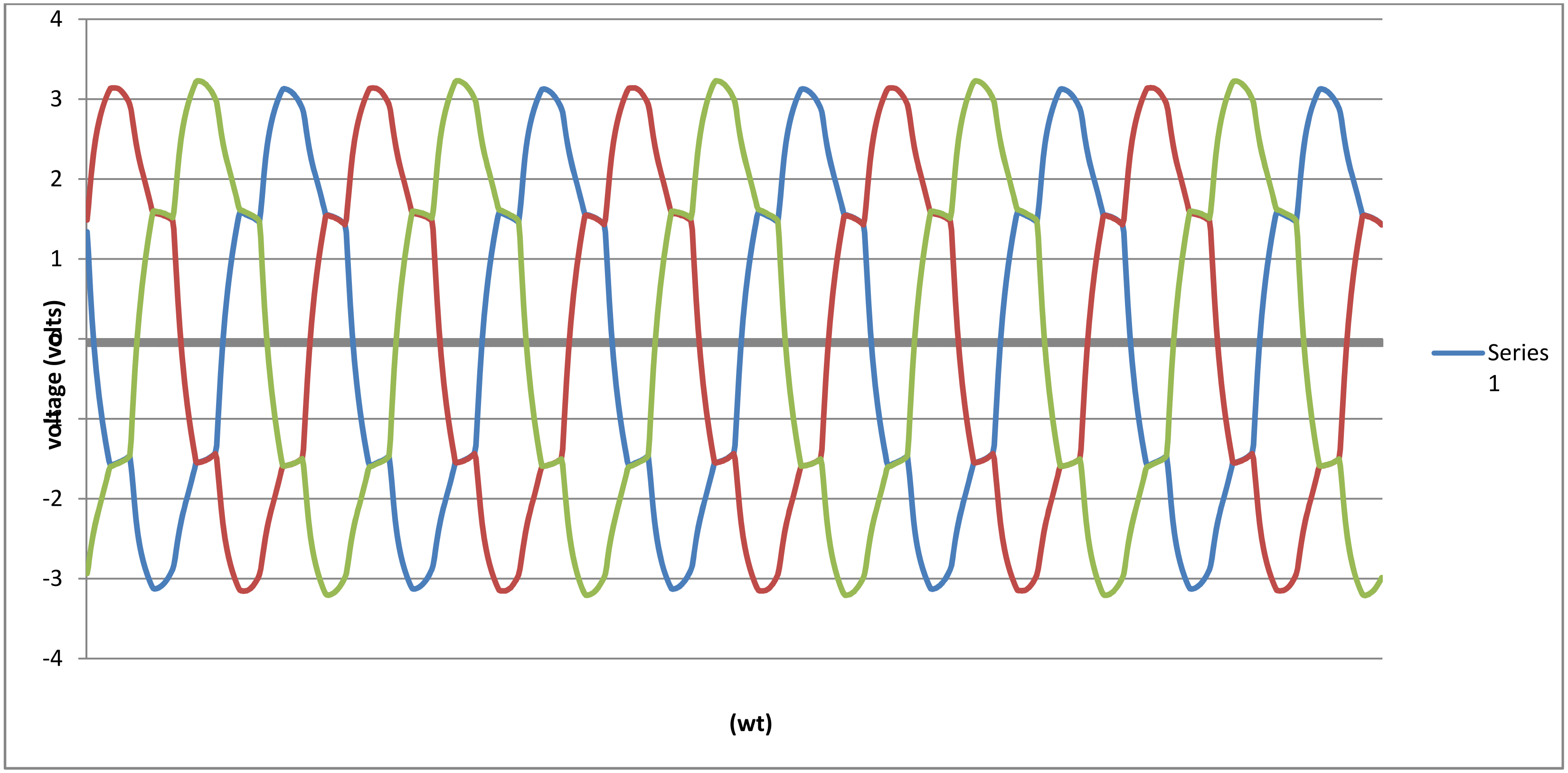


Fig 3.12.2-1 Load voltage waveform for r-y-b phase (without compensation)

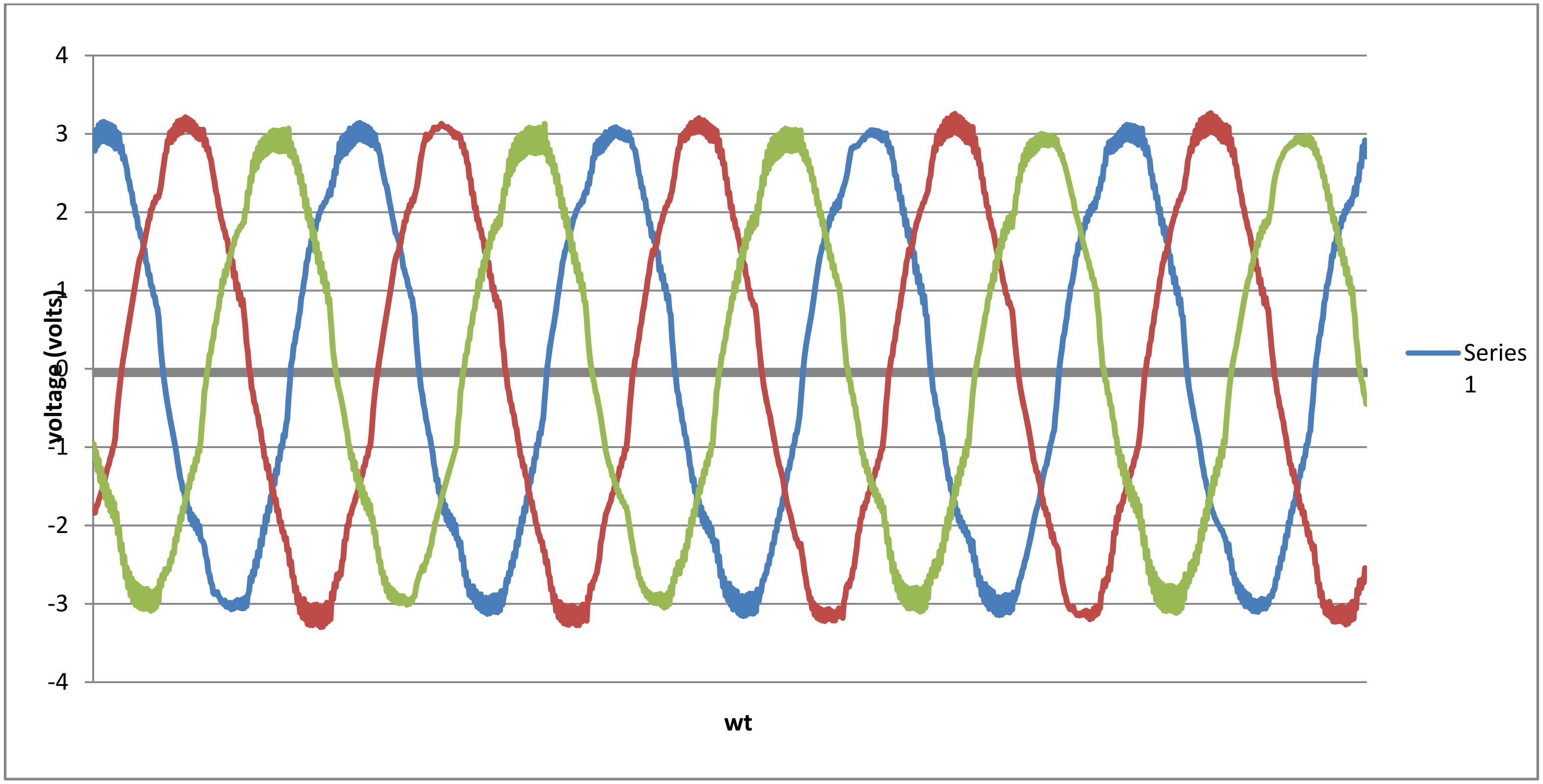


Fig 3.12.2-2 Load voltage waveform for r-y-b phase (with compensation)

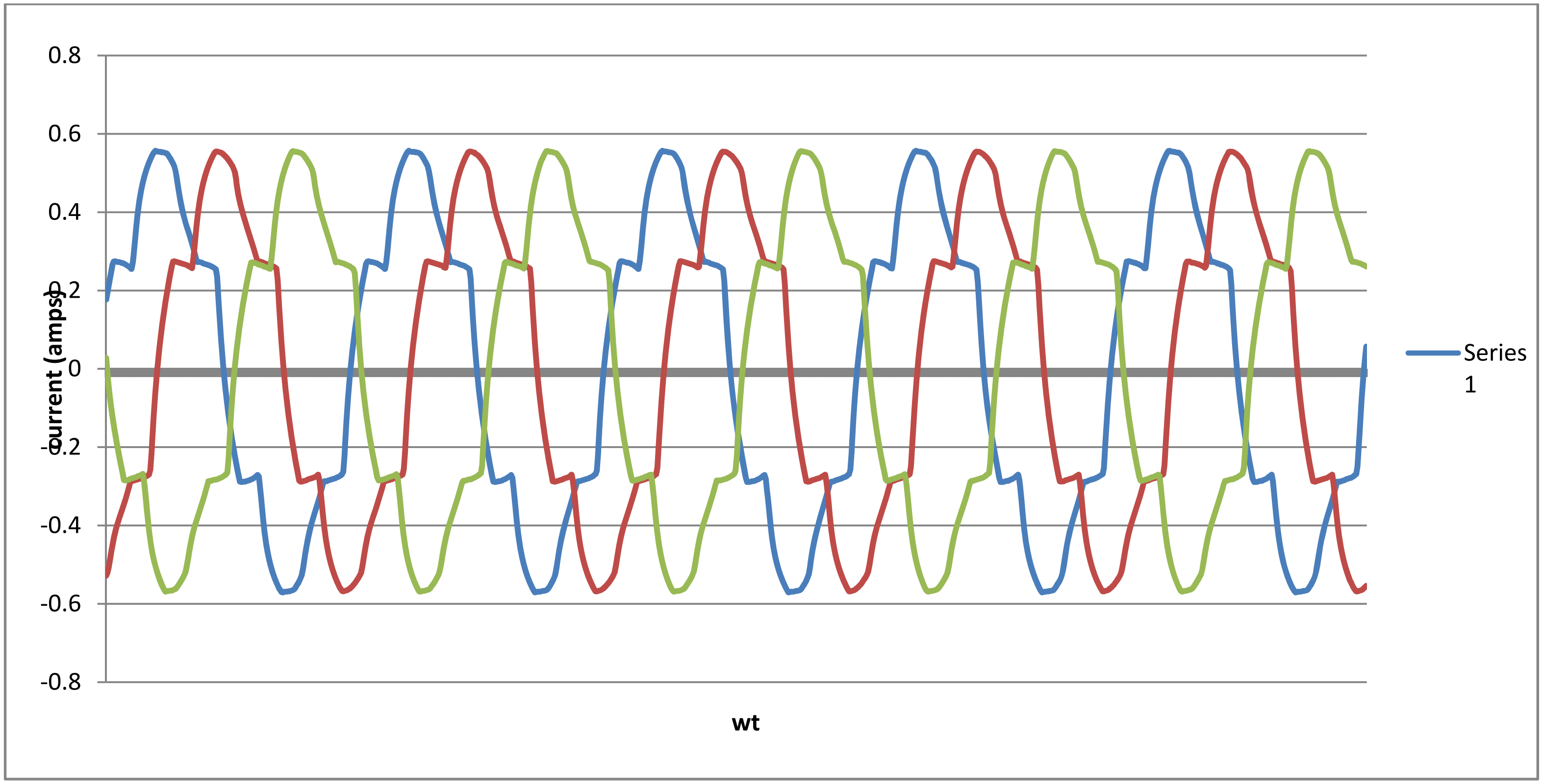


Fig 3.12.2-3 Load current waveform for r-y-b-phase (without compensation)

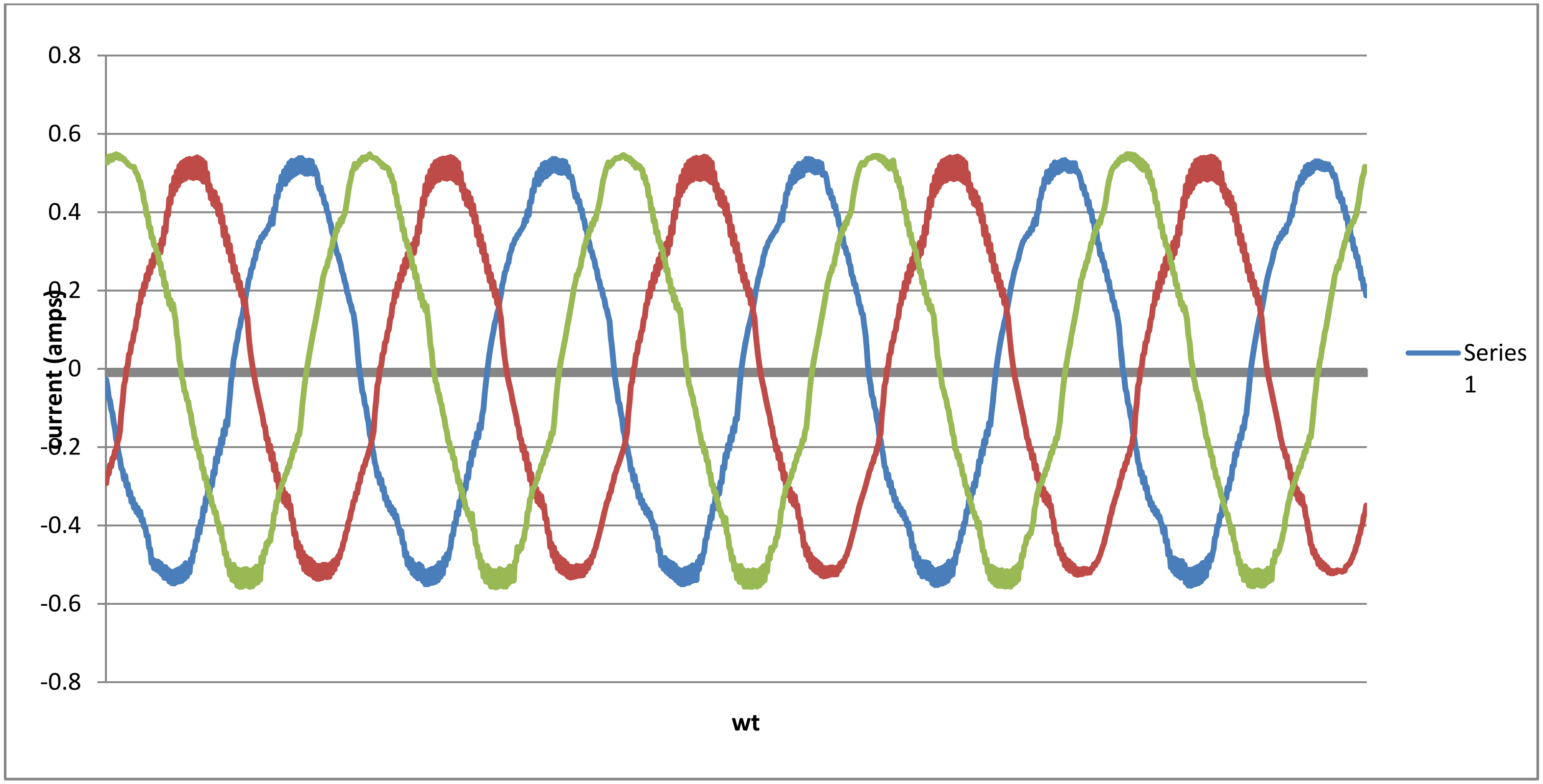


Fig 3.12.2-4 Load current waveform for r-y-b-phase (with compensation)

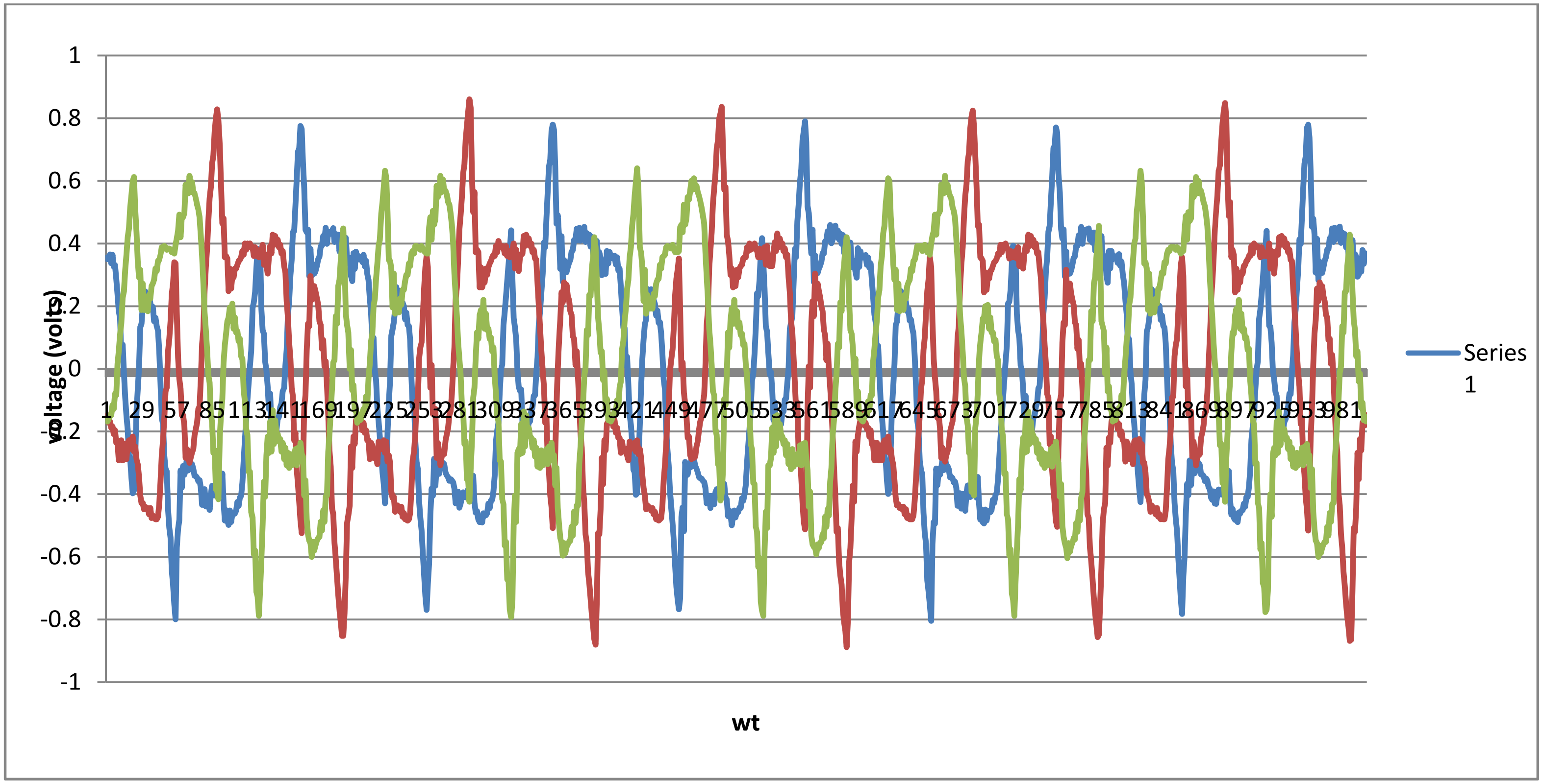


Fig 3.12.2-5 Transformer output voltage waveform for r-y-b phase (with fundamental added)

3.12.3 ANALYSIS OF RESULTS

Table 3.12.3-1: Harmonic analysis of the load voltages with series active filter Bypass

Harm no.	Voltage magnitude VRN Volts	Voltage VRN %	Voltage magnitude VYN Volts	Voltage VYN %	Voltage magnitude VBN Volts	Voltage VBN %
	208.00	100.00	212.00	100.00	206.00	100.00
2	0.40	0.19	0.40	0.19	0.10	0.05
3	0.80	0.38	0.40	0.19	1.00	0.49
4	0.00	0.00	0.00	0.00	0.00	0.00
5	30.00	14.42	30.00	14.15	29.60	14.37
6	0.00	0.00	0.00	0.00	0.00	0.00
7	16.00	7.69	16.00	7.55	15.70	7.62
8	0.00	0.00	0.00	0.00	0.00	0.00
9	8.00	3.85	0.80	0.38	0.30	0.15
10	0.00	0.00	0.00	0.00	0.00	0.00
11	4.80	2.31	5.20	2.45	5.20	2.52
12	0.00	0.00	0.00	0.00	0.00	0.00
13	3.60	1.73	3.60	1.70	3.60	1.75
14	0.00	0.00	0.00	0.00	0.00	0.00
15	0.40	0.19	0.00	0.00	0.10	0.05
16	0.00	0.00	0.00	0.00	0.00	0.00
17	3.20	1.54	3.20	1.51	3.30	1.60
18	0.00	0.00	0.00	0.00	0.00	0.00
19	2.80	1.35	2.80	1.32	2.60	1.26
20	0.00	0.00	0.00	0.00	0.00	0.00
21	0.00	0.00	0.00	0.00	0.10	0.05
22	0.00	0.00	0.00	0.00	0.00	0.00
23	0.80	0.38	0.80	0.38	1.00	0.49
24	0.00	0.00	0.00	0.00	0.00	0.00
25	0.80	0.38	0.80	0.38	0.90	0.44
VOLT/THD	211.05	17.18	214.85	16.45	208.85	16.70

Table 3.12.3-2 : Harmonic analysis of load voltages with series active filter in operation

Harm no.	Voltage magnitude VRN Volts	Voltage VRN %	Voltage magnitude VYN Volts	Voltage VYN %	Voltage magnitude VBN Volts	Voltage VBN %
1	221.60	100.00	208.80	100.00	217.30	100.00
2	3.20	1.44	2.00	0.96	2.20	1.01
3	2.80	1.26	5.20	2.49	9.60	4.42
4	1.60	0.72	1.20	0.57	0.90	0.41
5	4.40	1.99	6.40	3.07	6.40	2.95
6	1.20	0.54	1.20	0.57	0.50	0.23
7	6.00	2.71	5.20	2.49	4.90	2.25
8	0.40	0.18	0.80	0.38	0.70	0.32
9	2.40	1.08	0.80	0.38	1.70	0.78
10	0.80	0.36	0.80	1.34	0.80	0.37
11	2.00	0.90	2.80	0.00	1.90	0.87
12	1.20	0.54	0.00	0.57	0.30	0.14
13	0.80	0.36	1.20	0.38	1.50	0.69
14	0.80	0.36	0.80	0.38	0.50	0.23
15	0.80	0.36	0.80	0.00	1.20	0.55
16	0.00	0.00	0.00	0.77	0.20	0.09
17	1.60	0.72	1.60	0.19	1.20	0.55
18	0.80	0.36	0.40	0.38	0.20	0.09
19	0.80	0.36	0.80	0.19	0.70	0.32
20	0.40	0.18	0.40	0.00	0.00	0.00
21	0.40	0.18	0.00	0.19	0.60	0.28
22	0.40	0.18	0.40	0.19	0.30	0.14
23	0.40	0.18	0.40	0.00	0.20	0.09
24	0.00	0.00	0.00	0.00	0.20	0.09
25	0.40	0.18	0.00	0.00	0.20	0.09
VOLT/THD	221.82	4.42	209.08	5.19	217.71	6.13

Table 3.12.3-3 : Harmonic analysis of the load currents with series active filter Bypass

Harm no.	Current magnitude IRN Amps	Current IRN %	Current magnitude IYN Amps	Current IYN %	Current magnitude IBN Amps	Current IBN %
1	**3.74**	**100.00**	**3.69**	**100.00**	**3.70**	**100.00**
2	**0.02**	**0.53**	**0.01**	**0.27**	**0.03**	**0.81**
3	**0.02**	**0.53**	**0.02**	**0.54**	**0.02**	**0.54**
4	**0.01**	**0.27**	**0.01**	**0.27**	**0.02**	**0.54**
5	**0.50**	**13.37**	**0.52**	**14.09**	**0.49**	**13.24**
6	**0.01**	**0.27**	**0.02**	**0.54**	**0.02**	**0.54**
7	**0.28**	**7.49**	**0.26**	**7.05**	**0.24**	**6.49**
8	**0.01**	**0.27**	**0.00**	**0.00**	**0.01**	**0.27**
9	**0.02**	**0.53**	**0.02**	**0.54**	**0.01**	**0.27**
10	**0.01**	**0.27**	**0.00**	**0.00**	**0.00**	**0.00**
11	**0.08**	**2.14**	**0.09**	**2.44**	**0.09**	**2.43**
12	**0.01**	**0.27**	**0.01**	**0.27**	**0.01**	**0.27**
13	**0.06**	**1.60**	**0.06**	**1.63**	**0.05**	**1.35**
14	**0.00**	**0.00**	**0.00**	**0.00**	**0.01**	**0.27**
15	**0.01**	**0.27**	**0.00**	**0.00**	**0.01**	**0.27**
16	**0.01**	**0.27**	**0.00**	**0.00**	**0.01**	**0.27**
17	**0.04**	**1.07**	**0.05**	**1.36**	**0.03**	**0.81**
18	**0.00**	**0.00**	**0.01**	**0.27**	**0.01**	**0.27**
19	**0.04**	**1.07**	**0.03**	**0.81**	**0.02**	**0.54**
20	**0.00**	**0.00**	**0.00**	**0.00**	**0.01**	**0.27**
21	**0.00**	**0.00**	**0.00**	**0.00**	**0.01**	**0.27**
22	**0.00**	**0.00**	**0.00**	**0.00**	**0.00**	**0.00**
23	**0.01**	**0.27**	**0.01**	**0.27**	**0.01**	**0.27**
24	**0.00**	**0.00**	**0.00**	**0.00**	**0.00**	**0.00**
25	**0.00**	**0.00**	**0.01**	**0.27**	**0.01**	**0.27**
VOLT/ THD	**3.74**	**15.67**	**3.74**	**16.14**	**3.74**	**15.12**

Table 3.12.3-4 : Harmonic analysis of the load currents with series active filter in operation

Harm no.	Current magnitude IRN Amps	Current IRN %	Current magnitude IYN Amps	Current IYN %	Current magnitude IBN Amps	Current IBN %
1	3.78	100.00	3.66	100.00	3.80	100.00
2	0.06	1.59	0.02	0.55	0.04	1.05
3	0.17	4.50	0.10	2.73	0.03	0.79
4	0.03	0.79	0.04	1.09	0.03	0.79
5	0.11	2.91	0.11	3.01	0.09	2.37
6	0.03	0.79	0.01	0.27	0.02	0.53
7	0.08	2.12	0.09	2.46	0.10	2.63
8	0.01	0.26	0.01	0.27	0.00	0.00
9	0.03	0.79	0.02	0.55	0.04	1.05
10	0.00	0.00	0.01	0.27	0.01	0.26
11	0.03	0.79	0.05	1.37	0.04	1.05
12	0.01	0.26	0.01	0.27	0.02	0.53
13	0.02	0.53	0.02	0.55	0.02	0.53
14	0.01	0.26	0.01	0.27	0.02	0.53
15	0.01	0.26	0.01	0.27	0.02	0.53
16	0.00	0.00	0.01	0.27	0.00	0.00
17	0.01	0.26	0.02	0.55	0.04	1.05
18	0.00	0.00	0.01	0.27	0.01	0.26
19	0.01	0.26	0.01	0.27	0.01	0.26
20	0.00	0.00	0.00	0.00	0.01	0.26
21	0.00	0.00	0.00	0.00	0.01	0.26
22	0.00	0.00	0.00	0.00	0.01	0.26
23	0.00	0.00	0.00	0.00	0.01	0.26
24	0.00	0.00	0.00	0.00	0.00	0.00
25	0.00	0.00	0.00	0.00	0.01	0.26
VOLT/THD	3.79	6.24	3.67	5.24	3.80	4.49

Table 3.12.3-5 : Harmonic analysis of the output of the inverter

Harm no.	Voltage magnitude VRN Volts	Voltage VRN %	Voltage magnitude VYN Volts	Voltage VYN %	Voltage magnitude VBN Volts	Voltage VBN %
1	28.80	100.00	28.40	100.00	30.30	100.00
2	4.40	15.28	2.40	8.45	2.00	6.60
3	4.00	13.89	4.40	15.49	3.00	9.90
4	2.00	6.94	2.00	7.04	1.50	4.95
5	16.40	56.94	15.60	54.93	13.70	45.21
6	0.80	2.78	0.40	1.41	0.40	1.32
7	13.60	47.22	11.20	39.44	11.60	38.28
8	0.80	2.78	0.80	2.82	0.40	1.32
9	0.40	1.39	1.20	4.23	0.20	0.66
10	0.80	2.78	0.40	1.41	0.50	1.65
11	4.40	15.28	3.60	12.68	3.50	11.55
12	0.00	0.00	0.00	0.00	0.10	0.33
13	2.40	8.33	2.00	7.04	2.50	8.25
14	0.80	2.78	0.80	2.82	0.60	1.98
15	0.40	1.39	0.40	1.41	0.10	0.33
16	0.40	1.39	0.00	0.00	0.40	1.32
17	1.20	4.17	1.20	4.23	1.20	3.96
18	0.00	0.00	0.00	0.00	0.20	0.66
19	1.20	4.17	1.20	4.23	1.30	4.29
20	0.40	1.39	0.40	1.41	0.30	0.99
21	0.40	1.39	0.00	0.00	0.20	0.66
22	0.40	1.39	0.40	1.41	0.20	0.66
23	1.20	4.17	0.80	2.82	1.20	3.96
24	0.00	0.00	0.00	0.00	0.10	0.33
25	0.80	2.78	0.80	2.82	0.80	2.64
VOLT/THD	36.83	79.70	35.06	72.38	35.79	62.84

3.13 CONCLUSION

This whole control system is realized with analog approach. The reason behind for selecting analog approach is to make control faster. Here in this system various high speed analog IC's are used to make system faster. If this system is realized in digital way it takes lot of time to work out calculation, particularly matrix calculation. If load is like arc furnace in which load current is changing number of times in one cycle, it is very difficult to generate corresponding compensation currents for particular harmonic current. For digital approach, the possible realization is made by using either 16 bit or more processor or by using DSP. This can be taken as future extension of this project. This approach is very useful for harmonic elimination in distribution system as now a day most of the home appliance are equipped with modern sophisticated power electronic control which affect the performance of the nearby equipment connected in the same bus. Also same circuit with small modification can be used as a STATCOM, Static Reactive Power Compensator, for reactive power compensation. For high voltage application hybrid active power filter is used for harmonic elimination, which can be also taken as future project.

With this novel approach on line & instantaneous voltage harmonics detection is possible which is simple and easily implementable. The harmonics in the load bus is reduced from 17 % to 6% at rated voltage of 415 volts & load of 10 Amp. Experimental waveform taken at site along with FFT are shown in section 3.13.1

3.13.1 EXPERIMENTAL WAVEFORMS TAKEN AT SITE ALONG WITH FFT

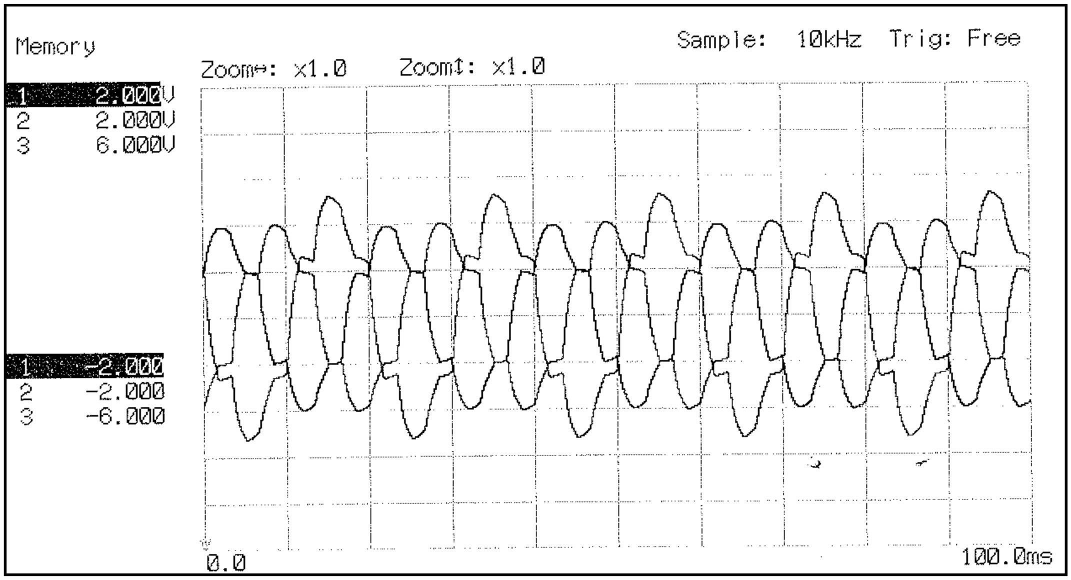


Figure 3.13.1-1 Load voltage waveform for R-Y-B phase (without compensation)

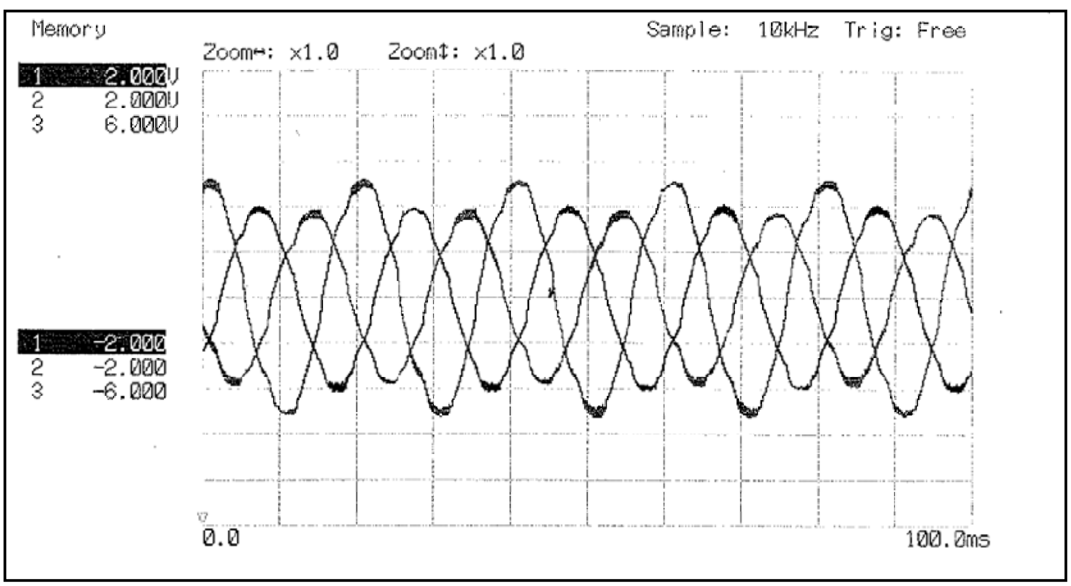
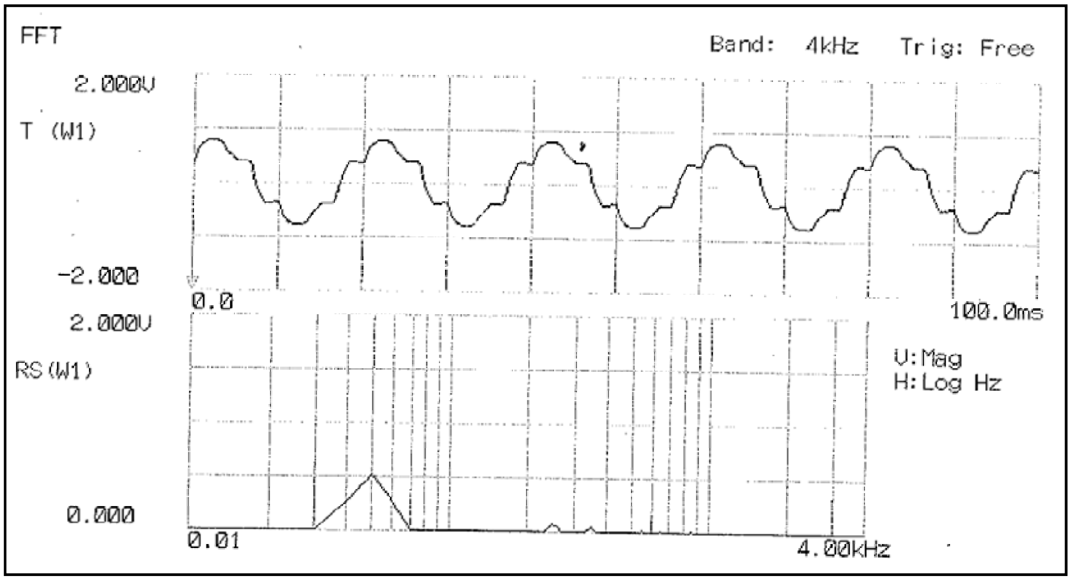
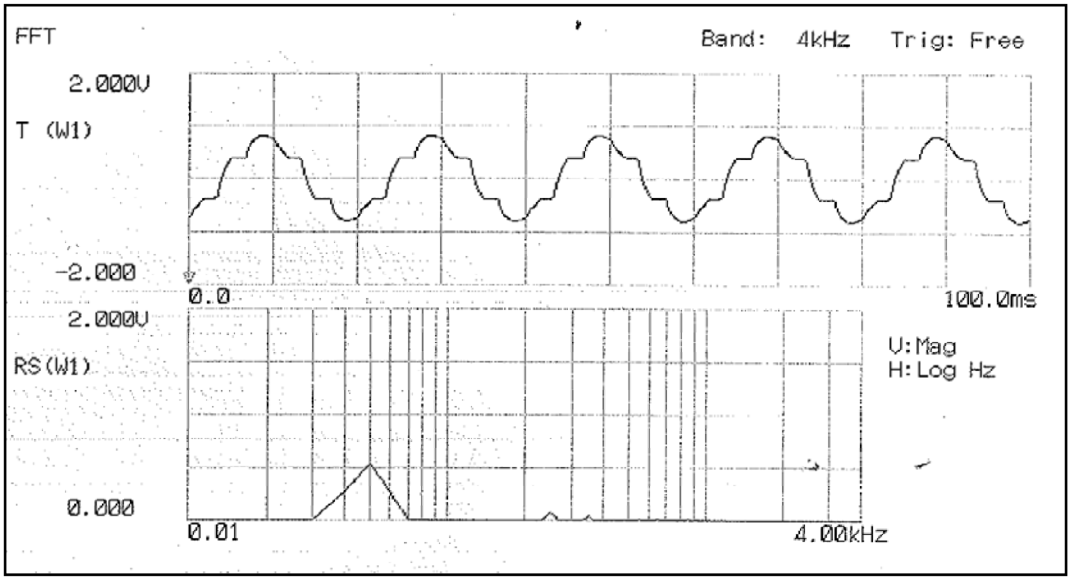


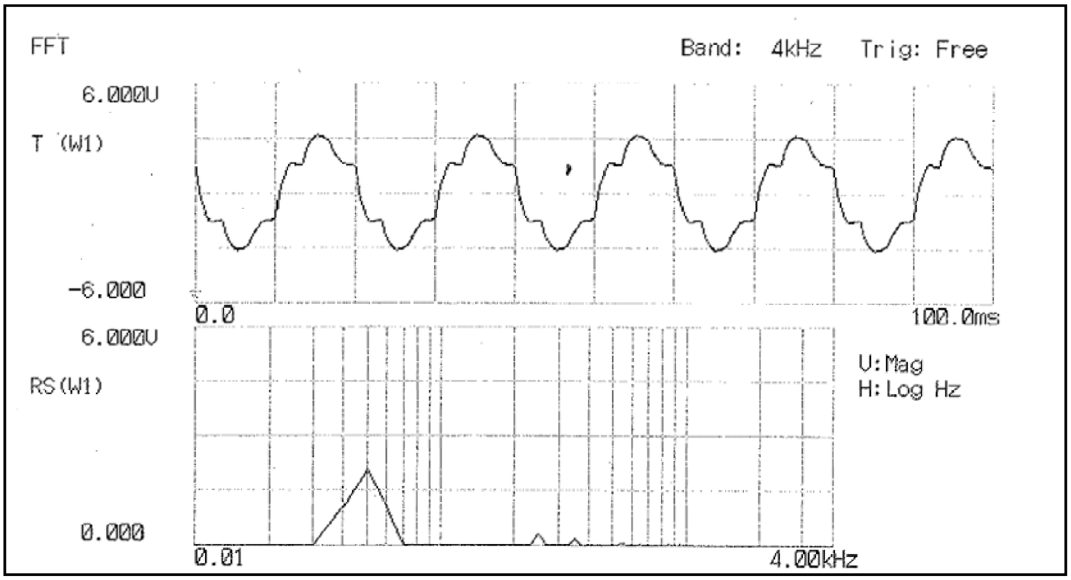
Figure 3.13.1-2 Load voltage waveform for R-Y-B phase (with compensation)



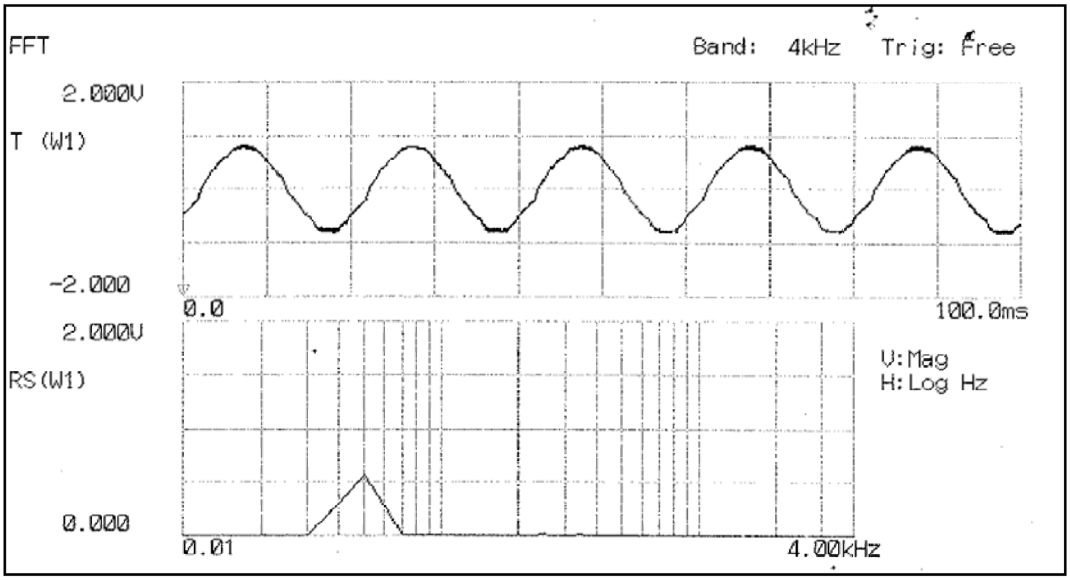
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	208.00
2	0.19	0.40
3	0.38	0.80
4	0.00	0.00
5	14.42	30.00
6	0.00	0.00
7	7.69	16.00
8	0.00	0.00
9	0.38	0.80
10	0.00	0.00
11	2.31	4.80
12	0.00	0.00
13	1.73	3.60
14	0.00	0.00
15	0.19	0.40
16	0.00	0.00
17	1.54	3.20
18	0.00	0.00
19	1.35	2.80
20	0.00	0.00
21	0.00	0.00
22	0.00	0.00
23	0.38	0.80
24	0.00	0.00
25	0.38	0.80
THD %	16.74	
Parameter measured	210.90	



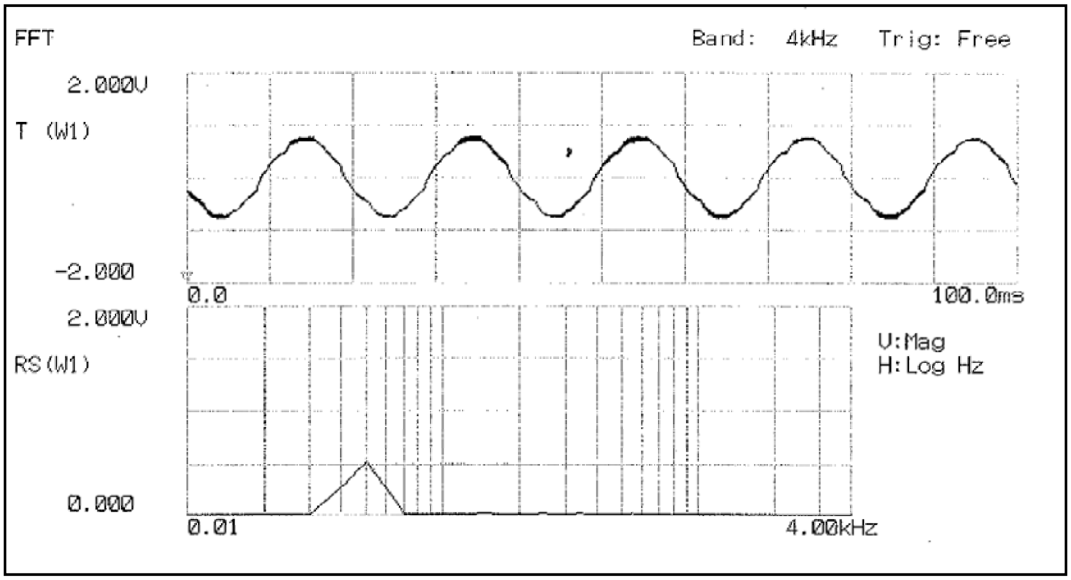
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	212.00
2	0.19	0.40
3	0.19	0.40
4	0.00	0.00
5	14.15	30.00
6	0.00	0.00
7	7.55	16.00
8	0.00	0.00
9	0.38	0.80
10	0.00	0.00
11	2.45	5.20
12	0.00	0.00
13	1.70	3.60
14	0.00	0.00
15	0.00	0.00
16	0.00	0.00
17	1.51	3.20
18	0.00	0.00
19	1.32	2.80
20	0.00	0.00
21	0.00	0.00
22	0.00	0.00
23	0.38	0.80
24	0.00	0.00
25	0.38	0.80
THD %	16.45	
Parameter measured	214.85	



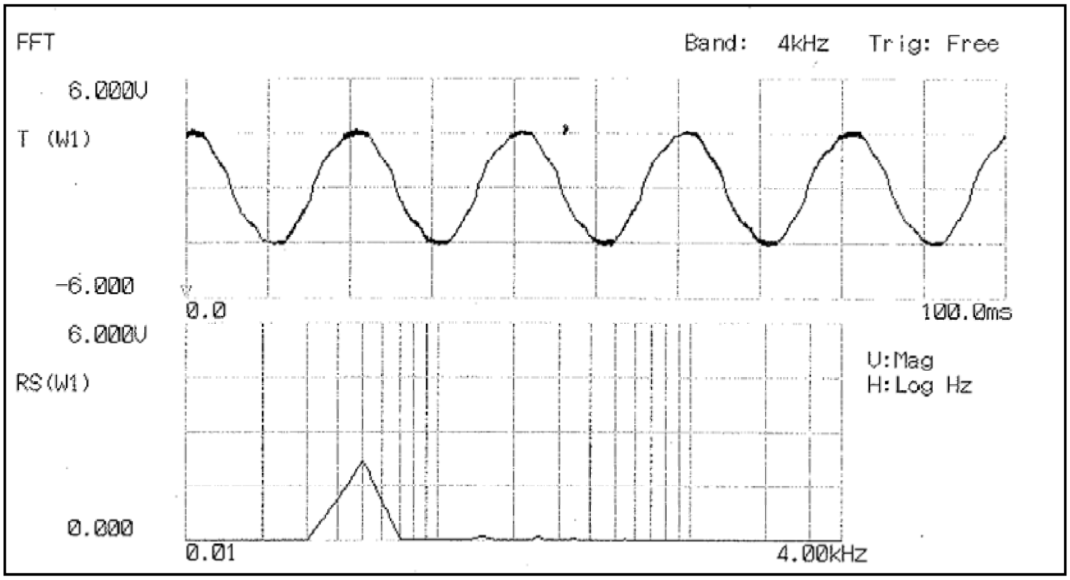
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	206.00
2	0.05	0.10
3	0.49	1.00
4	0.00	0.00
5	14.37	29.60
6	0.00	0.00
7	7.62	15.70
8	0.00	0.00
9	0.15	0.30
10	0.00	0.00
11	2.52	5.20
12	0.00	0.00
13	1.80	3.70
14	0.00	0.00
15	0.05	0.10
16	0.00	0.00
17	1.60	3.30
18	0.00	0.00
19	1.26	2.60
20	0.00	0.00
21	0.05	0.10
22	0.00	0.00
23	0.49	1.00
24	0.00	0.00
25	0.44	0.90
THD %	16.70	
Parameter measured	208.85	



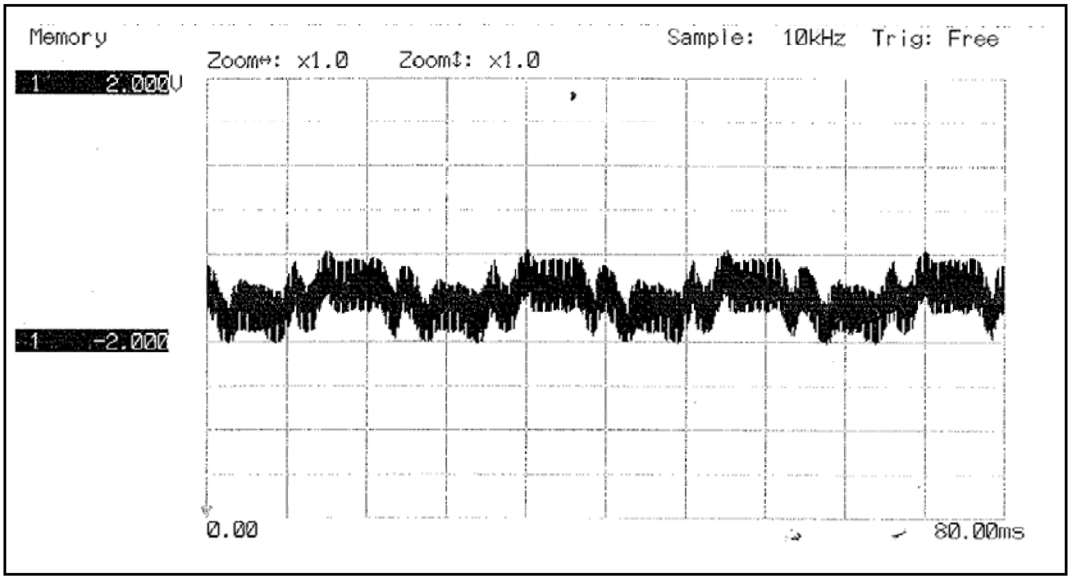
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	221.60
2	0.90	2.00
3	0.90	2.00
4	0.90	2.00
5	1.99	4.40
6	0.72	1.60
7	2.89	6.40
8	0.18	0.40
9	0.90	2.00
10	0.18	0.40
11	0.90	2.00
12	0.54	1.20
13	0.36	0.80
14	0.36	0.80
15	0.54	1.20
16	0.00	0.00
17	0.90	2.00
18	0.36	0.80
19	0.36	0.80
20	0.18	0.40
21	0.18	0.40
22	0.18	0.40
23	0.18	0.40
24	0.00	0.00
25	0.18	0.40
THD %	4.36	
Parameter measured	221.81	



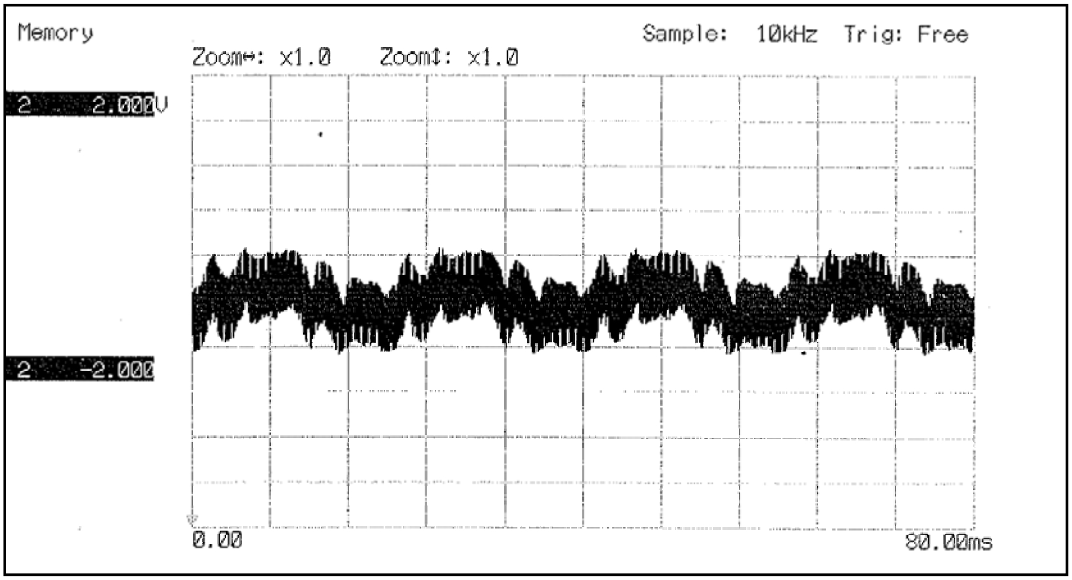
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	208.40
2	0.38	0.80
3	2.69	5.60
4	0.96	2.00
5	3.07	6.40
6	0.38	0.80
7	2.50	5.20
8	0.19	0.40
9	0.58	1.20
10	0.38	0.80
11	1.54	3.20
12	0.19	0.40
13	0.58	1.20
14	0.38	0.80
15	0.38	0.80
16	0.19	0.40
17	0.96	2.00
18	1.92	4.00
19	0.58	1.20
20	0.19	0.40
21	0.00	0.00
22	0.19	0.40
23	0.19	0.40
24	0.00	0.00
25	0.19	0.40
THD %	5.72	
Parameter measured	208.74	



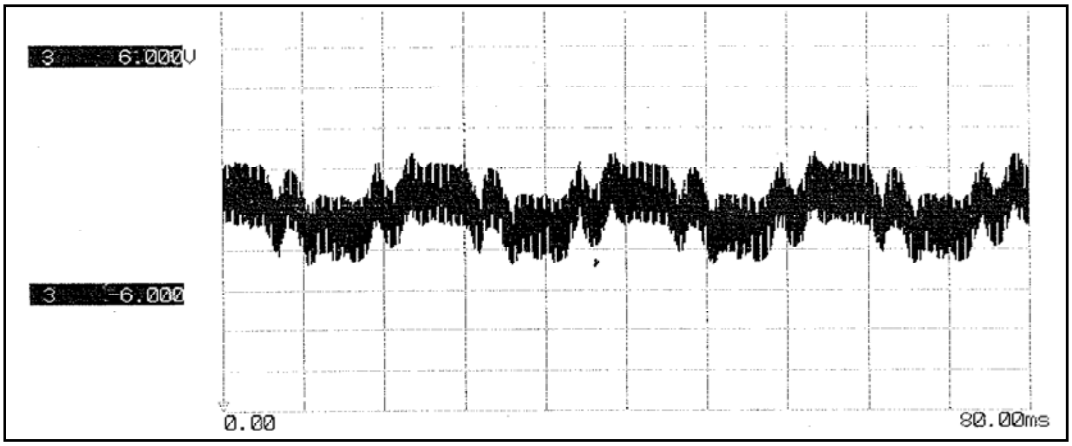
Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	217.90
2	0.83	1.80
3	4.54	9.90
4	0.32	0.70
5	2.89	6.30
6	0.18	0.40
7	2.34	5.10
8	0.37	0.80
9	0.73	1.60
10	0.37	0.80
11	0.96	2.10
12	0.09	0.20
13	0.78	1.70
14	0.23	0.50
15	0.64	1.40
16	0.05	0.10
17	0.64	1.40
18	0.09	0.20
19	0.46	1.00
20	0.00	0.00
21	0.37	0.80
22	0.14	0.30
23	0.14	0.30
24	0.09	0.20
25	0.09	0.20
THD %	6.24	
Parameter measured	218.32	



Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	2.00
2	20.00	0.40
3	40.00	0.80
4	0.00	0.00
5	0.00	0.00
6	20.00	0.40
7	20.00	0.40
8	40.00	0.80
9	80.00	1.60
10	1780.00	35.60
11	40.00	0.80
12	0.00	0.00
13	0.00	0.00
14	0.00	0.00
15	0.00	0.00
16	0.00	0.00
17	0.00	0.00
18	0.00	0.00
19	80.00	1.60
20	400.00	8.00
21	20.00	0.40
22	20.00	0.40
23	0.00	0.00
24	0.00	0.00
25	0.00	0.00
THD %	1829.75	
Parameter measured	36.65	



Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	4.40
2	36.36	1.60
3	9.09	0.40
4	9.09	0.40
5	9.09	0.40
6	9.09	0.40
7	9.09	0.40
8	18.18	0.80
9	54.55	2.40
10	1227.27	54.00
11	9.09	0.40
12	0.00	0.00
13	0.00	0.00
14	0.00	0.00
15	0.00	0.00
16	0.00	0.00
17	0.00	0.00
18	9.09	0.40
19	45.45	2.00
20	200.00	8.80
21	27.27	1.20
22	9.09	0.40
23	0.00	0.00
24	0.00	0.00
25	0.00	0.00
THD %	1246.71	
Parameter measured	55.03	



Harm No.	Voltage V in %	Voltage V In Volts
1	100.00	4.30
2	34.88	1.50
3	6.98	0.30
4	4.65	0.20
5	13.95	0.60
6	9.30	0.40
7	11.63	0.50
8	11.63	0.50
9	39.53	1.70
10	746.51	32.10
11	6.98	0.30
12	4.65	0.20
13	2.33	0.10
14	4.65	0.20
15	0.00	0.00
16	4.65	0.20
17	0.00	0.00
18	2.33	0.10
19	48.84	2.10
20	348.84	15.00
21	9.30	0.40
22	6.98	0.30
23	0.00	0.00
24	2.33	0.10
25	2.33	0.10
THD %	827.66	
Parameter measured	35.85	

CHAPTER-IV

NEGATIVE SEQUENCE & ZERO SEQUENCE COMPENSATOR

TABLE OF CONTENTS

4.1 INTRODUCTION

This chapter presents a novel and online method, which is also a novel contribution of this research work for detecting negative & zero sequence voltages and its compensation. It is based on Instantaneous Active Reactive Power (IARP) theory. Using this concept, negative sequence and zero sequence components are computed online. The α-β transformation is real unlike the complex transformation matrix in case of symmetrical components. So implementation of α-β analysis on line is simple. This theory is verified through simulation and experimentation results.

This chapter describes the effect of negative sequence & zero sequence on electric distribution systems mainly induction motor & energy saved due to compensation.

Modern power systems are three-phase systems that can be balanced or unbalanced and will have mutual coupling between the phases. In many instances, the analysis of these systems is performed using what is known as ''per-phase analysis.'' In this chapter, more generally applicable approach to system analysis known as ''symmetrical components" will be described. The concept of symmetrical components was first proposed for power system analysis by C.L. Fortescue in a classic paper devoted to consideration of the general N-phase case (1918). The effect of negative & zero sequence components can be reduced using negative & zero sequence compensator. New method was introduced to detect negative sequence along with zero sequence components. This method detects the negative sequence components instantaneously and on line. After detecting the negative & zero sequence these signal are converted into three reference signals which are given to voltage source inverter to nullify the effect of negative and zero sequence voltage components in the distribution network.

4.2 GENERATION OF NEGATIVE SEQUENCE & ZERO SEQUENCE

4.2.1 VOLTAGE IMBALANCE

Voltage imbalance (also called voltage unbalance) is sometimes defined as the maximum deviation from the average of the three-phase voltages or currents, divided by the average of the three-phase voltages or currents, expressed in percent. Imbalance is more rigorously defined in the standards [A17]-[A20] using symmetrical components. The ratio of either the negative- or zero sequence components to the positive-sequence component can be used to specify the percent unbalance. The most recent standards11 specify that the negative-sequence method be used. Figure 4.2.1-1 shows an example of negative sequence component to the positive-sequence component in current for a 24 hours trend of imbalance on an industrial feeder.

The primary source of voltage unbalances is single-phase loads on a three-phase circuit. Voltage unbalance can also be the result of blown fuses in one phase of a three-phase capacitor bank. Severe voltage unbalance (greater than 5 percent) can result from single-phasing conditions.

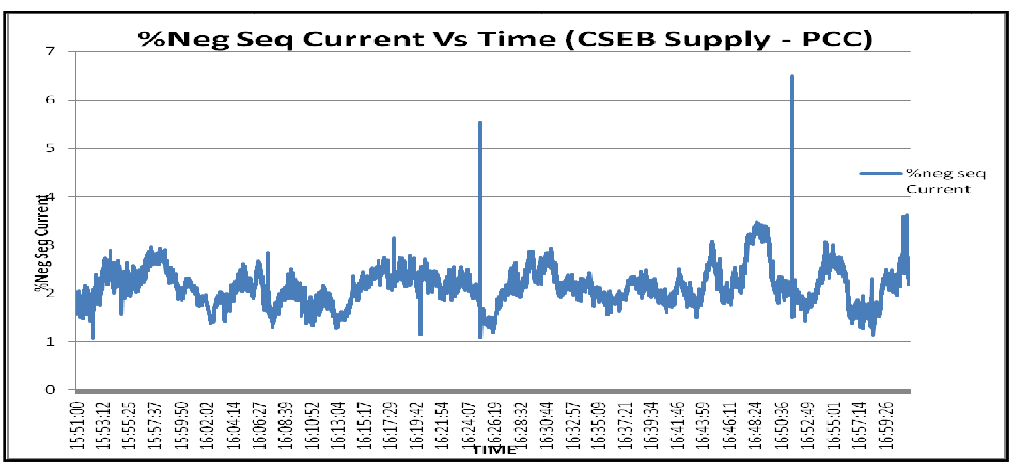


Figure 4.2.1-1: % Negative Sequence Current for an industrial feeder

4.2.2 HARMONIC PHASE SEQUENCES

Power engineers have traditionally used symmetrical components to help describe three-phase system behavior. The three-phase system is transformed into three single-phase systems that are much simpler to analyze. The method of symmetrical components can be employed for analysis of the system's response to harmonic currents provided care is taken not to violate the fundamental assumptions of the method. The method allows any unbalanced set of phase currents (or voltages) to be transformed into three balanced sets. The *positive-sequence* set contains three sinusoids displaced 120° from each other, with the normal A-B-C phase rotation (e.g., 0°, -120°, 120°). The sinusoids of the *negative-sequence* set are also displaced 120°, but have opposite phase rotation (A-C-B, e.g., 0°, 120°, -120°). The sinusoids of the *zero sequence* are in phase with each other (e.g., 0°, 0°, 0°).

In a perfect balanced three-phase system, the harmonic phase sequence can be determined by multiplying the harmonic number h with the normal positive-sequence phase rotation. For example, for the second harmonic, h=2, hence 2 is multiplied with normal phase rotation i.e. 2 x (0, -120°, 120°) or (0°, 120°, -120°), which is the negative sequence. For the third harmonic, h = 3, hence 3 is multiplied with normal phase rotation i.e. 3 x (0°, -120°, 120°) or (0°, 0°, 0°), which is the zero sequence. Phase sequences for all other harmonic orders can be determined in the same fashion. Since a symmetrical distorted waveform in power systems contains only odd-harmonic components, only odd-harmonic phase sequence rotations are summarized here:

■ Harmonics of order h = 1, 7, 13, are generally positive sequence.

■ Harmonics of order h = 5, 11, 17, are generally negative sequence.

■ Triplens (h = 3, 9, 15,…) are generally zero sequence. Impacts of sequence harmonics on various power system components are detailed in section 1.9

4.2.3 TRIPLEN HARMONICS

As previously mentioned, triplen harmonics are the odd multiples of the third harmonic (h = 3, 9, 15, 21,…). They deserve special consideration because the system response is often considerably different for triplens than for the rest of the harmonics. Triplens become an important issue for grounded-wye systems with current flowing on the neutral. Two typical problems are overloading the neutral and telephone interference. One also hears occasionally of devices that mis-operate because the line-to-neutral voltage is badly distorted by the triplen harmonic voltage drop in the neutral conductor. For the system with perfectly balanced single-phase loads illustrated in Figure 4.2.3-1, an assumption is made that fundamental and third-harmonic components are present. Summing the currents at node *N,* the fundamental current components in the neutral are found to be zero, but the third-harmonic components are 3 times those of the phase currents because they naturally coincide in phase and time. Transformer winding connections have a significant impact on the flow of triplen harmonic currents from single-phase nonlinear loads. Two cases are shown in Figure 4.2.3-2. In the wye-delta transformer (top), the triplen harmonic currents are shown entering the wye side. Since they are in phase, they add in the neutral. The delta winding provides ampere-turn balance so that they can flow, but they remain trapped in the delta and do not show up in the line currents on the delta side. When the currents are balanced, the triplen harmonic currents behave exactly as zero-sequence currents, which is precisely what they are. This type of transformer connection is the most common employed in utility distribution substations with the delta winding connected to the transmission feed. Using grounded-wye windings on both sides of the transformer (bottom) allows balanced triplens to flow from the low-voltage system to the high-voltage system unimpeded. They will be present in equal proportion on both sides. Many loads in the United States are served in this fashion. Some important implications of this related to power quality analysis are

1. Transformers, particularly the neutral connections, are susceptible to overheating when serving single-phase loads on the wye side that have high third-harmonic content.

2. Measuring the current on the delta side of a transformer will not show the triplens and, therefore, not give a true idea of the heating the transformer is being subjected to.

3. The flow of triplen harmonic currents can be interrupted by the appropriate isolation transformer connection. Removing the neutral connection in one or both wye windings blocks the flow of triplen harmonic current.

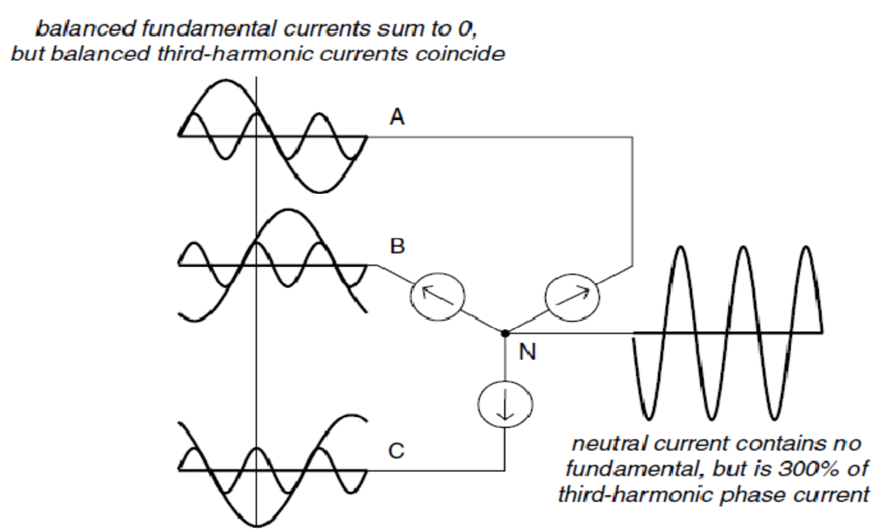


Figure 4.2.3-1: High neutral currents in circuits serving single-phase nonlinear loads.

There is no place for ampere-turn balance. Likewise, a delta winding blocks the flow from the line. One should note that three-legged core transformers behave as if they have a "phantom" delta tertiary winding. Therefore, a wye-wye connection with only one neutral point grounded will still be able to conduct the triplen harmonics from that side. These rules about triplen harmonic current flow in transformers apply only to *balanced* loading conditions. When the phases are not balanced, currents of normal triplen harmonic frequencies may very well show up where they are not expected. The normal mode for triplen harmonics is to be zero sequence. During imbalances, triplen harmonics may have positive or negative sequence components, too. One notable case of this is a three-phase arc furnace. The furnace is nearly always fed by a delta-delta connected transformer to block the flow of the zero sequence currents as shown in Figure 4.2.3-3. Thinking that third harmonics are synonymous with zero sequence, many engineers are surprised to find substantial third-harmonic current present in large magnitudes in the line current.

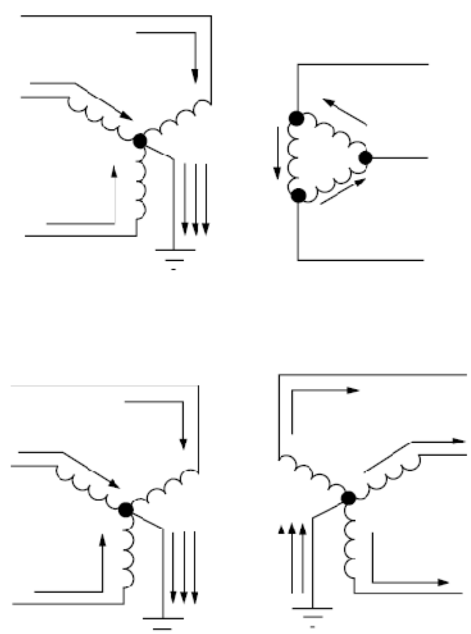


Figure 4.2.3-2: Flow of third-harmonic current in three-phase transformers.

However, during scrap meltdown, the furnace will frequently operate in an unbalanced mode with only two electrodes carrying current. Large third-harmonic currents can then freely circulate in these two phases just as in a single-phase circuit. However, they are not zero-sequence currents. The third-harmonic currents have equal amounts of positive- and negative-sequence currents. But to the extent that the system is *mostly* balanced, triplens mostly behave in the manner described.

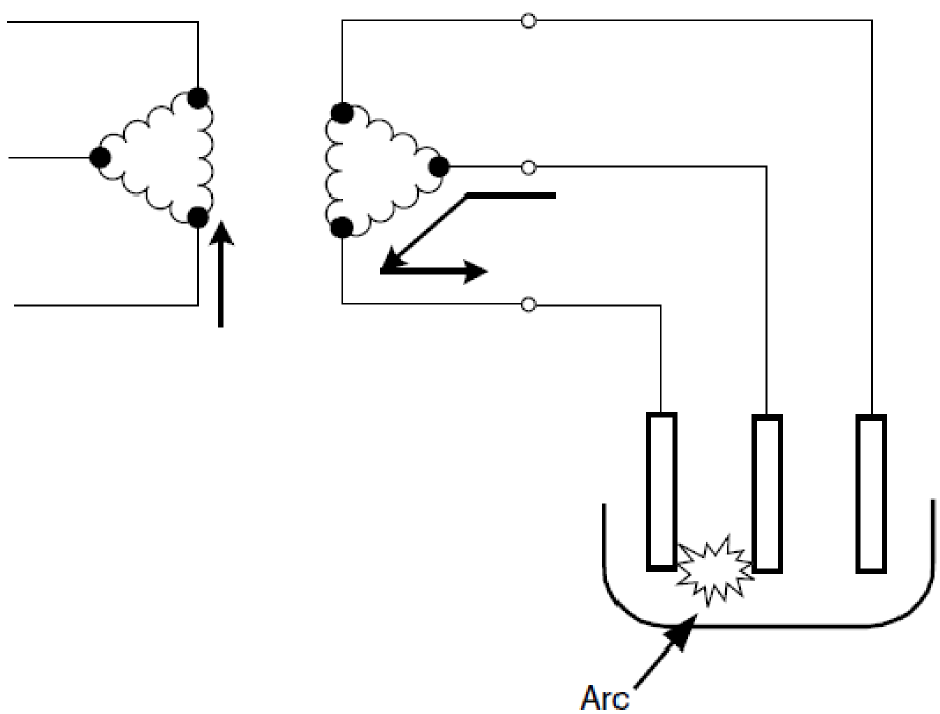


Figure 4.2.3-3: Arc furnace operation in an unbalanced mode allows triplen harmonics to reach the power system despite a delta connected transformer.

4.3 ANALYSIS OF VOLTAGE UNBALANCE PROBLEM

Induction motors have been used in the past mainly in applications requiring a constant speed because conventional methods of their speed control have either been expensive or highly inefficient. Variable speed applications have been dominated by DC drives. Availability of thyristors, power transistors, IGBT and GTO has allowed the development of variable speed induction motor drives. The main drawback of de motors is the presence of commutator and brushes, which require frequent maintenance and make them unsuitable for explosive and dirty environments. On the other hand, induction motors, particularly squirrel-cage are rugged, cheaper, lighter, smaller, more efficient, require lower maintenance and can operate in dirty and explosive environments. Although variable speed induction motor drives are generally expensive than dc drives, they are used in a number of applications such as fans, blowers, mill run-out tables, cranes, conveyers, traction etc. because of the advantages of induction motors. Other dominant applications are underground and underwater installations, and explosive and dirty environments.

Use of 3-ϕ induction motor (IM) in various industries is almost 80% of their total drive requirements. Performance of these induction motors is affected by the unbalance in bus voltage caused by negative sequence currents, which heat up the rotor and generate Braking Torque. Due to the very low value of negative sequence impedance offered by the induction motor as compared to its positive sequence impedance, a very small negative sequence voltage component in the input may give rise to considerable negative sequence current causing significance deterioration in the performance of induction motor.

Therefore even low unbalance factor (ratio of negative sequence to positive sequence) in input voltage has to be carefully looked into. Voltage unbalances occur quite often in distribution systems. Figure 4.3-1 shows the schematic of this conventional induction motor drive and the effect of input voltage unbalance.

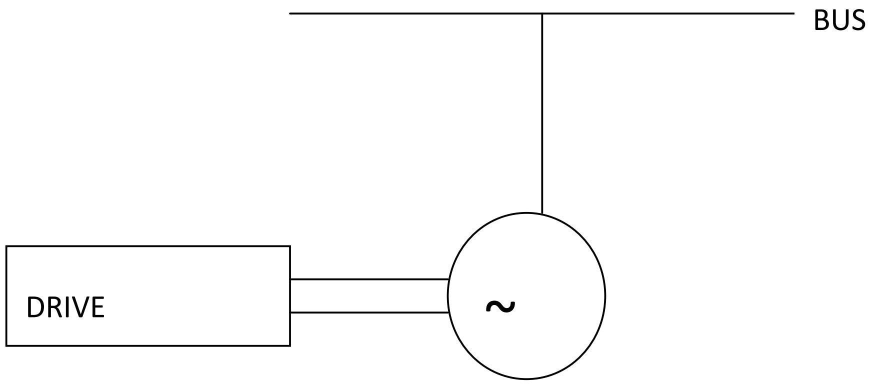


Figure 4.3-1(a): Schematic of conventional induction motor drive

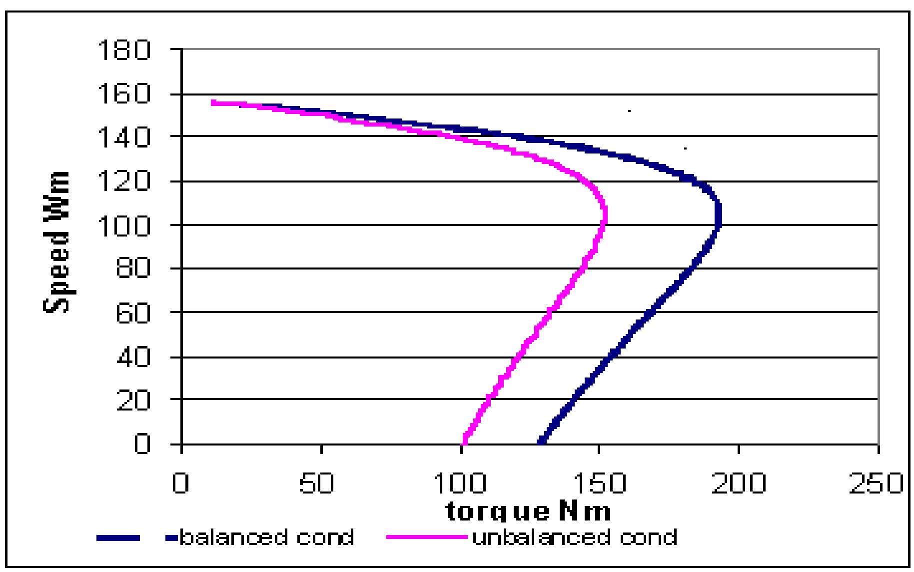


Figure 4.3-1(b): Torque v/s speed characteristic with balanced and unbalanced condition

4.4 SYMMETRICAL COMPONENTS FOR POWER SYSTEMS ANALYSIS

The case for per-phase analysis can be made by considering the simple three-phase system illustrated in Figure 4.4-1. The steady-state circuit response can be obtained by solution of the three loop equations presented in Equation (4.4-1a) through (4.4-1c). By solving these loop equations for the three line currents, Equation (4.4-2a) through (4.4-2c) are obtained. Now, if completely balanced source operation (the impedances are defined to be balanced) is assumed, then the line currents will also form a balanced three-phase set. Hence, their sum, and the neutral current, will be zero. As a result, the line current solutions are as presented in Equation (4.4-3a) through (4.4-3c).

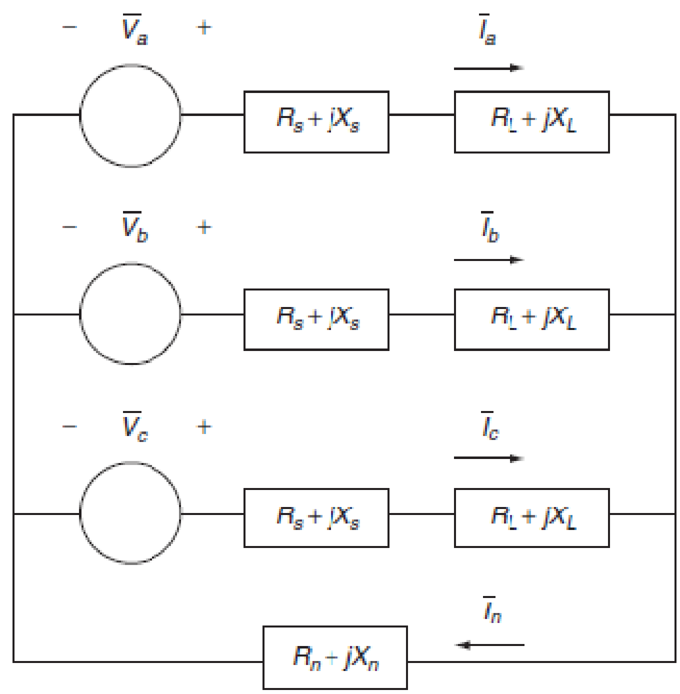


Figure 4.4-1: A simple three-phase system

$$\bar{V}_a - \bar{I}_a\,(R_S + jX_S) - \bar{I}_a\,(R_L + jX_L) - \bar{I}_n\,(R_n + jX_n) = 0 \tag{4.4-1a}$$

$$\bar{V}_b - \bar{I}_b\,(R_S + jX_S) - \bar{I}_b\,(R_L + jX_L) - \bar{I}_n\,(R_n + jX_n) = 0 \tag{4.4-1b}$$

$$\bar{V}_c - \bar{I}_c\,(R_S + jX_S) - \bar{I}_c\,(R_L + jX_L) - \bar{I}_n\,(R_n + jX_n) = 0 \tag{4.4-1c}$$

$$\bar{I}_a = \frac{\bar{V}_a - \bar{I}_n\,(R_n + jX_n)}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-2a}$$

$$\bar{I}_b = \frac{\bar{V}_b - \bar{I}_n\,(R_n + jX_n)}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-2b}$$

$$\bar{I}_c = \frac{\bar{V}_c - \bar{I}_n\,(R_n + jX_n)}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-2c}$$

$$\bar{I}_a = \frac{\bar{V}_a}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-3a}$$

$$\bar{I}_b = \frac{\bar{V}_b}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-3b}$$

$$\bar{I}_c = \frac{\bar{V}_c}{(R_S + R_L) + j(X_S + X_L)} \tag{4.4-3c}$$

The circuit synthesis of Equation (4.4-3a) through (4.4-3c) is illustrated in Figure 4.4-2. Particular notice should be taken of the fact the response of each phase is independent of the other two phases. Thus, only one phase need be solved, and three-phase symmetry may be applied to determine the solutions for the other phases.

This solution technique is the per-phase analysis method. If one considers the introduction of an unbalanced source or mutual coupling between the phases in Figure 4.4-1, then per-phase analysis will not result in three decoupled networks as shown in Figure 4.4-2. In fact, in the general sense, no immediate circuit reduction is available without some form of reference frame transformation. The symmetrical component transformation represents such a transformation, which will enable decoupled analysis in the general case and single-phase analysis in the balanced case.

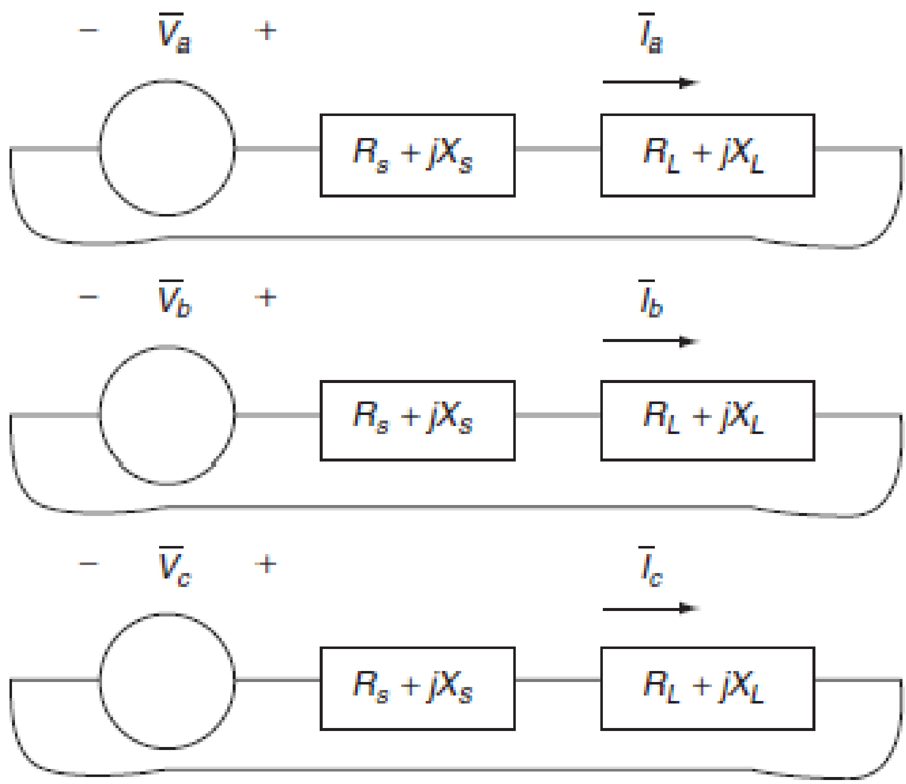


Figure 4.4-2: Decoupled phases of the three phase system

4.4.1 FUNDAMENTAL DEFINITIONS

4.4.1.1 VOLTAGE AND CURRENT TRANSFORMATION

To develop the symmetrical components, [J11]-[J14] let us first consider an arbitrary (no assumptions on balance) three-phase set of voltages as defined in Equation (4.4.1.1-1a) through (4.4.1.1-1c). Note that it could just as easily be considering current for the purposes at hand, but voltage was selected arbitrarily. Each voltage is defined by a magnitude and phase angle. Hence, six degrees of freedom is available to fully define this arbitrary voltage set.

$$\bar{V}_a = V_a \angle \theta_a \tag{4.4.1.1-1a}$$

$$\bar{V}_b = V_b \angle \theta_b \tag{4.4.1.1-1b}$$

$$\bar{V}_c = V_c \angle \theta_c \tag{4.4.1.1-1c}$$

Each of the three given voltages can be represented as the sum of three components as illustrated in Equation (4.4.1.1-2a) through (4.4.1.1-2c). For now, these components are considered to be completely arbitrary except for their sum. The 0, 1 and 2 subscripts are used to denote the zero, positive and negative sequence components of each phase voltage, respectively. Examination of Equation (4.4.1.1-2a-c) reveals that 6 degrees of freedom exist on the left-hand side of the equations while 18 degrees of freedom exist on the right-hand side. Therefore, for the relationship between the voltages in the a-b-c frame of reference and the voltages in the 0-1-2 frame of reference to be unique, the right-hand side of Equation (4.4.1.1-2) must be considered.

$$\bar{V}_a = \bar{V}_{a0} + \bar{V}_{a1} + \bar{V}_{a2} \tag{4.4.1.1-2a}$$

$$\bar{V}_b = \bar{V}_{b0} + \bar{V}_{b1} + \bar{V}_{b2} \tag{4.4.1.1-2b}$$

$$\bar{V}_c = \bar{V}_{c0} + \bar{V}_{c1} + \bar{V}_{c2} \tag{4.4.1.1-2c}$$

It is begin by forcing the a_0, b_0, and c_0 voltages to have equal magnitude and phase. This is defined in Equation (2.6). The zero sequence components of each phase voltage are all defined by a single magnitude and a single phase angle. Hence, the zero sequence components have been reduced from 6 degrees of freedom to 2.

$$\bar{V}_{a0} = \bar{V}_{b0} = \bar{V}_{c0} \equiv \bar{V}_0 = V_0\angle\theta_0 \tag{4.4.1.1-3}$$

Second, the a_1, b_1, and c_1 voltages are forced to form a balanced three-phase set with positive phase sequence. This is mathematically defined in Equation (4.4.1.1-4a-c). This action reduces the degrees of freedom provided by the positive sequence components from 6 to 2.

$$\bar{V}_{a1} = \bar{V}_1 = V_1\angle\theta_1 \tag{4.4.1.1-4a}$$

$$\bar{V}_{b1} = V_1\angle(\theta_1 - 120^0) = \bar{V}_1.1\angle-120^0 \tag{4.4.1.1-4b}$$

$$\bar{V}_{c1} = V_1\angle(\theta_1 + 120^0) = \bar{V}_1.1\angle+120^0 \tag{4.4.1.1-4c}$$

And finally, the a_2, b_2, and c_2 voltages are forced to form a balanced three-phase set with negative phase sequence. This is mathematically defined in Equation (4.4.1.1-5a-c). As in the case of the positive sequence components, the negative sequence components have been reduced from 6 to 2 degrees of freedom.

$$\bar{V}_{a2} = \bar{V}_2 = V_2\angle\theta_2 \tag{4.4.1.1-5a}$$

$$\bar{V}_{b2} = V_2\angle(\theta_2 + 120^0) = \bar{V}_2.1\angle+120^0 \tag{4.4.1.1-5b}$$

$$\bar{V}_{c2} = V_2\angle(\theta_2 - 120^0) = \bar{V}_2.1\angle-120^0 \tag{4.4.1.1-5c}$$

Now, the right and left hand sides of Equation (4.4.1.1-2a) through (4.4.1.1-2c) each have 6 degrees of freedom. Thus, the relationship between the symmetrical component voltages and the original phase voltages is unique. The final relationship is presented in Equation (4.4.1.1-6a) through (4.4.1.1-6c). Note that the constant ''a'' has been defined as indicated in Equation (4.4.1.1-7).

$$\bar{V}_a = \bar{V}_0 + \bar{V}_1 + \bar{V}_2 \qquad (4.4.1.1\text{-}6a)$$

$$\bar{V}_b = \bar{V}_0 + \bar{a}^2\bar{V}_1 + \bar{a}\bar{V}_2 \qquad (4.4.1.1\text{-}6b)$$

$$\bar{V}_c = \bar{V}_0 + \bar{a}\bar{V}_1 + \bar{a}^2\bar{V}_2 \qquad (4.4.1.1\text{-}6c)$$

$$\bar{a} = 1\angle 120^0 \qquad (4.4.1.1\text{-}7)$$

Equation (4.4.1.1-6) is more easily written in matrix form, as indicated in Equation (4.4.1.1-8) in both expanded and compact form. In Equation (4.4.1.1-8), the [T] matrix is constant, and the inverse exists. Thus, the inverse transformation can be defined as indicated in Equation (4.4.1.1-9). The over tilde (˜) indicates a vector of complex numbers.

$$\begin{bmatrix} \bar{V}_a \\ \bar{V}_b \\ \bar{V}_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a}^2 & \bar{a} \\ 1 & \bar{a} & \bar{a}^2 \end{bmatrix} \begin{bmatrix} \bar{V}_0 \\ \bar{V}_1 \\ \bar{V}_2 \end{bmatrix}$$

$$\tilde{V}_{abc} = [\bar{T}]\,\tilde{V}_{012} \qquad (4.4.1.1\text{-}8)$$

$$\begin{bmatrix} \bar{V}_0 \\ \bar{V}_1 \\ \bar{V}_2 \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a} & \bar{a}^2 \\ 1 & \bar{a}^2 & \bar{a} \end{bmatrix} \begin{bmatrix} \bar{V}_a \\ \bar{V}_b \\ \bar{V}_c \end{bmatrix}$$

$$\tilde{V}_{012} = [\bar{T}]^{-1}\, \tilde{V}_{abc} \qquad (4.4.1.1\text{-}9)$$

Equations (4.4.1.1-10) and (4.4.1.1-11) define an identical transformation and inverse transformation for current.

$$\begin{bmatrix} \bar{I}_a \\ \bar{I}_b \\ \bar{I}_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a}^2 & \bar{a} \\ 1 & \bar{a} & \bar{a}^2 \end{bmatrix} \begin{bmatrix} \bar{I}_0 \\ \bar{I}_1 \\ \bar{I}_2 \end{bmatrix}$$

$$\tilde{I}_{abc} = [\bar{T}]\, \tilde{I}_{012} \qquad (4.4.1.1\text{-}10)$$

$$\begin{bmatrix} \bar{I}_0 \\ \bar{I}_1 \\ \bar{I}_2 \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a} & \bar{a}^2 \\ 1 & \bar{a}^2 & \bar{a} \end{bmatrix} \begin{bmatrix} \bar{I}_a \\ \bar{I}_b \\ \bar{I}_c \end{bmatrix}$$

$$\tilde{I}_{012} = [\bar{T}]^{-1}\, \tilde{I}_{abc} \qquad (4.4.1.1\text{-}11)$$

4.4.1.2 IMPEDANCE TRANSFORMATION

In order to assess the impact of the symmetrical component transformation on systems impedances, Figure 4.4.1.2-1 was considered. Note that the balanced case has been assumed. Kirchhoff's Voltage Law for the circuit dictates equations Equation (4.4.1.2-1a -c), which are written in matrix form in Equation (4.4.1.2-2) and even more simply in Equation (4.4.1.2-3).

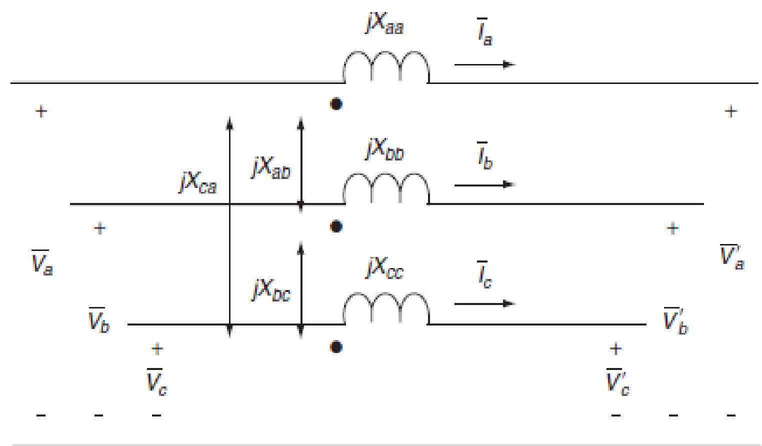


Figure 4.4.1.2-1: Mutually Coupled Series Impedances

$$\bar{V}_a - \bar{V}_a' = jX_{aa}\bar{I}_a + jX_{ab}\bar{I}_b + jX_{ca}\bar{I}_c \quad (4.4.1.2\text{-}1a)$$

$$\overline{V_b} - \bar{V}_b' = jX_{ab}\bar{I}_a + jX_{bb}\bar{I}_b + jX_{bc}\bar{I}_c \quad (4.4.1.2\text{-}1b)$$

$$\bar{V}_c - \bar{V}_c' = jX_{ca}\bar{I}_a + jX_{bc}\bar{I}_b + jX_{cc}\bar{I}_c \quad (4.4.1.2\text{-}1c)$$

$$\begin{bmatrix} \bar{V}_a \\ \bar{V}_b \\ \bar{V}_c \end{bmatrix} - \begin{bmatrix} \bar{V}_a' \\ \bar{V}_b' \\ \bar{V}_c' \end{bmatrix} = j \begin{bmatrix} X_{aa} & X_{ab} & X_{ca} \\ X_{ab} & X_{bb} & X_{bc} \\ X_{ca} & X_{bc} & X_{cc} \end{bmatrix} \begin{bmatrix} \bar{I}_a \\ \bar{I}_b \\ \bar{I}_c \end{bmatrix} \quad (4.4.1.2\text{-}2)$$

$$\tilde{V}_{abc} - \tilde{V}_{abc}' = [\bar{Z}_{abc}]\,\bar{I}_{abc} \quad (4.4.1.2\text{-}3)$$

Multiplying both sides of Equation (4.4.1.2-3) by [T]$^{-1}$ yields Equation (4.4.1.2-4). Then, substituting Equation (4.4.1.1-9) and (4.4.1.1-10) into the result leads to the sequence equation presented in Equation (4.4.1.2-5). The equation is written strictly in the 012 frame reference in Equation (4.4.1.2-6) where the sequence impedance matrix is defined in Equation (4.4.1.2-7).

$$[\bar{T}]^{-1}\tilde{V}_{abc} - [\bar{T}]^{-1}\tilde{V}'_{abc} = [\bar{T}]^{-1}[\bar{Z}_{abc}]\,\tilde{I}_{abc} \qquad (4.4.1.2\text{-}4)$$

$$\tilde{V}_{012} - \tilde{V}'_{012} = [\bar{T}]^{-1}[\bar{Z}_{abc}]\,[\bar{T}]\,\tilde{I}_{012} \qquad (4.4.1.2\text{-}5)$$

$$\tilde{V}_{012} - \tilde{V}'_{012} = [\bar{Z}_{012}]\,\tilde{I}_{012} \qquad (4.4.1.2\text{-}6)$$

$$[\bar{Z}_{012}] = [T]^{-1}\,[\bar{Z}_{abc}]\,[\bar{T}] = \begin{bmatrix} \bar{Z}_{00} & \bar{Z}_{01} & \bar{Z}_{02} \\ \bar{Z}_{10} & \bar{Z}_{11} & \bar{Z}_{12} \\ \bar{Z}_{20} & \bar{Z}_{21} & \bar{Z}_{22} \end{bmatrix} \qquad (4.4.1.2\text{-}7)$$

4.4.1.3 POWER CALCULATIONS

The impact of the symmetrical components on the computation of complex power can be easily derived from the basic definition. Consider the source illustrated in Figure 4.4.1.3-1. The three-phase complex power supplied by the source is defined in Equation (4.4.1.3-1). The algebraic manipulation to Equation (4.4.1.3-1) is presented, and the result in the sequence domain is presented in Equation (4.4.1.3-2) in matrix form and in Equation (4.4.1.3-3) in scalar form.

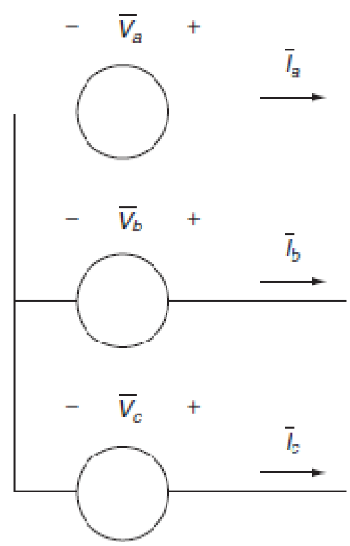


Figure 4.4.1.3-1: Three-Phase Wye Connected Source

$$S_{3\emptyset} = \bar{V}_a\,\bar{I}_a^* + \bar{V}_b\,\bar{I}_b^* + \bar{V}_c\,\bar{I}_c^* = \tilde{V}_{abc}^T\,\tilde{I}_{abc}^* \qquad (4.4.1.3\text{-}1)$$

$$S_{3\emptyset} = \tilde{V}_{abc}^T\,\tilde{I}_{abc}^* = \{[\bar{T}]\,\tilde{V}_{012}\}^T\,\{[\bar{T}]\,\tilde{I}_{012}\}^*$$

$$= \tilde{V}_{012}^T\,[\bar{T}]^T\,[\bar{T}]^*\,\tilde{I}_{012}^*$$

$$[\bar{T}]^T\,[\bar{T}]^* = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a}^2 & \bar{a} \\ 1 & \bar{a} & \bar{a}^2 \end{bmatrix} \begin{bmatrix} 1 & 1 & 1 \\ 1 & \bar{a} & \bar{a}^2 \\ 1 & \bar{a}^2 & \bar{a} \end{bmatrix}$$

$$= \begin{bmatrix} 3 & 0 & 0 \\ 0 & 3 & 0 \\ 0 & 0 & 3 \end{bmatrix} = 3\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

$$\bar{S}_{3\emptyset} = 3\, \tilde{V}_{012}^{T}\, \tilde{I}_{012}^{*} \tag{4.4.1.3-2}$$

$$\bar{S}_{3\emptyset} = 3\, \{\bar{V}_0\, \bar{I}_0^{*} + \bar{V}_1\, \bar{I}_1^{*} + \bar{V}_2\, \bar{I}_2^{*}\} \tag{4.4.1.3-3}$$

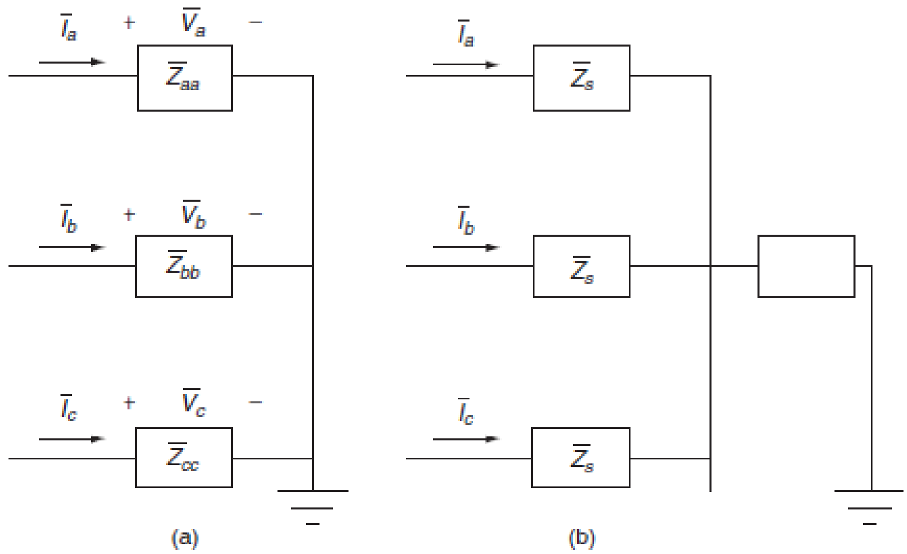


Figure 4.4.1.3-2: Three-Phase Impedance Load Model

Note that the nature of the symmetrical component transformation is not one of power invariance, as indicated by the multiplicative factor of 3 in Equation (4.4.1.3-3). However, this will prove useful in the analysis of balanced systems, which will be seen later. Power invariant transformations do exist as minor variations of the one defined herein. However, they are not typically employed, although the results are just as mathematically sound.

4.4.2 SUMMARY OF THE SYMMETRICAL COMPONENTS IN THE GENERAL THREE-PHASE CASE

The general symmetrical component transformation process has been defined in this section. Table 4.4.2-1 is a short form reference for the utilization of these procedures in the general case (i.e., no assumption of balanced conditions).

Table 4.4.2-1 Summary of the Symmetrical Components in the General Case

Quantity	Transformation Equations: abc ⇒ 012	Transformation Equations: 012 ⇒ abc
Voltage	$\begin{bmatrix} \overline{V}_0 \\ \overline{V}_1 \\ \overline{V}_2 \end{bmatrix} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & \overline{a} & \overline{a}^2 \\ 1 & \overline{a}^2 & \overline{a} \end{bmatrix}\begin{bmatrix} \overline{V}_a \\ \overline{V}_b \\ \overline{V}_c \end{bmatrix}$ $\tilde{\boldsymbol{V}}_{012} = \left[\overline{T}\right]^{-1}\tilde{\boldsymbol{V}}_{abc}$	$\begin{bmatrix} \overline{V}_a \\ \overline{V}_b \\ \overline{V}_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \overline{a}^2 & \overline{a} \\ 1 & \overline{a} & \overline{a}^2 \end{bmatrix}\begin{bmatrix} \overline{V}_0 \\ \overline{V}_1 \\ \overline{V}_2 \end{bmatrix}$ $\tilde{\boldsymbol{V}}_{abc} = \left[\overline{T}\right]\tilde{\boldsymbol{V}}_{012}$
Current	$\begin{bmatrix} \overline{I}_0 \\ \overline{I}_1 \\ \overline{I}_2 \end{bmatrix} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & \overline{a} & \overline{a}^2 \\ 1 & \overline{a}^2 & \overline{a} \end{bmatrix}\begin{bmatrix} \overline{I}_a \\ \overline{I}_b \\ \overline{I}_c \end{bmatrix}$ $\overline{I}_{012} = \left[\overline{T}\right]^{-1}\overline{I}_{abc}$	$\begin{bmatrix} \overline{I}_a \\ \overline{I}_b \\ \overline{I}_c \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 \\ 1 & \overline{a}^2 & \overline{a} \\ 1 & \overline{a} & \overline{a}^2 \end{bmatrix}\begin{bmatrix} \overline{I}_0 \\ \overline{I}_1 \\ \overline{I}_2 \end{bmatrix}$ $\overline{I}_{abc} = \left[\overline{T}\right]\overline{I}_{012}$
Impedance	$\left[\overline{Z}_{012}\right] = \left[\overline{T}\right]^{-1}\left[\overline{Z}_{abc}\right]\left[\overline{T}\right]$	
Power	$\overline{S}_{3\phi} = \overline{V}_a\overline{I}_a^* + \overline{V}_b\overline{I}_b^* + \overline{V}_c\overline{I}_c^* = \tilde{\boldsymbol{V}}_{abc}^T\tilde{\boldsymbol{I}}_{abc}^*$ $\overline{S}_{3\phi} = 3\left\{\overline{V}_0\overline{I}_0^* + \overline{V}_2\overline{I}_2^* + \overline{V}_3\overline{I}_3^*\right\} = 3\tilde{\boldsymbol{V}}_{012}^T\tilde{\boldsymbol{I}}_{012}^*$	

Application of these relationships defined in Table 4.4.2-1 will enable the power system analyst to draw the zero, positive and negative sequence networks for the system under study. These networks can then be analyzed in the 0-1-2 reference frame, and the results can be easily transformed back into the a-b-c reference frame.

Table 4.4.2-2 Summary of the Symmetrical Components in the Balanced Case

Quantity	Transformation Equations: abc $\Rightarrow$ 012	Transformation Equations: 012 $\Rightarrow$ abc
Voltage	Positive Phase Sequence: $\begin{bmatrix}\overline{V}_0\\\overline{V}_1\\\overline{V}_2\end{bmatrix}=[\overline{T}]^{-1}\begin{bmatrix}\overline{V}_a\\\overline{V}_b\\\overline{V}_c\end{bmatrix}=\begin{bmatrix}0\\\overline{V}_a\\0\end{bmatrix}$	Positive Phase Sequence: $\begin{bmatrix}\overline{V}_a\\\overline{V}_b\\\overline{V}_c\end{bmatrix}=[\overline{T}]\begin{bmatrix}\overline{V}_0\\\overline{V}_1\\\overline{V}_2\end{bmatrix}=\begin{bmatrix}\overline{V}_1\\\overline{a}^2\overline{V}_1\\\overline{a}\overline{V}_1\end{bmatrix}$
	Negative Phase Sequence: $\begin{bmatrix}\overline{V}_0\\\overline{V}_1\\\overline{V}_2\end{bmatrix}=[\overline{T}]^{-1}\begin{bmatrix}\overline{V}_a\\\overline{V}_b\\\overline{V}_c\end{bmatrix}=\begin{bmatrix}0\\0\\\overline{V}_a\end{bmatrix}$	Negative Phase Sequence: $\begin{bmatrix}\overline{V}_a\\\overline{V}_b\\\overline{V}_c\end{bmatrix}=[\overline{T}]\begin{bmatrix}\overline{V}_0\\\overline{V}_1\\\overline{V}_2\end{bmatrix}=\begin{bmatrix}\overline{V}_2\\\overline{a}\overline{V}_2\\\overline{a}^2\overline{V}_2\end{bmatrix}$
Current	Positive Phase Sequence: $\begin{bmatrix}\overline{I}_0\\\overline{I}_1\\\overline{I}_2\end{bmatrix}=[\overline{T}]^{-1}\begin{bmatrix}\overline{I}_a\\\overline{I}_b\\\overline{I}_c\end{bmatrix}=\begin{bmatrix}0\\\overline{I}_a\\0\end{bmatrix}$	Positive Phase Sequence: $\begin{bmatrix}\overline{I}_a\\\overline{I}_b\\\overline{I}_c\end{bmatrix}=[\overline{T}]\begin{bmatrix}\overline{I}_0\\\overline{I}_1\\\overline{I}_2\end{bmatrix}=\begin{bmatrix}\overline{I}_1\\\overline{a}^2\overline{I}_1\\\overline{a}\overline{I}_1\end{bmatrix}$
	Negative Phase Sequence: $\begin{bmatrix}\overline{I}_0\\\overline{I}_1\\\overline{I}_2\end{bmatrix}=[\overline{T}]^{-1}\begin{bmatrix}\overline{I}_a\\\overline{I}_b\\\overline{I}_c\end{bmatrix}=\begin{bmatrix}0\\0\\\overline{I}_a\end{bmatrix}$	Negative Phase Sequence: $\begin{bmatrix}\overline{I}_a\\\overline{I}_b\\\overline{I}_c\end{bmatrix}=[\overline{T}]\begin{bmatrix}\overline{I}_0\\\overline{I}_1\\\overline{I}_2\end{bmatrix}=\begin{bmatrix}\overline{I}_1\\\overline{a}\overline{I}_1\\\overline{a}^2\overline{I}_1\end{bmatrix}$
Impedance	$[\overline{Z}_{012}]=[\overline{T}]^{-1}[\overline{Z}_{abc}][\overline{T}]=\begin{bmatrix}\overline{Z}_s+2\overline{Z}_m+3\overline{Z}_n & 0 & 0\\0 & \overline{Z}_s-\overline{Z}_m & 0\\0 & 0 & \overline{Z}_2-\overline{Z}_m\end{bmatrix}$	
Power	$\overline{S}_{3\phi}=\overline{V}_a\overline{I}_a^*+\overline{V}_b\overline{I}_b^*+\overline{V}_c\overline{I}_c^*=3\overline{V}_a\overline{I}_a^*$ $\overline{S}_{3\phi}=\{\overline{V}_0\overline{I}_0^*+\overline{V}_1\overline{I}_1^*+\overline{V}_2\overline{I}_2^*\}=\begin{cases}3\overline{V}_1\overline{I}_1^* \textit{ positive ph. seq}\\3\overline{V}_2\overline{I}_2^* \textit{ negative ph. seq}\end{cases}$	

4.4.3 INSTANTANEOUS SYMMETRICAL COMPONENTS THEORY FOR COMPENSATION

A three-phase, four-wire compensated system is shown in Figure 4.4.3-1, where the three-phase load may be unbalanced and nonlinear, while the supply voltage may be unbalanced and distorted. A shunt APF (or compensator) and the load are connected at the PCC. For the sake of illustrating the concept, the compensator is considered to be ideal and it is comprised of three ideal current sources as shown in the Figure 4.4.3-1.

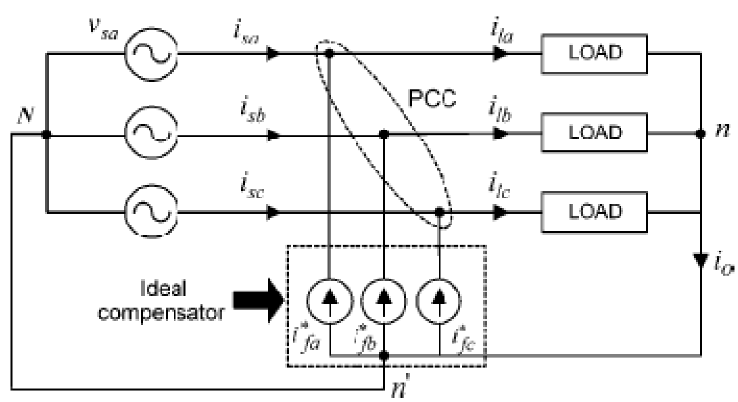


Figure 4.4.3-1: Schematic of Three-Phase, Four-Wire Compensated System

Let the unbalanced and distorted voltages be represented by

$$v_{sa}(t) = \sum_{n=1}^{k} V_{ma\,n} \sin\left(n\omega t + \emptyset_{va\,n}\right) \quad (4.4.3\text{-}1a)$$

$$v_{sb}(t) = \sum_{n=1}^{k} V_{mb\,n} \sin\left(n\omega t + \emptyset_{vb\,n}\right) \quad (4.4.3\text{-}1b)$$

$$v_{sc}(t) = \sum_{n=1}^{k} V_{mb\,n} \sin\left(n\omega t + \emptyset_{vc\,n}\right) \quad (4.4.3\text{-}1c)$$

The subscript “s” stands for supply, “a”, “b”, “c” for the three phases notation, “m” for maximum or peak value and “n” for the harmonic number. The term “k” is the order of maximum harmonic considered in the supply voltages.

Let us denote instantaneous positive, negative, and zero sequence voltages for phase-a by $v_{sa}^{+}(t)$, $v_{sa}^{-}(t)$ and $v_{sa}^{0}(t)$, respectively. These sequence voltages are expressed using symmetrical transformation as the following:

$$\begin{bmatrix} v_{sa}^{0}(t) \\ v_{sa}^{+}(t) \\ v_{sa}^{-}(t) \end{bmatrix} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & a & a^2 \\ 1 & a^2 & a \end{bmatrix}\begin{bmatrix} v_{sa}(t) \\ v_{sb}(t) \\ v_{sc}(t) \end{bmatrix} \tag{4.4.3-2}$$

Similarly, the instantaneous load currents can be resolved into its instantaneous zero, positive, negative sequence components as

$$\begin{bmatrix} i_{la}^{0}(t) \\ i_{la}^{+}(t) \\ i_{la}^{-}(t) \end{bmatrix} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & a & a^2 \\ 1 & a^2 & a \end{bmatrix}\begin{bmatrix} i_{la}(t) \\ i_{lb}(t) \\ i_{lc}(t) \end{bmatrix} \tag{4.4.3-3}$$

Here "a" is a complex operator equal to $e^{j2\pi/3}$. The aforementioned sequence components are denoted by boldface letters as they are complex quantities as a function of time. The zero sequence components are however not complex quantities, still they are represented by bold faced letters to maintain uniformity in equations. In a balanced distorted system with a fundamental frequency of ω and for a non-negative integer n, the harmonics of the order of $3n+1$, $3n+2$, and $3n+3$ follow the positive, negative, and zero sequences, respectively [J22]. Then$v_{sa}^{+}(t)$, $v_{sa}^{-}(t)$ and $v_{sa}^{0}(t)$ in (4.4.3-2), have $3n+1$, $3n+2$, and $3n+3$ harmonics, respectively. But in the unbalanced distorted system$v_{sa}^{+}(t)$, $v_{sa}^{-}(t)$ and $v_{sa}^{0}(t)$ have all of the harmonic orders. The balanced steady state harmonic components can be determined by using the following expressions. In general, for the harmonic order, it is written as follows:

$$\boldsymbol{V}_{sa(3n+1)}^{+} = \frac{\sqrt{2}}{T}\int_{t1}^{t1+T} v_{sa}^{+}\,(t)e^{-j((3n+1)\omega t-\pi/2} \tag{4.4.3-4}$$

where $n = 0, 1, 2, \dots \dots k$. The term is an arbitrary instant and T is the time period of a cycle. The aforementioned integration is carried out using a moving average filter to have a fast response of the compensator. From the expression just shown $\left|\boldsymbol{V}_{sa(3n+1)}^{+}\right|$ and $\angle\boldsymbol{V}_{sa(3n+1)}^{+}$ can be obtained. The balanced voltage harmonics are given by

$$\boldsymbol{V}_{sa(3n+2)}^{-} = \frac{\sqrt{2}}{T}\int_{t1}^{t1+T} v_{sa}^{-}(t)e^{-j((3n+2)\omega t-\pi/2} \qquad (4.4.3\text{-}5)$$

From (4.4.3-5), $\left|\boldsymbol{V}_{sa(3n+2)}^{-}\right|$ and $\angle\boldsymbol{V}_{sa(3n+2)}^{-}$ can be obtained. Similarly, the $3n+3$ voltage harmonics can be obtained as follows:

$$\boldsymbol{V}_{sa(3n+3)}^{0} = \frac{\sqrt{2}}{T}\int_{t1}^{t1+T} v_{sa}^{0}(t)e^{-j((3n+3)\omega t-\pi/2} \qquad (4.4.3\text{-}6)$$

Using (4.4.3-6), $\left|\boldsymbol{V}_{sa(3n+3)}^{0}\right|$ and $\angle\boldsymbol{V}_{sa(3n+3)}^{0}$ are computed.

Thus, the balanced quantities in the case of unbalanced and distorted supply voltages can be given by (4.4.3-7)–(4.4.3-9), shown at the next page, where $n = 0, 1, 2, \ldots\ldots k$. in (4.4.3-7)–(4.4.3-9).

$$v_{sa(3n+1)}^{+}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+1)}^{+}\right| \sin\left((3n+1)\omega t) + \angle\boldsymbol{V}_{sa(3n+1)}^{+}\right) \qquad (4.4.3\text{-}7a)$$

$$v_{sb(3n+1)}^{+}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+1)}^{+}\right| \sin\left((3n+1)\omega t - 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+1)}^{+}\right) \qquad (4.4.3\text{-}7b)$$

$$v_{sc(3n+1)}^{+}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+1)}^{+}\right| \sin\left((3n+1)\omega t + 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+1)}^{+}\right) \qquad (4.4.3\text{-}7c)$$

$$v_{sa(3n+2)}^{-}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+2)}^{-}\right| \sin\left((3n+2)\omega t) + \angle\boldsymbol{V}_{sa(3n+2)}^{-}\right) \qquad (4.4.3\text{-}8a)$$

$$v_{sb(3n+2)}^{-}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+2)}^{-}\right| \sin\left((3n+2)\omega t - 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+2)}^{-}\right) \qquad (4.4.3\text{-}8b)$$

$$v_{sc(3n+2)}^{-}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+2)}^{-}\right| \sin\left((3n+2)\omega t + 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+2)}^{-}\right) \qquad (4.4.3\text{-}8c)$$

$$v_{sa(3n+3)}^{0}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+3)}^{0}\right| \sin\left((3n+3)\omega t) + \angle\boldsymbol{V}_{sa(3n+2)}^{0}\right) \qquad (4.4.3\text{-}9a)$$

$$v_{sb(3n+3)}^{0}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+3)}^{0}\right| \sin\left((3n+3)\omega t - 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+2)}^{0}\right) \qquad (4.4.3\text{-}9b)$$

$$v_{sc(3n+3)}^{0}(t) = \sqrt{2}\left|\boldsymbol{V}_{sa(3n+3)}^{0}\right| \sin\left((3n+3)\omega t + 2\pi/3) + \angle\boldsymbol{V}_{sa(3n+2)}^{0}\right) \qquad (4.4.3\text{-}9c)$$

The three-phase balanced set of voltages in the case of unbalanced and distorted supply voltages can be written as

$$v'_{sa}(t) = \sum_{n=0}^{k} \left(v^{+}_{sa(3n+1)}(t) + v^{-}_{sa(3n+2)}(t) + v^{0}_{sa(3n+3)}(t) \right) \quad \text{(4.4.3-10a)}$$

$$v'_{sb}(t) = \sum_{n=0}^{k} \left(v^{+}_{sb(3n+1)}(t) + v^{-}_{sb(3n+2)}(t) + v^{0}_{sb(3n+3)}(t) \right) \quad \text{(4.4.3-10b)}$$

$$v'_{sc}(t) = \sum_{n=0}^{k} \left(v^{+}_{sc(3n+1)}(t) + v^{-}_{sc(3n+2)}(t) + v^{0}_{sc(3n+3)}(t) \right) \quad \text{(4.4.3-10c)}$$

Now v'_{sa} , v'_{sb} and v'_{sc} form balanced quantities for the available distorted and unbalanced supply voltages.

Let the source currents after compensation be i_{sa} , i_{sb} and i_{sc} in phases *a*, *b* and *c* respectively, as shown in Figure 4.4.3-1. In order to meet the requirements of load compensation, the following three conditions are to be satisfied. The first condition is that the neutral current after compensation must be zero. Therefore

$$i_{sa} + i_{sb} + i_{sc} = 0 \quad \text{(4.4.3-11)}$$

The second objective of compensation is that the reactive power delivered from the source is controlled by the phase angle between the positive sequence voltage and current, hence

$$\angle\left(\boldsymbol{v^{+}_{sa}(t)}\right) = \angle\left(i^{+}_{sa}(t)\right) + \emptyset^{+} \quad \text{(4.4.3-12)}$$

This implies that

$$\angle(v_{sa} + a v_{sb} + a^2 v_{sc}) = \angle(i_{sa} + a i_{sb} + a^2 i_{sc}) + \emptyset^{+}$$

Simplifying the equation that was just shown leads to

$$(v_{sb} - v_{sc} - 3\gamma v_{sa})i_{sa} + (v_{sc} - v_{sa} - 3\gamma v_{sb})i_{sb} + (v_{sa} - v_{sb} - 3\gamma v_{sc})i_{sc} = 0$$

(4.4.3-13)

Where $\gamma = \tan \emptyset^{+}/\sqrt{3}$ and $\emptyset^{+}$ is the angle between the instantaneous phasors $\boldsymbol{v}_{sa}^{+}$ and $\boldsymbol{i}_{sa}^{+}$. In order to control the reactive power from the source through the negative sequence quantities, the following condition should be met:

$$\angle(\boldsymbol{v}_{sa}^{-}(\boldsymbol{t})) = \angle(i_{sa}^{-}(t)) + \emptyset^{-} \quad (4.4.3\text{-}14)$$

This gives

$$(v_{sc} - v_{sb} - 3\beta v_{sa})i_{sa} + (v_{sa} - v_{sc} - 3\beta v_{sb})i_{sb} + (v_{sb} - v_{sa} - 3\beta v_{sc})i_{sc} = 0$$

(4.4.3-15)

where $\beta = \tan \emptyset^{-}/\sqrt{3}$ and $\emptyset^{-}$ is the angle between the instantaneous phasors $\boldsymbol{v}_{sa}^{-}$ and $\boldsymbol{i}_{sa}^{-}$. By keeping both $\gamma = 0$ and $\beta = 0$ means that the reactive power supplied from the source through the positive sequence and negative sequence components of voltages and currents is zero. Alternatively, the positive sequence currents and negative sequence currents are in phase with the positive sequence voltages and negative sequence voltages, respectively. It results in the following equation:

$$(v_{sb} - v_{sc})i_{sa} + (v_{sc} - v_{sa})i_{sb} + (v_{sa} - v_{sb})i_{sc} = 0 \quad (4.4.3\text{-}16)$$

The third condition is that the source should supply the average load power P_{lavg} i.e., $v_{sa}i_{sa} + v_{sb}i_{sb} + v_{sc}i_{sc} = P_{lavg}$ (4.4.3-17)

By solving (4.4.3-11), (4.4.3-16), and (4.4.3-17), the reference source currents for compensation are obtained as follows:

$$i_{sa} = \frac{v_{sa} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}18a)$$

$$i_{sb} = \frac{v_{sb} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}18b)$$

$$i_{sc} = \frac{v_{sc} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}18c)$$

Where $v_{sa}^0 = (1/3) \sum_{j=a,b,c} v_{sj}$ and $\Delta = \sum_{j=a,b,c} v_{sj}^2 - 3\,(v_{sa}^0)^2$

For the balanced sinusoidal conditions, is zero. For the supply voltages with unbalanced fundamental and unbalanced harmonics, it is not zero. Even for balanced distorted supply voltages, it is nonzero because of the triplen harmonics. By knowing the reference source currents, the APF reference currents can be obtained as shown

$$i_{fa}^* = i_{la} - i_{sa} = i_{la} - \frac{v_{sa} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}19a)$$

$$i_{fb}^* = i_{lb} - i_{sb} = i_{lb} - \frac{v_{sb} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}19b)$$

$$i_{fc}^* = i_{lc} - i_{sc} = i_{lc} - \frac{v_{sc} - v_{sa}^0}{\Delta} P_{lavg} \quad (4.4.3\text{-}19c)$$

While deriving the reference compensator currents in (4.4.3-19), in general, it is not assumed that supply voltages should be balanced and sinusoidal. However, the compensator meets the requirements as follows.

• There is no neutral current after compensation.

• The source supplies average load power and the rest of the load power is supplied by the compensator.

4.5 INDUCTION MOTOR ANALYSIS

Three-phase induction motors are of two types: squirrel-case and wound-rotor. In squirrel-cage, the rotor consists of longitudinal conductor-bars shorted by circular connectors at the two ends while in wound-rotor motor, the rotor also has a balanced three-phase distributed winding having same poles as stator winding. However, in both, stator carries a three-phase balanced distributed winding.

4.5.1 PER PHASE EQUIVALENT CIRCUIT OF MOTOR

Per-phase equivalent circuit of a three-phase induction motor is shown in Figure 4.5.1-1(a). *R'* and X' are the stator referred values of rotor resistance *R,* and rotor reactance *X.* Slip is defined by

$$s = \frac{\omega_{ms} - \omega_m}{\omega_{ms}} \tag{4.5.1-1}$$

where ω_m and ω_{ms} are rotor and synchronous speeds, respectively. Further

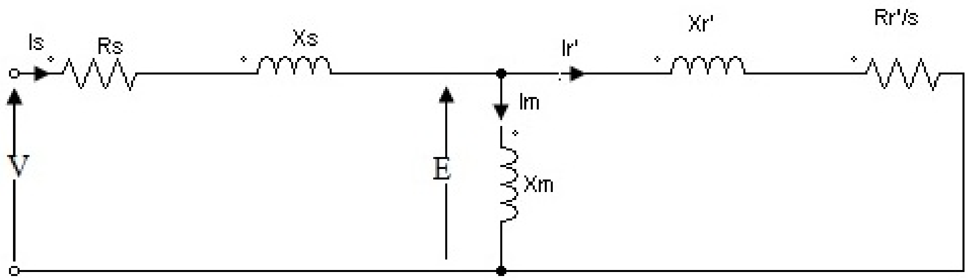
$$\omega_{ms} = \frac{4\Pi f}{p} \ rad/sec \tag{4.5.1-2}$$

Where f and *p* are supply frequency and number of poles, respectively.

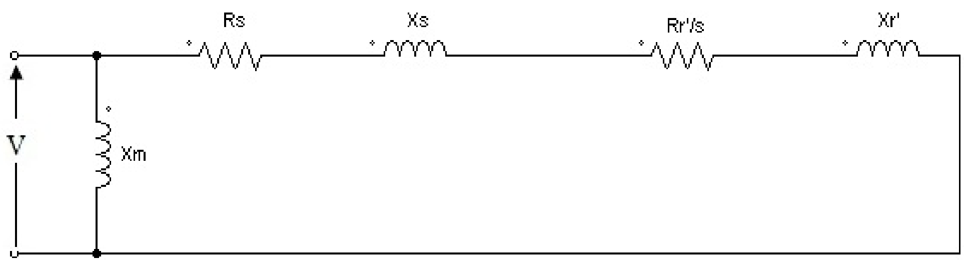
Since, stator impedance drop is generally negligible compared to terminal voltage V, the equivalent circuit can be Simplified to that shown in Figure 4.5.1-1(b).

Also from Equation (4.5.1-1)

$$\omega_m = \omega_{ms} (1 - s) \tag{4.5.1-3}$$



(a)



(b)

Figure 4.5.1-1: Per-Phase Stator Referred Equivalent Circuits of An Induction Motor

From Figure 4.5.1-1(b);

$$\bar{I}_p' = \frac{V}{\left(R_s + \frac{R_r'}{s}\right) + j\,(X_s + X_r')} \qquad (4.5.1\text{-}4)$$

Power transferred to rotor (or air-gap power)

$$P_g = 3\, I_r'^2\, R_r' \,/\, s \qquad (4.5.1\text{-}5)$$

Rotor copper loss is

$$p_{cu} = 3\, I_r'^2\, R_r' \qquad (4.5.1\text{-}6)$$

Electrical power converted into mechanical power

$$P_m = P_g - p_{cu} = 3\, I_r'^2\, R_r' \left(\frac{1-s}{s}\right) \tag{4.5.1-7}$$

Torque developed by motor

$$T = P_m \,/\, \omega_m \tag{4.5.1-8}$$

Substituting from Equations (4.5.1-3)and (4.5.1-7) yields

$$T = \frac{3}{\omega_{ms}}\, I_r'^2\, \frac{R_r'}{s} \tag{4.5.1-9}$$

Substituting from Equation (4.5.1-4) gives

$$T = \frac{3}{\omega_{ms}} \left[\frac{V^2\, R_{r/s}'}{\left(R_s + \frac{R_r'}{s}\right)^2 + \left(X_s + X_r'\right)^2}\right] \tag{4.5.1-10}$$

A comparison of Equations (4.5.1-5) and (4.5.1-9) suggests that

$$T = P_g \,/\, \omega_{ms} \tag{4.5.1-11}$$

Motor output torque at the shaft is obtained by deducting friction windage and *core-loss* torques from the developed torque. The developed torque is a function of slip only (Equation 4.5.1-10). Differentiating *T* (4.5.1-10) with respect to s and equating to zero gives the slip for maximum torque

$$s_m = \pm \frac{R_r'}{\sqrt{R_S^2 + \left(X_s + X_r'\right)^2}} \tag{4.5.1-12}$$

Substituting from Equation (4.5.1-12) into (4.5.1-10) yields an expression for maximum torque

$$T_{max} = \frac{3}{2\omega_{ms}} \left[\frac{V^2}{R_s \pm \sqrt{R_S^2 + \left(X_s + X_r'\right)^2}}\right] \tag{4.5.1-13}$$

Maximum torque is also known as breakdown torque. While it is independent of rotor resistance, s_m is directly proportional to rotor resistance.

The natures of speed-torque and speed-rotor current characteristics are shown in Figure 4.3-1(b). Both rotor-current and torque are zero at synchronous speed. With decrease in speed, both increase, While torque reduces after reaching breakdown value, the rotor-current continues to increase, reaching a maximum value at zero speed, Drop in speed from no load to full load depends on the rotor resistance. When rotor resistance is low, the drop is quite small, and therefore, motor operates essentially at a constant speed. The breakdown torque is a measure of short-time torque overload capability of the motor.

Motor runs in the direction of the rotating field. Direction of rotating field and therefore motor speed can be reversed by reversing the phase sequence. Phase sequence can be reversed by interchanging any two terminals of the motor.

4.5.2 OPERATION WITH UNBALANCED SOURCE VOLTAGES AND SINGLE PHASING

As Supply voltage may sometimes become unbalanced, it is useful to know the effect of unbalanced voltages on motor performance. Further, motor terminal voltage may be unbalanced intentionally for speed control or starting as described later. A three-phase set of unbalanced voltages (V_a, V_b, and V_c) can be resolved into three-phase balanced positive sequence (V_p), negative sequence (V_n) and zero sequence (V_o) voltages, using symmetrical component relations as mentioned in section 4.4:

Motor performance can be calculated for positive and negative sequence voltages separately. Resultant performance is obtained by the principle of superposition by assuming motor to be a linear system. Positive sequence voltages produce an air-gap flux wave which rotates at synchronous speed in the forward direction. For a forward rotor speed $w_m{}'$ slip s is given by Equation (4.5.1-1). For positive sequence voltages, equivalent circuits are same as shown in Figure 4.5.1-1, except that V is replaced by V_p. The positive sequence rotor current and torque are obtained by replacing V *by* V_p in Equations (4.5.1-4) and (4.5.1-10). Thus

$$I'_{rp} = \frac{V_p}{\left(R_s + \frac{R'_r}{s}\right) + j\,(X_s + X'_r)}$$

$$T_p = \frac{3}{\omega_{ms}}\left[\frac{V_p^2\, R'_{r/s}}{\left(R_s + \frac{R'_r}{s}\right)^2 + \left(X_s + X'_r\right)^2}\right] \qquad (4.5.2\text{-}1)$$

Negative sequence voltages produce an air-gap flux wave which rotates at synchronous speed in the reverse direction. The slip is

$$s_n = \frac{-\omega_{ms} - \omega_m}{-\omega_{ms}}$$

Substitution from Equation (4.5.1-3) gives

$$s_n = (2 - s) \tag{4.5.2-2}$$

Again, equivalent circuits of Figure 4.5.1-1 are applicable when *s* is replaced by (2 - *s)* or s_n and *V are* replaced by V_n. Proceeding as in Sec. 4.5.1, following expressions are obtained for rotor current and torque:

$$I'_{rn} = \frac{V_n}{\left(R_s + \frac{R'_r}{2-s}\right) + j\,(X_s + X'_r)}$$

$$T_n = -\frac{3}{\omega_{ms}}\left[\frac{V_n^2\, R'_{r/(2-s)}}{\left(R_s + \frac{R'_r}{2-s}\right)^2 + (X_s + X'_r)^2}\right] \tag{4.5.2-3}$$

Torque has a negative sign because for negative sequence voltages the synchronous speed is $(-\omega_{ms})$. The rms rotor current and torque are given by

$$I'_r = \left(I'^2_{rp} + I'^2_{rn}\right)^{1/2} \tag{4.5.2-4}$$

$$T = T_p + T_n$$

$$= \frac{3}{\omega_{ms}}\left[\frac{V_p^2\, R'_{r/s}}{\left(R_s + \frac{R'_r}{s}\right)^2 + (X_s + X'_r)^2} - \frac{V_n^2\, R'_{r/(2-s)}}{\left(R_s + \frac{R'_r}{2-s}\right)^2 + (X_s + X'_r)^2}\right] \tag{4.5.2-5}$$

Positive sequence and negative sequence speed-torque characteristics are shown in Figure 4.5.2-1(a). *Single phasing (when supply to anyone phase fails) is the extreme case of unbalancing, when V_p = V_n .* At zero speed, *s* is also equal to s_n *consequently* starting torque is zero. Speed-torque curves for single phasing are shown in Figure 4.5.2-1(b). Interaction between positive sequence air-gap flux wave and positive sequence rotor currents produce positive sequence torque T_p. Negative sequence torque T_n is produced due to interaction between negative sequence flux wave and negative sequence rotor currents.

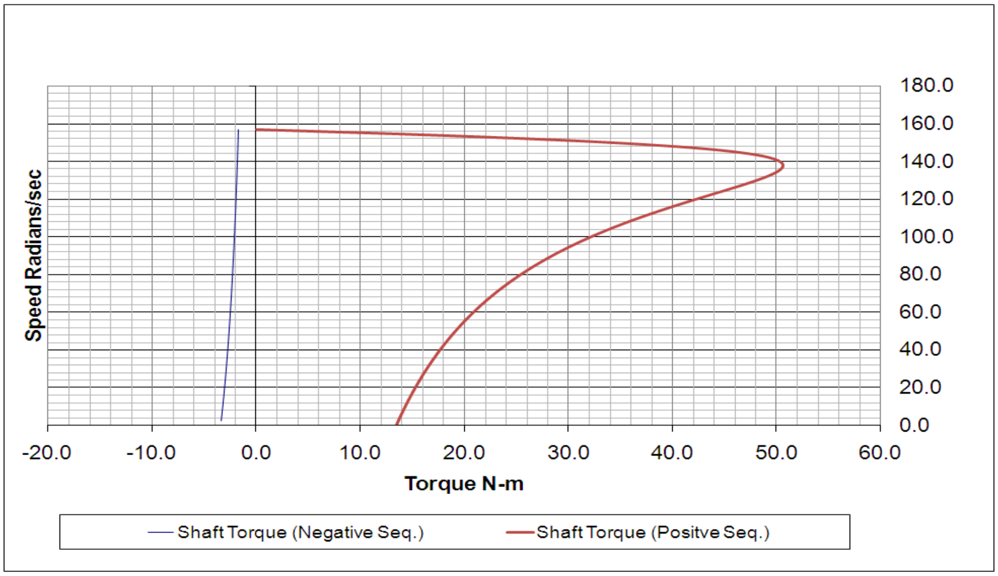


Figure 4.5.2-1(a): Speed-Torque Curves of An Induction Motor With Unbalance Stator Voltages

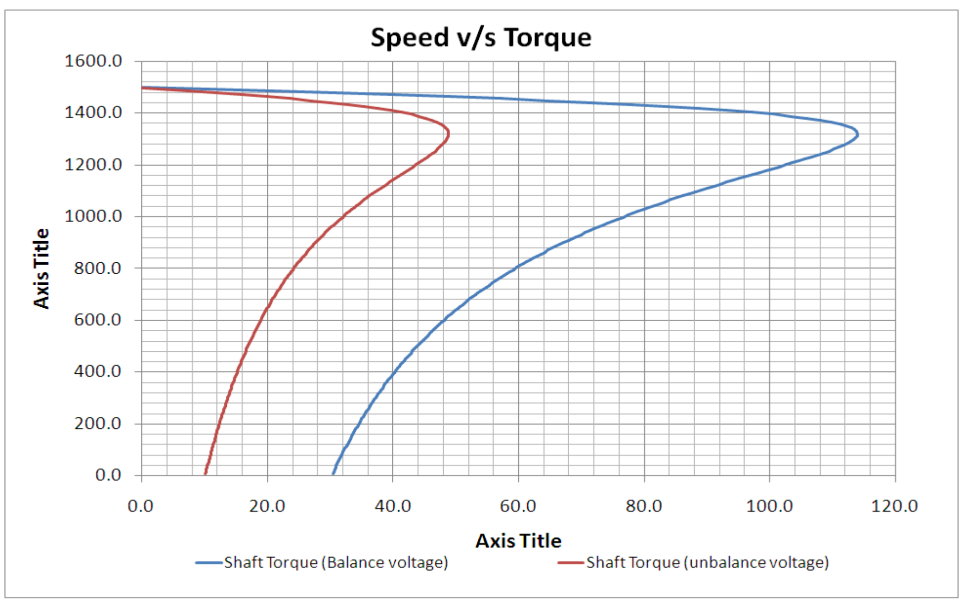


Figure 4.5.2-1(b): Speed-Torque Curves Of An Induction Motor With Unbalance Stator Voltages & Balance Stator Voltage For Single Phasing

Interactions between positive sequence flux wave and negative sequence rotor currents, and negative sequence flux wave and positive sequence rotor currents, *also* produce torques. However, these torques are pulsating in nature with zero average values. The pulsating torques cause vibrations which reduce the life of motor and produce hum.

Equations (4.5.2-5) and (4.5.2-4) suggest that while the torque is reduced, copper losses (and also core losses) are increased. Thus, the unbalanced operation substantially reduces the motor torque capability and efficiency. To prevent burning of the motor, it is not allowed to run for a prolonged period when the unbalance in voltages is more than 5%. For the same reason, motor is disconnected from the source whenever single phasing occurs, unless the single phasing is always accompanied by a light load.

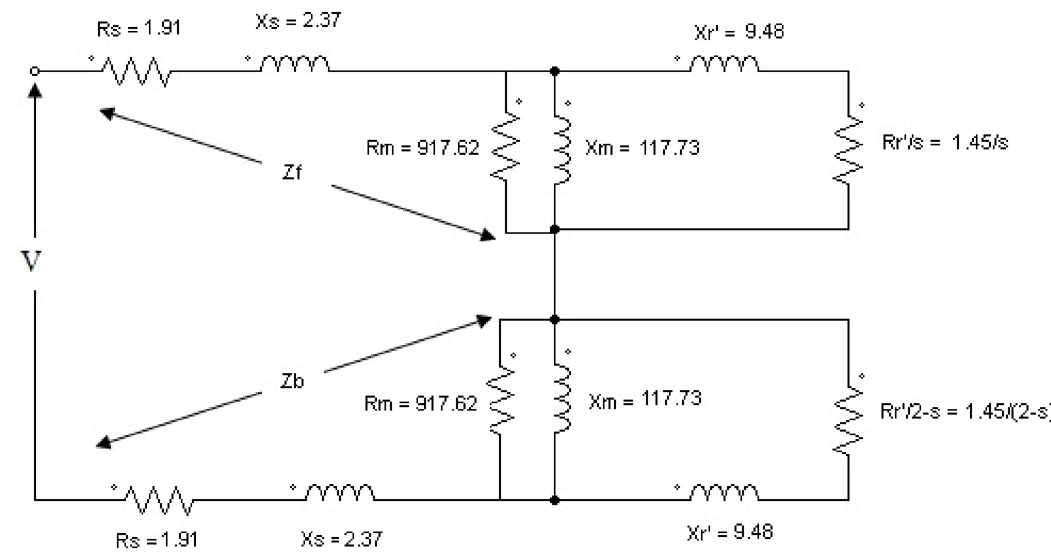


Figure 4.5.2-2: Per phase Equivalent circuit of Induction motor for calculation of speed torque characteristics under unbalance stator voltage condition.

From equivalent circuit of Figure 4.5.2-2, one can obtain speed-torque curve for the motor. This curve and the load speed-torque curve are plotted on a graph. Intersection provides the values of steady state speed and torque. As a sample motor current and torque for a slip of 0.035 is calculated in Annexure-III.

4.5.3 ANALYSIS OF INDUCTION MOTOR FED FROM NON·SINUSOIDAL VOLTAGE SUPPLY

When fed from an inverter or cyclo-converter, the motor terminal voltage is non-sinusoidal but it has half-wave symmetry. A non-sinusoidal waveform can be resolved into fundamental and harmonic components using Fourier analysis. Because of half-wave symmetry only odd harmonics will be present. The harmonics can be divided into positive sequence, negative sequence and zero sequence. The harmonics, which have the same phase sequence as that of fundamental are called positive sequence harmonics. The harmonics having phase sequence opposite to fundamental are called negative sequence harmonics. The harmonics, which have all three-phase voltages in phase, are called zero sequence harmonics.

Consider the fundamental phase voltage components V_{AN} = Vl sin wt, V_{BN}=Vl sin(wt-2π/3) and V_{CN} = Vl sin(wt-4π/3) with the phase sequence ABC. The corresponding 5th and 7th harmonic phase voltages are

$$V_{AN} = V_5 \sin 5\omega t$$

$$V_{BN} = V_5 \sin 5(\omega t - 2\Pi/3) = V_5 \sin(5\omega t - 4\Pi/3)$$

$$V_{CN} = V_5 \sin 5(\omega t - 4\Pi/3) = V_5 \sin (5\omega t - 2\Pi/3)$$

and

$$V_{AN} = V_7 \sin 7\omega t$$

$$V_{BN} = V_7 \sin 7(\omega t - 2\Pi/3) = V_7 \sin (7\omega t - 2\Pi/3)$$

$$V_{CN} = V_7 \sin 7(\omega t - 4\Pi/3) = V_7 \sin (7\omega t - 4\Pi/3)$$

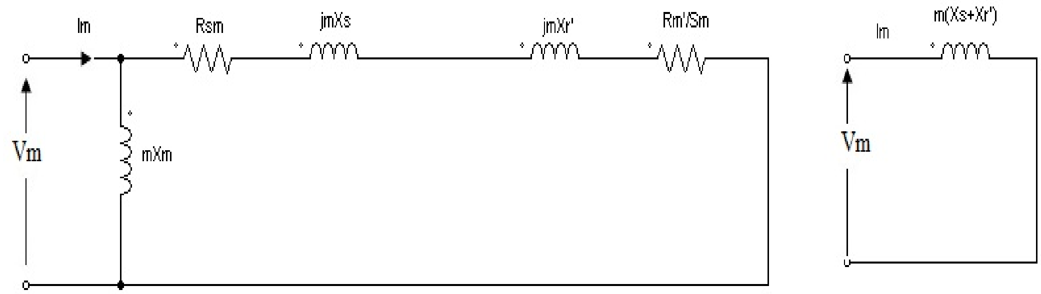
The above equations show that 7th harmonic has the phase sequence A-B-C, which is the same as that of fundamental. Hence it is a positive sequence harmonic. The 5th harmonic has a phase sequence A-C-B, hence it is a negative sequence harmonic. It can be shown that the harmonic voltages and currents of the order m = *6k* + 1 (where *k* is an integer) are of positive sequence and harmonic voltages of the order m = *6k* - I are of negative sequence. Similarly it can be shown that harmonics of the order m = *3k* are of zero sequence. A positive sequence harmonic m will produce a rotating field, which moves in the same direction as the fundamental at a speed m times that of the fundamental field. Similarly rotating field produced by a negative sequence harmonic III will move in the direction opposite to the fundamental at III times its speed. Zero sequence components do not produce a rotating field.

For fundamental component the equivalent circuits of Figure 4.5.1-1 will be applicable. For any m^{th} harmonic, equivalent circuit will be as shown in Figure 4.5.3-1(a). Each reactance has been increased by a factor III. Due to the skin effect resistances will also be increased several times. Slip s_m for the m^{th} harmonic is given by

$$s_m = \frac{m\omega_{ms} \mp \omega_m}{m\omega_{ms}}$$

Negative sign is applicable to harmonics which produce forward rotating fields and the positive sign to those which produce backward rotating fields. Since s_m is close to unity, resistance (R_m'/s_m) has a small value. As reactance are very large compared to resistance, equivalent circuit of Figure 4.5.3-1(a) can be replaced by the simplified circuit of Figure 4.5.3-1(b). When fed from a semiconductor converter, it can be shown that the net torque produced by harmonics is close to zero.

In view of this motor torque can be evaluated from equivalent circuits of Figure 4.5.3-1(b), using Equation (4.5.1-10), where *V* is the fundamental component of supply voltage.



(a) (b)

Figure 4.5.3-1: Harmonic Equivalent Circuits of an Induction Motor

Fundamental component of rotor current is obtained from Equation (4.5.1-4) and the mth harmonic current is calculated from Figure 4.5.3-1(b) as

$$I_m = \frac{V_m}{mX} \tag{4.5.3-1}$$

where X=X_s + X_r'.

Generally, supply will have odd harmonics. When stator is star-connected triplen harmonics (third harmonic and its multiples) will not flow. The rms motor current I_{rms} will then be

$$I_{rms}^2 = I_s^2 + \sum_{m=5,7,11,\ldots} I_m^2 \tag{4.5.3-2}$$

When motor is delta-connected, triplen harmonics will circulate in delta, but will not flow in the source. The source current therefore can be obtained by multiplying I_{rms}. given by Equation (4.5.3-2) by $\sqrt{3}$. The rms motor phase current will be obtained by

$$I_{rms}^2 = I_s^2 + \sum_{m=3,57\ \ldots} I_m^2 \tag{4.5.3-3}$$

For a given motor torque and power, rms current flowing through the motor has a higher value. Further due to skin effect harmonic rotor resistance has higher value. Therefore presence of harmonics, increase the copper loss substantially. Core losses are also increased by harmonics. Because of increase in losses, motor has to be derated in the sense that the power output that can be obtained from machine for the same temperature rise has to be smaller. The efficiency is also reduced due to increase in losses.

Another important effect of non-sinusoidal supply is the production of pulsating torques due to interaction between the rotating field produced by one harmonic and rotor current of another harmonic. Harmonic 5, 7, 11 and 13 are major contributors of torque pulsations. 5^{th} harmonic produces backward rotating field whereas 7^{th} harmonic produces forward rotating field. Therefore, relative speed between the field produced by the fundamental and 5^{th} and 7^{th} harmonics is six time the speed of fundamental. Consequently, torque pulsations produced due to the interaction of 5^{th} and 7^{th} harmonic currents and the fundamental rotating field has a frequency six times the fundamental. It can be similarly shown that harmonics 11 and 13 produce torque pulsations whose frequency is 12 times the fundamental. When motor supply frequency is not very low, the frequency of torque pulsations is large enough to be filtered out by motor inertia. Consequently the torque pulsations do not have significant effect on motor speed, although they do increase noise and reduce motor life due to vibrations. However, when motor supply frequency is low, these torque pulsations cause pulsations in speed. The motor then does not move smoothly but have jerky motion.

4.6 SOLUTION TO NEGATIVE SEQUENCE & ZERO SEQUENCE PROBLEM

This section presents a novel and online method, which is also a novel contribution of this research work for detecting negative & zero sequence voltages and its compensation. It is based on instantaneous Active Reactive power theory. Using this concept, negative sequence and zero sequence components are computed online. The α-β transformation is real unlike the complex transformation matrix in case of symmetrical components. So implementation of α-β analysis on line is simple. The theory is verified through simulation and experimentation results.

The section also analyzes theoretically the effect of negative sequence voltages on induction motors, energy saved due to compensation.

4.6.1 ELECTROMAGNETIC COMPENSATOR:

Figure 4.6.1-1 shows the schematic connection diagram for this system. It requires another 3ϕ squirrel cage motor (much smaller in capacity) whose rotor is mounted in the same shaft and whose stator windings are connected with reverse phase sequence. Stator windings of both the motors are electrically connected in series with the system voltage. The theory behind this method is that the second motor with negative sequence winding connection will offer large impedance to the negative sequence currents in the input lines (caused by the mains voltage unbalance) and low impedance to the positive sequence currents. Thus this connection will reduce the magnitude of negative sequence currents in the stator of the main motor for the same input voltage unbalance factor. Improvement in the performance of the main motor by this method has been reported in reference [J1].

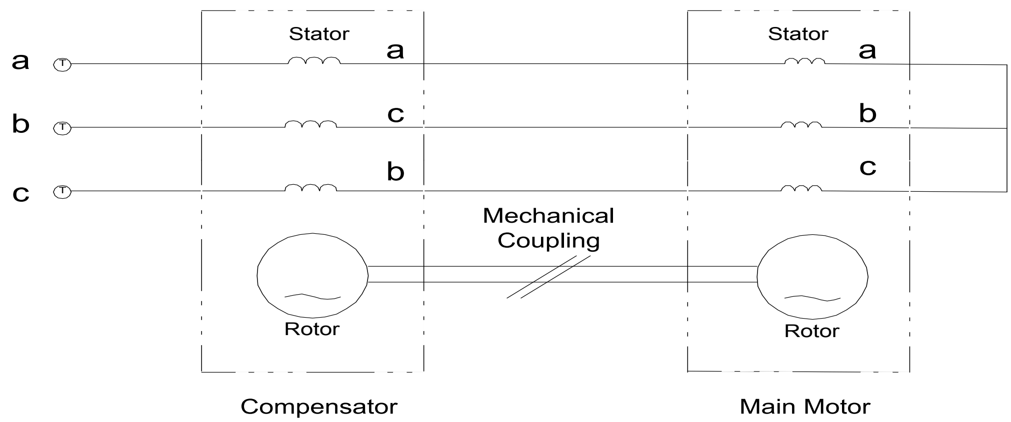


FIGURE 4.6.1-1: Electromagnetic Compensator

The drawback of this method in addition to the cost of another motor is that the stator winding of the second motor which is connected in series with the stator of the main motor while reducing the magnitude of negative sequence current also reduces the positive sequence voltage applied to the main motor since the series winding has some impedance offered to the flow of positive sequence currents. Since the output of induction motor is approximately proportional to the square of the voltage applied, this method results in reduced output of the motor while improving its efficiency due to reduction in negative sequence torque. Further the positive sequence currents flowing through the stator windings of the second motor generate the breaking torque on the rotor reduces the net torque developed by main motor. It is very difficult to retrofit this type of compensator. Also this type of compensator is applicable for motors only.

4.6.2 ELECTRONIC COMPENSATOR

The above problem can be solved by the use of electronic compensator connected in series with the stator of main motor, which will compensate only the negative sequence voltage in the mains, and the positive sequence voltage remains unaffected. This compensator can be used for other types of loads also which are sensitive to negative sequence voltages. This chapter describes the principle of this electronic compensator.

Electronic negative sequence compensator is a voltage source inverter [J2]-[J3] whose output is changed through pulse width modulation (PWM). PWM method utilizes a reference signal (one for each phase), which is compared with a high frequency triangular wave, and the points of intersection of both these signals decide the time of switching on and off the devices connected to that particular phase. The output of PWM inverter follows the reference signal used.

If the reference signal is derived from the negative sequence voltage with opposite polarity then the negative sequence compensator voltages obtained from the PWM inverter which are connected in series with the main motor windings, will cancel the negative sequence component voltages presents in the mains applied voltage.

4.7 CONTORLLING OF NEGATIVE SEQUENCE COMPENSATOR

To control Voltage source inverter, in series with load for negative & zero sequence compensation, reference signal having information of negative sequence & zero sequence of the source is required. For extracting negative sequence and zero sequence from the bus voltage a novel technique using instantaneous active reactive power theory were used. After extracting the signal, the extracted signal was used as reference signal to voltage source inverter along with PI controller to control VSC so that it compensate the negative & zero sequence converter.

4.7.1 GENERATION OF REFERENCE SIGNALS:

It is well known that the average power flow due to positive sequence voltage and negative sequence current is zero

$$(V_m \sin(\omega t) I_m \sin(\omega t) + (V_m \sin(\omega t - 120) I_m \sin(\omega t + 120)) + (V_m \sin(\omega t + 120) I_m \sin(\omega t - 120)) = 0 \tag{4.7.1-1}$$

Similarly the negative sequence voltage will generate nonzero power with negative sequence current.

Hence this concept is used for generating the reference signal for negative sequence compensator using instantaneous Active reactive power theory [J4]-[J8].

To implement this theory, digitally generate reference unit rms amplitude 3-phase negative sequence voltages in synchronization with supply voltage v_a, v_b and v_c as follows.

$$v_{ar} = \sqrt{2}\,\sin(\omega t)$$

$$v_{br} = \sqrt{2}\ \sin\left(\omega t + {}^{2\pi}/_{3}\right)$$

$$v_{cr} = \sqrt{2}\ \sin\left(\omega t - {}^{2\pi}/_{3}\right) \qquad (4.7.1\text{-}2)$$

Convert these reference voltages to α-β-0 reference frame through the transformation matrix as below.

$$\begin{bmatrix} v_{0r} \\ v_{\alpha r} \\ v_{\beta r} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{ar} \\ v_{br} \\ v_{cr} \end{bmatrix} \qquad (4.7.1\text{-}3)$$

Let v_a, v_b and v_c be the phase to neutral unbalanced three phase voltages of the input power supply. It is assumed that these are of fundamental frequency and free of harmonics. These voltages are sensed through voltage sensor and converted into α-β-0 component through the transformation matrix as below.

$$\begin{bmatrix} v_{0} \\ v_{\alpha} \\ v_{\beta} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{a} \\ v_{b} \\ v_{c} \end{bmatrix} \qquad (4.7.1\text{-}4)$$

Now the following calculation is performed using these α-β-0 components.

$$\begin{bmatrix} p_{0} \\ p \\ q \end{bmatrix} = \begin{bmatrix} v_{0r} & 0 & 0 \\ 0 & v_{\alpha r} & v_{\beta r} \\ 0 & -v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} v_{0} \\ v_{\alpha} \\ v_{\beta} \end{bmatrix} \qquad (4.7.1\text{-}5)$$

p_o is zero because e_{ro} is zero and p & q can have dc and ac components as function of time.

In the instantaneous p-q theory DC components in p & q are responsible for the fundamental positive sequence or negative sequence power & this depends upon the reference voltage signal. If reference voltage v_{ra}, v_{rb}, v_{rc} are the positive sequence then DC components in the power p & q reflects the positive sequence power otherwise if reference voltage v_{ra}, v_{rb}, v_{rc} are of negative sequence, then DC components in p & q reflects the fundamental negative sequence power.

In Equation (4.7.1-5), the DC components are due to negative sequence voltages in the bus.

So DC components are filtered out using low pass filter to get the p_{dc} and $q_{dc.}$ Negative sequence component is computed using the following transformation.

$$\begin{bmatrix} v_{\alpha n} \\ v_{\beta n} \end{bmatrix} = \begin{bmatrix} v_{\alpha r} & v_{\beta r} \\ -v_{\beta r} & v_{\alpha r} \end{bmatrix}^{-1} \begin{bmatrix} -p_{dc} \\ -q_{dc} \end{bmatrix}$$

$$\begin{bmatrix} v_{\alpha n} \\ v_{\beta n} \end{bmatrix} = \frac{1}{(v_{\alpha r}{}^2 + v_{\beta r}{}^2)} \begin{bmatrix} v_{\alpha r} & -v_{\beta r} \\ v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} -p_{dc} \\ -q_{dc} \end{bmatrix} \qquad (4.7.1\text{-}6)$$

$v_{\alpha n}$ and $v_{\beta n}$ are the α, β components of negative sequence voltage of the supply, which can be transformed back to three-reference voltages for PWM voltage source inverter.

$$\begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1 & 0 \\ 1/\sqrt{2} & -1/2 & \sqrt{3}/2 \\ 1/\sqrt{2} & -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} -v_0 \\ v_{\alpha n} \\ v_{\beta n} \end{bmatrix} \qquad (4.7.1\text{-}7)$$

Where v_o is the zero sequence voltage component of the three phase bus voltages.

In the above manner, the instantaneous values of negative sequence voltages (including zero sequence components) of phase ‘a’ phase ‘b’ and phase ‘c’ are computed.

When once these three reference signals are obtained they are fed to the voltage source inverter as reference signals to generate corresponding voltages at the output of voltage source inverter which will completely eliminate the negative sequence and zero sequence component voltages in the mains supply and will not affect its positive sequence component voltages which is the problem with the previously proposed electromagnetic compensator [J1]. Further there will be no additional braking torque on the rotor due to the positive sequence currents flowing through negative sequence compensator, as was the case with the other compensator. So, the mechanical output will not be reduced. The active energy input to negative sequence compensator is negligible since its negative sequence output voltage sees positive sequence currents flowing to the main motor and so the negative sequence compensator output is only reactive in nature. Only very little amount of power loss takes place in the inverter and transformer. Further more positive sequence voltage applied to the motor can be improved to its rated voltage by adding a portion of the fundamental positive sequence mains voltage to the reference signals of the negative sequence compensator scheme. Hence further improvement in the effective torque of the motor is also possible.

4.7.2 RESULTS:

The simulation studies are carried out to predict the performance of the proposed new theory using MATLAB-SIMULINK-POWER SYSTEM BLOCK. For simulation of negative sequence compensator a voltage source inverter whose output is changed through pulse width modulation (PWM) is used. PWM method utilizes a reference signal (one for each phase), which is compared with a high frequency triangular wave, and the points of intersection of both these signals decide the time of switching on and off the devices connected to that particular phase. Block diagram for simulation is shown in Figure 4.7.2.1-1. Negative sequence condition is simulated using two methods.

One method is by adding balance three phase negative sequence voltage to the balance three-phase positive sequence voltage (here zero sequence components is absent) and second method is through unbalancing the supply voltage itself which can have zero sequence component.

4.7.2.1 SIMULATION RESULTS & WAVEFORMS

The simulation results for both the methods are shown in Figure 4.7.2.1-2(a) to Figure 4.7.2.1-2(d). In second method, by unbalancing the supply voltage if the zero sequence components are ignored for calculation of the three phase compensating signals, then the output remains still unbalance due to the presence of the zero sequence components in the system. This is also verified through simulation.

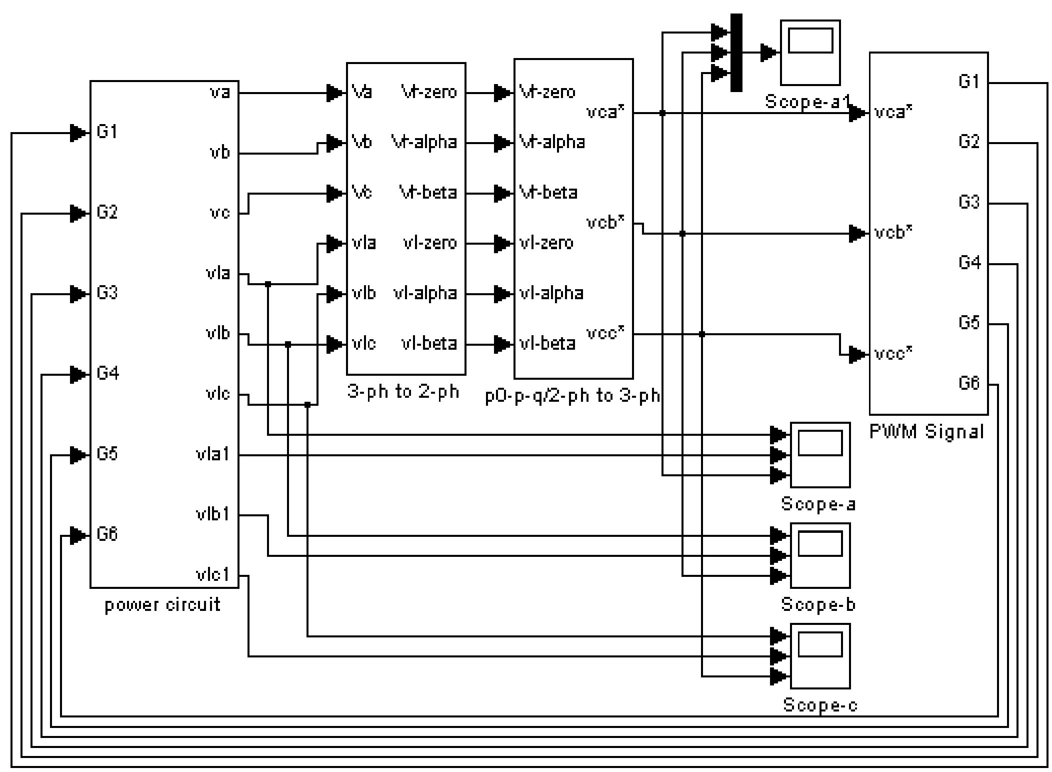


Figure 4.7.2.1-1: Block Diagram for Simulation

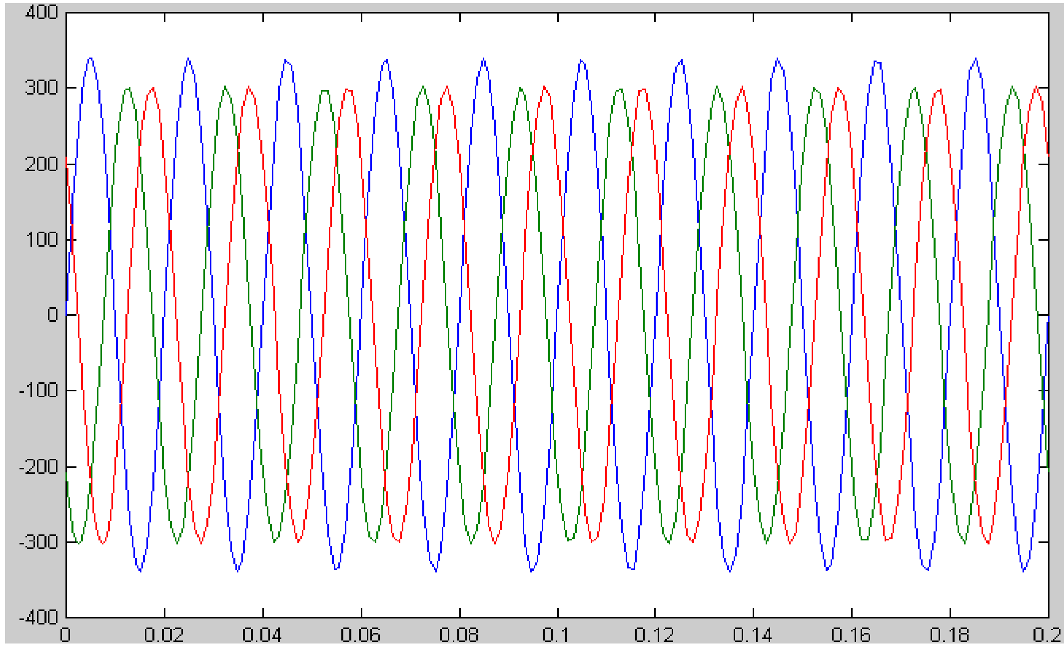


Figure 4.7.2.1-2(a)**:** Unbalanced Input Voltages

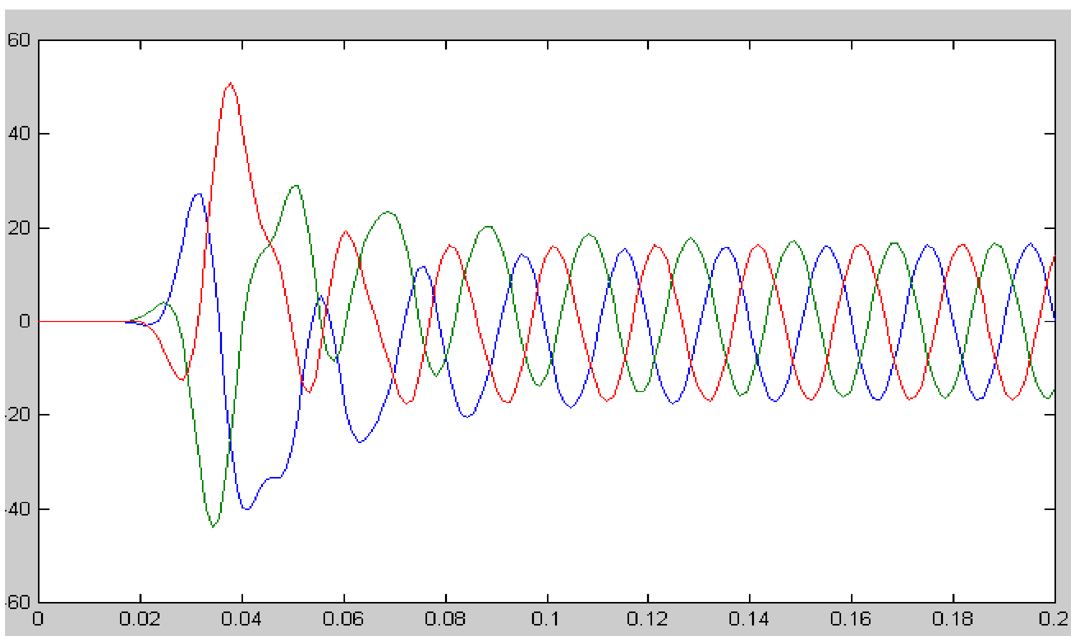


Figure 4.7.2.1-2(b)**:** Compensating Reference Voltages Without Accounting Zero Sequence Components.

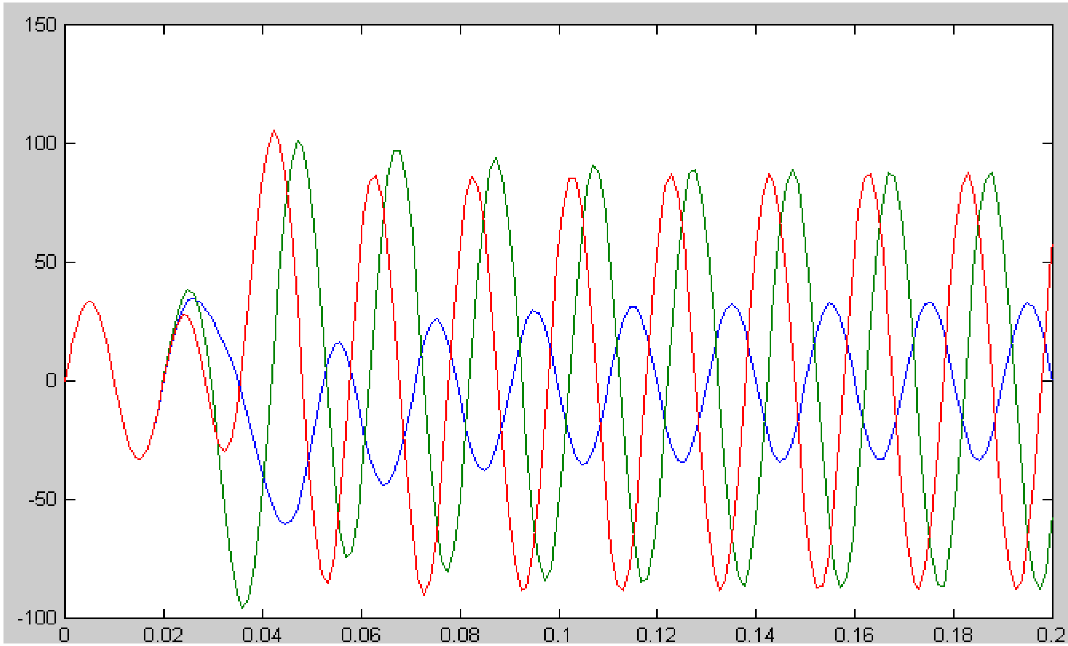


Figure 4.7.2.1-2(c): Compensating Reference Voltages With Accounting Zero Sequence Components

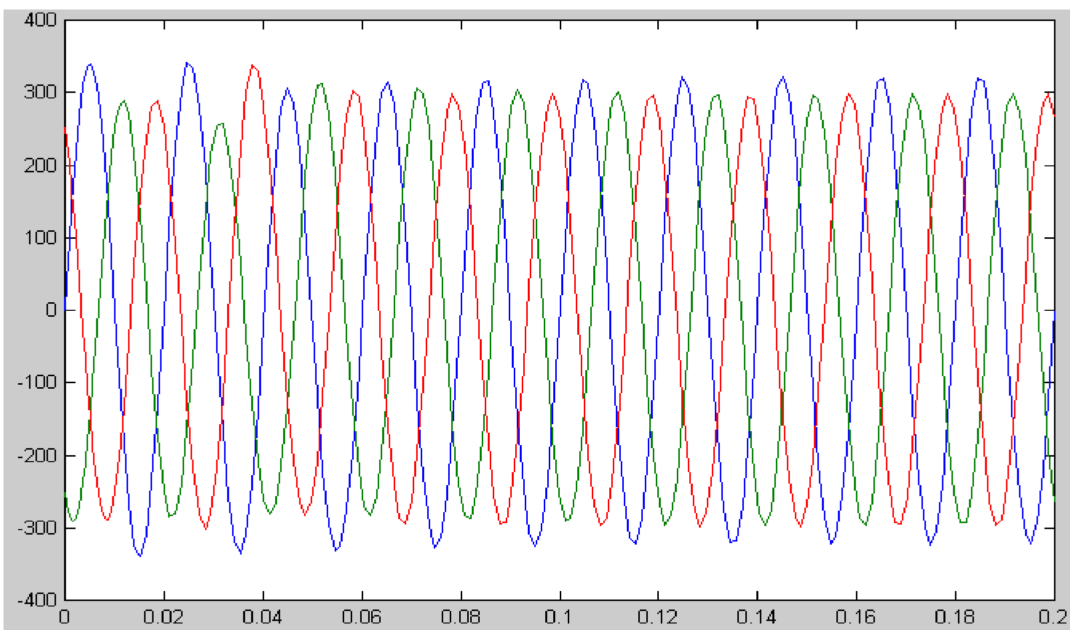


Figure 4.7.2.1-2(d): Load Voltages After Compensation Without Accounting Zero Sequence Components

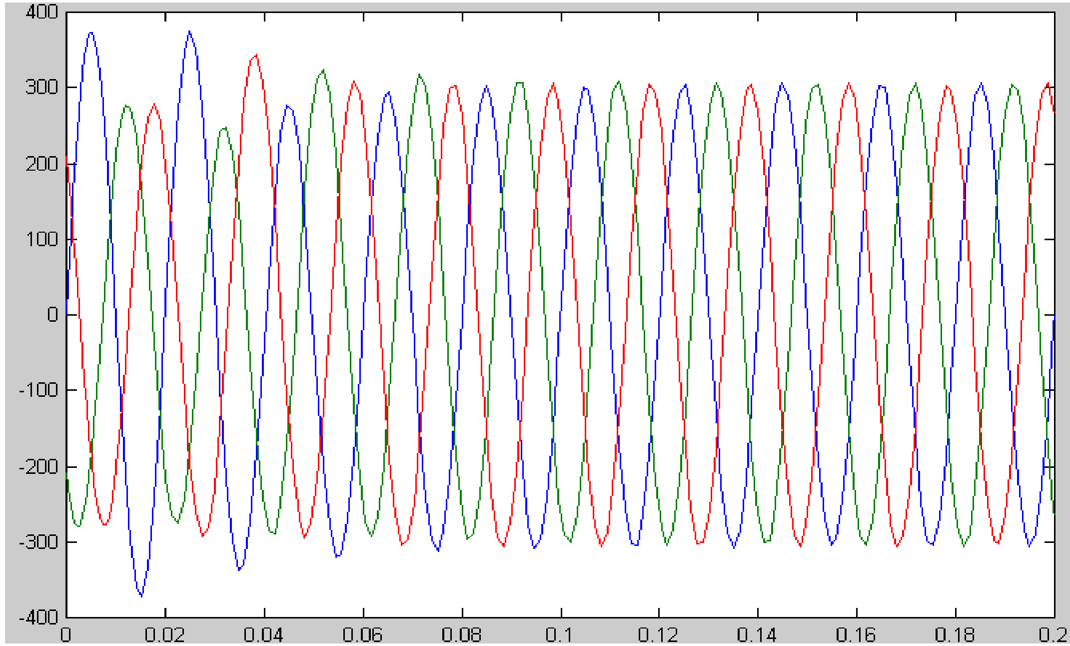


Figure 4.7.2.1-2(e): Load Voltages After Compensation With Accounting Zero Sequence Components

4.7.2.2 EXPERIMENTAL SETUP & WAVEFORMS:

The experimentation set up is prepared for detecting and compensating the negative sequence voltage in unbalance supply voltage. Schematic of experimental set up is given in Figure 4.7.2.2-1.

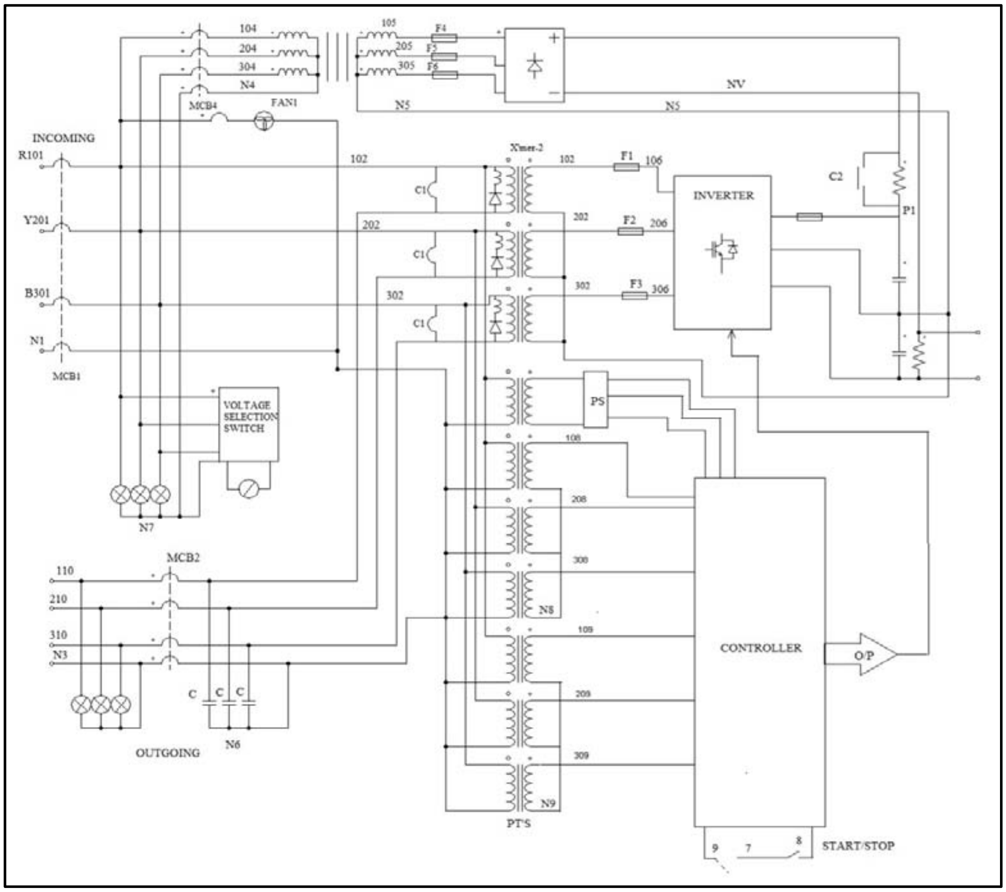


Figure 4.7.2.2-1: Schematic Diagram of Experimental Setup

The experiments set up were design for 12 amp 415 volt input supply voltage. The power circuit, as shown in figure, mainly consists of Inverter Bridge, high frequency transformer for injecting the voltage in between supply and load, LC filter. For bridge inverter, the IGBT used are SEMIKRON make SKM50 GB123 (50 Amps, 1200 V, dual module, with anti-parallel diode) for three phase application three such IGBT dual module were used. For DC bus capacitor C1, C2 = 6900 μF (4700 μF, 450 VDC Capacitor two in series and such three branch connected in parallel ALCON make with center point of each branch common) are used.

The high frequency transformer plays important role in the series compensation. The secondary of this transformer is connected in series between the source and the linear load. Rating of the transformer is decided based on the maximum unbalance to be compensated by negative sequence compensator. Considering this as based the rating of the high frequency transformer [J10] are fixed as follows:

Voltage ratings	:0-110-150//0-110-150volts per phase
Current rating	:15 amps per phase
Configuration	: three phase
VA ratings	:2250 VA per phase
Insulation Level	:3 kV
Leakage reactance	:10%.
Pri Turns	:120
Sec Turns	:132
Pri	: 8 Sq. mm
Sec	: 8 Sq. mm
Flux density	: 1.45 T,
Current density	: 1.87 A / mm^2

The high frequency isolation transformer is designed for high leakage reactance to filter out the effect of switching frequency. The frequency of switching is kept 4.8 kHz. The PWM signals (generated as a result of comparison of triangular wave with reference compensating voltages) are fed as gate signals to the IGBTs through the driver circuit. The Driver card is design to drive the IGBT using Mitsubishi hybrid gate Driver IC = M57962L so that it can be placed near the IGBT gate to reduce the parasitic impedance & to minimize the delay. Thus the output of the inverter contains the compensating fundamental voltage components plus the switching frequency. As the switching frequency component is not required, it is filtered out using the tuned LC filter. The value of the capacitor used in LC filter at the output of the negative sequence compensator as shown in the Figure are 0.47 μF, 2 kV.

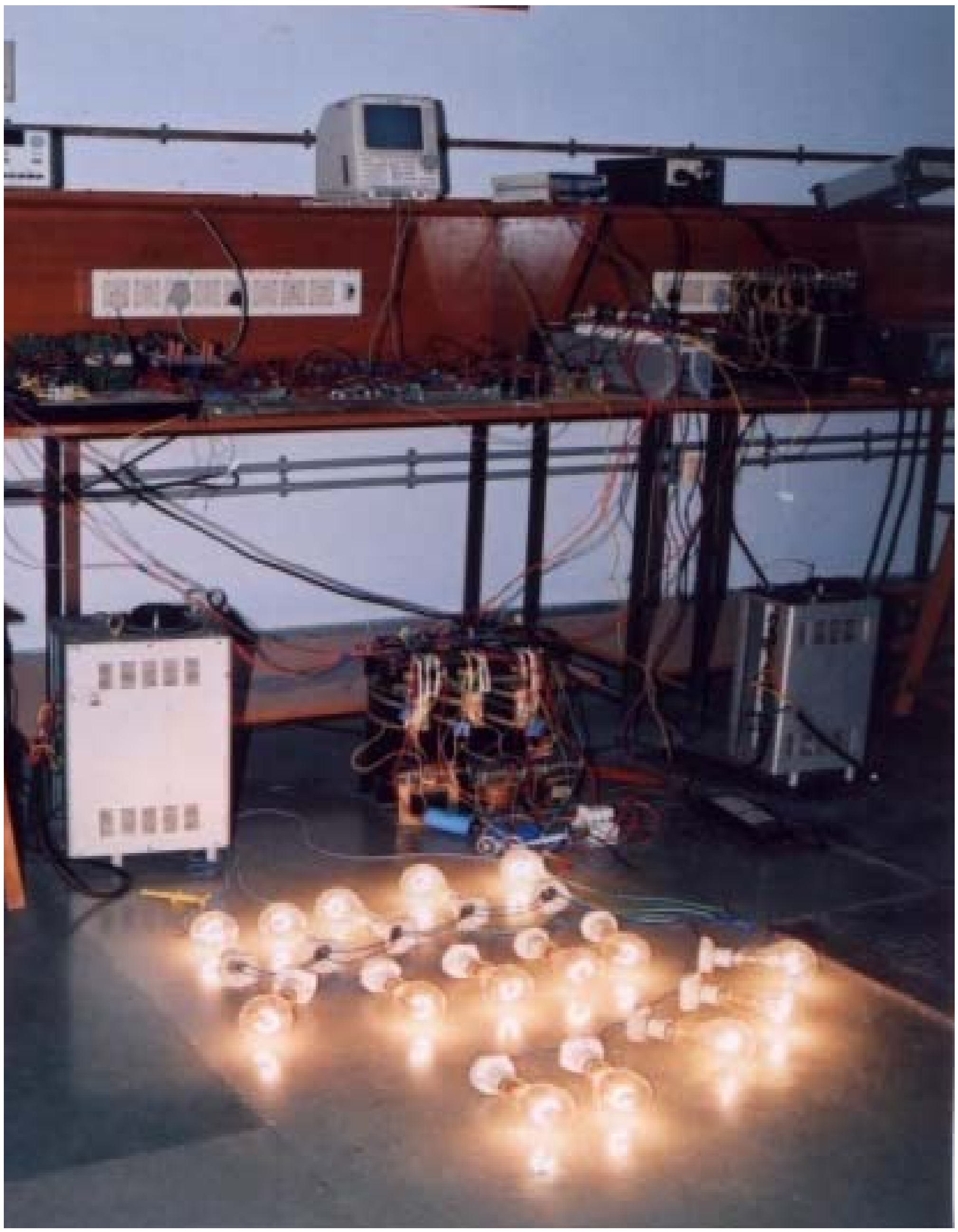


Figure 4.7.2.2-2: Experimental Set-Up

The technique explained in this paper is implemented using an analog circuit. The multiplication and division of the signal is done using low cost analog multiplier AD633 which is LASER trimmed 10 V scaling reference. Unbalance voltage is generated through three single-phase variacs. The unbalance voltages are sensed through PTs and negative sequence reference voltages are derived. The experimentation setup with control using analog IC, power circuit using IGBT, high frequency transformer, capacitor and lamp load is shown in Figure 4.7.2.2-2 and the experimental wave form are shown in the Figure 4.7.2.2-3 to Figure 4.7.2.2-10

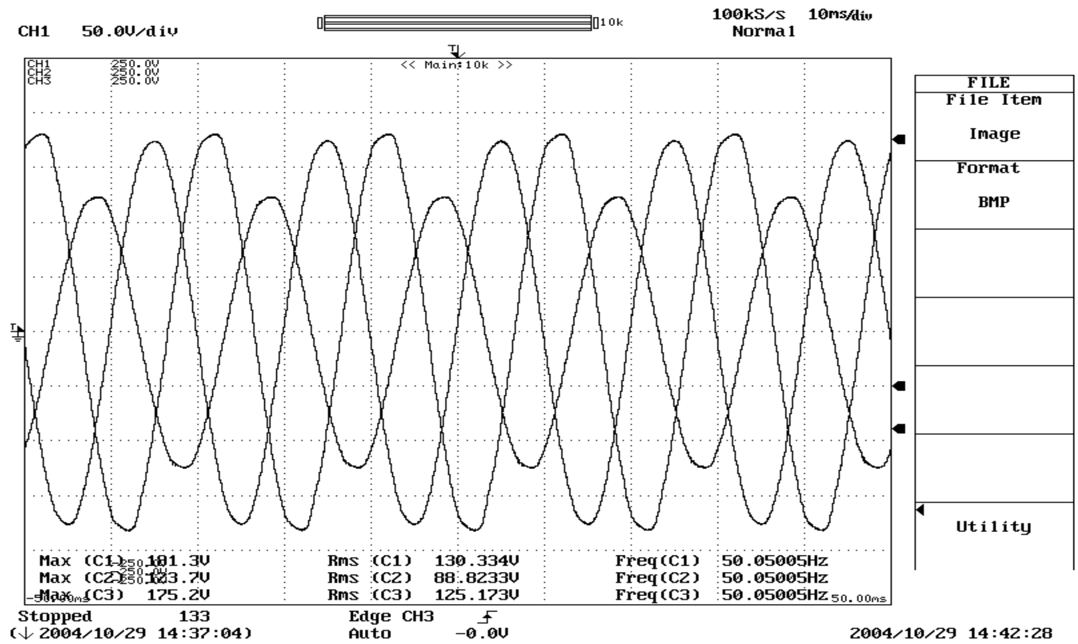


Figure 4.7.2.2-3: Unbalance Input Voltages

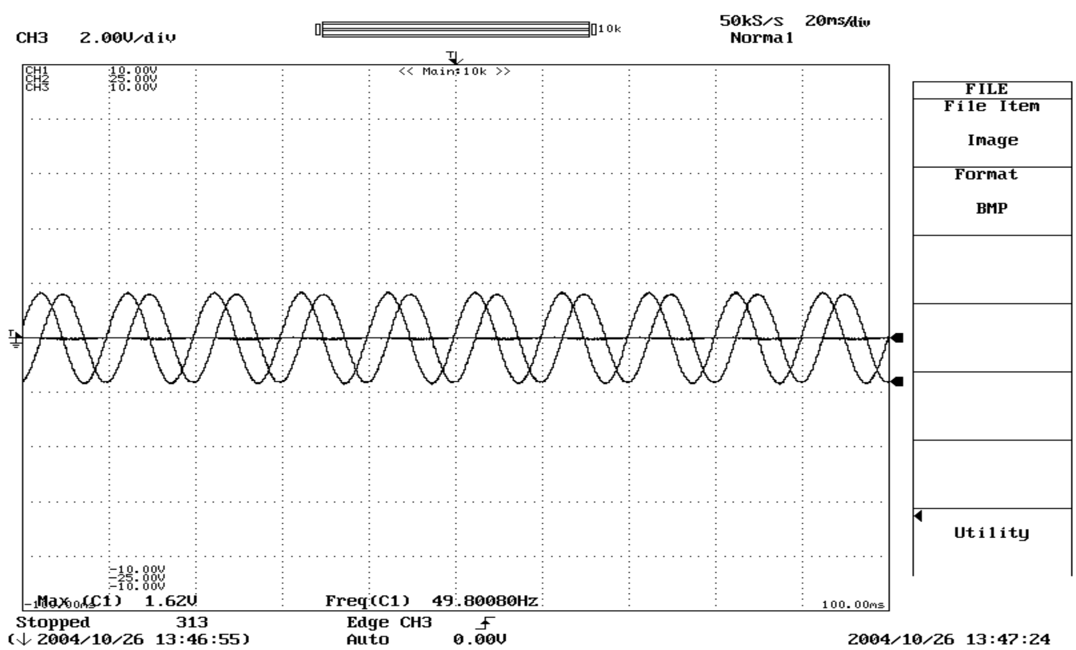


Figure 4.7.2.2-4: $V_{\alpha R}$ & $V_{\beta R}$ for Reference Input Voltages

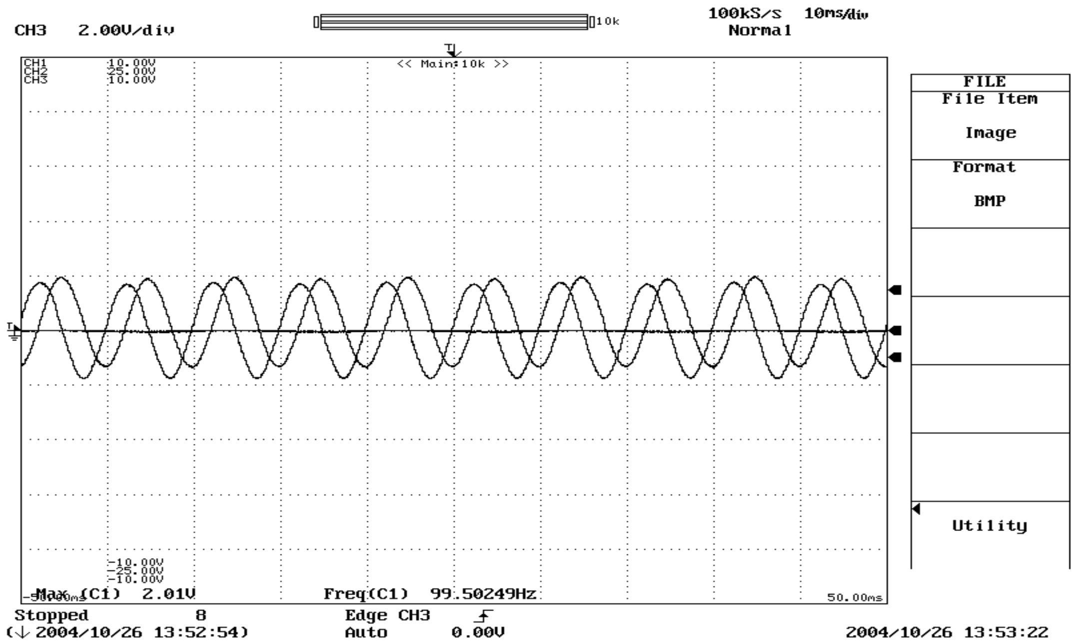


Figure 4.7.2.2-5: V_{α} & V_{β} for Input Voltages to Be Sensed

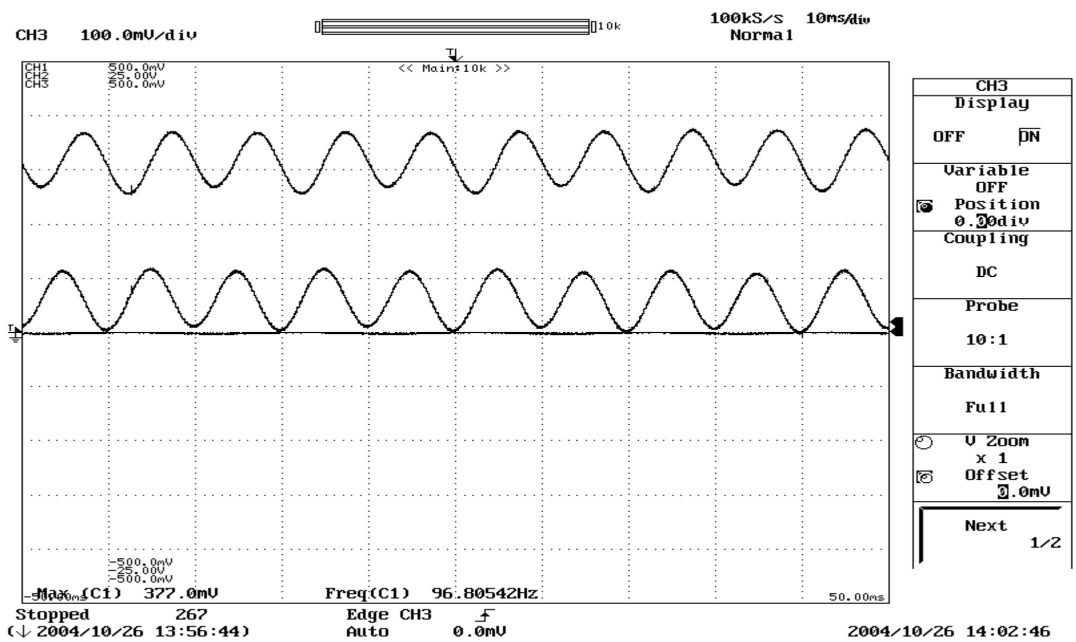


Figure 4.7.2.2-6: Active Power p & Reactive Power q In α-β Reference Frame

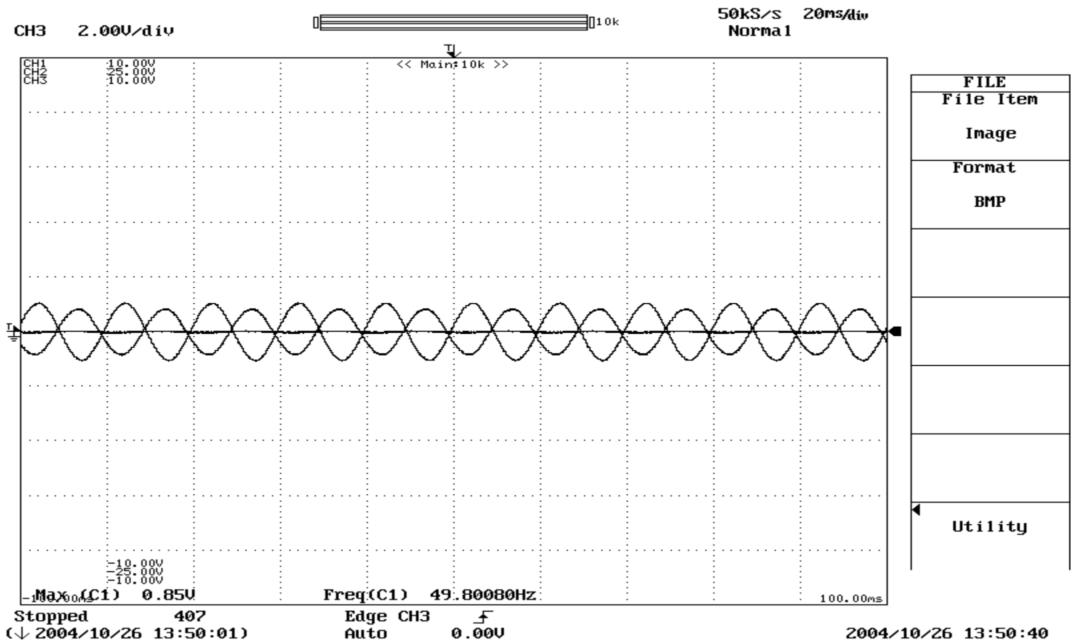


Figure 4.7.2.2-7: p_{ac} & q_{ac} in α-β Reference Frame after Filtration

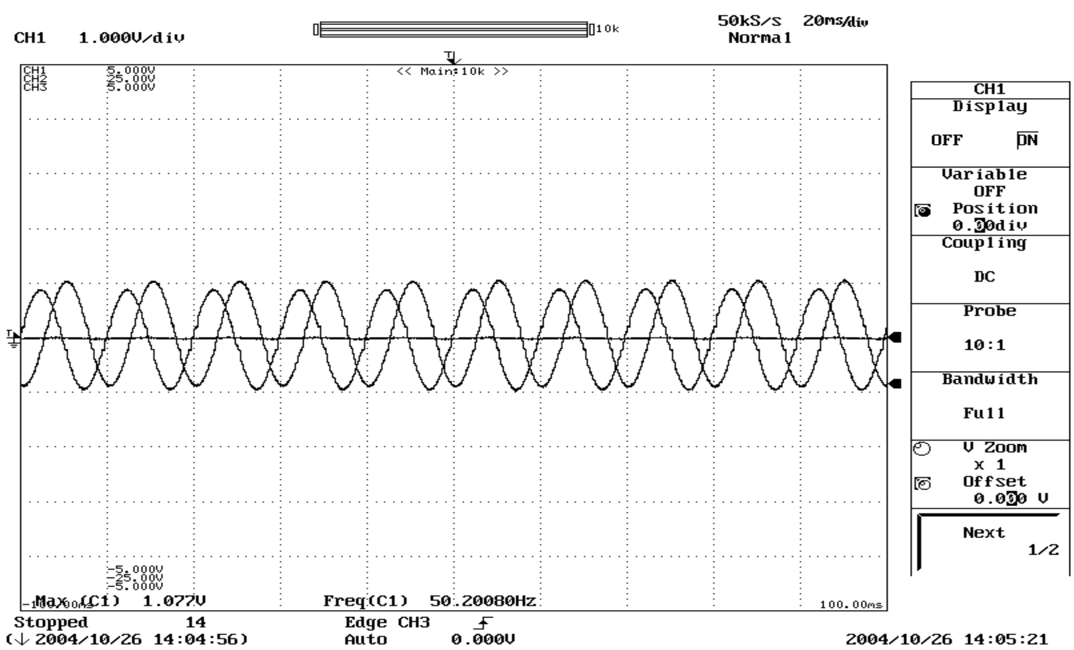


Figure 4.7.2.2-8: $V_{\alpha n}$ & $V_{\beta n}$ in α-β Reference Frame after Filtration

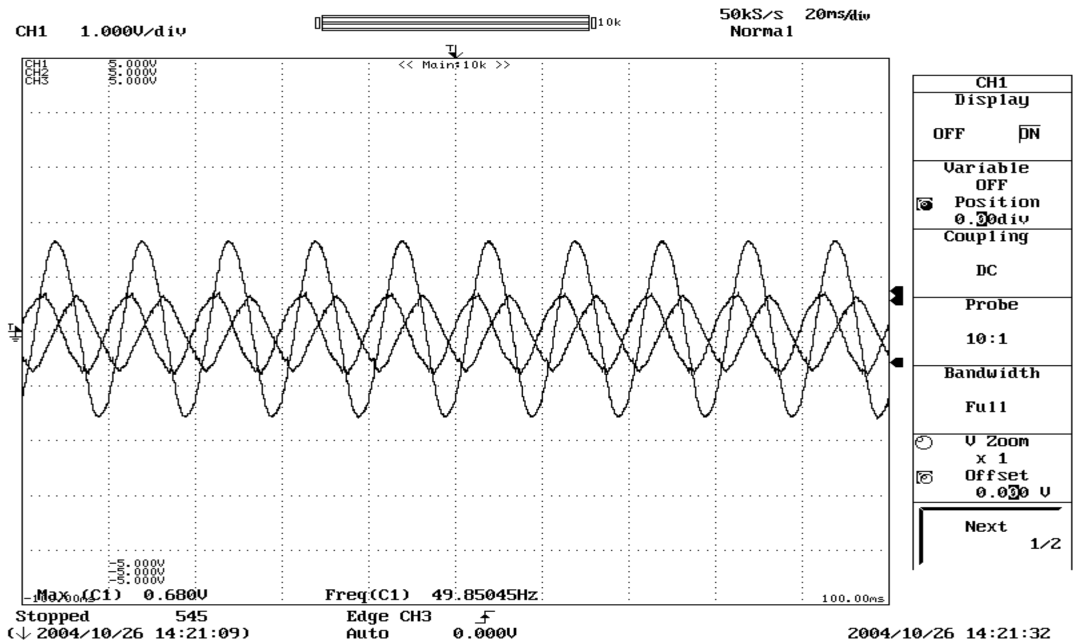


Figure 4.7.2.2-9: Compensating Reference Voltages

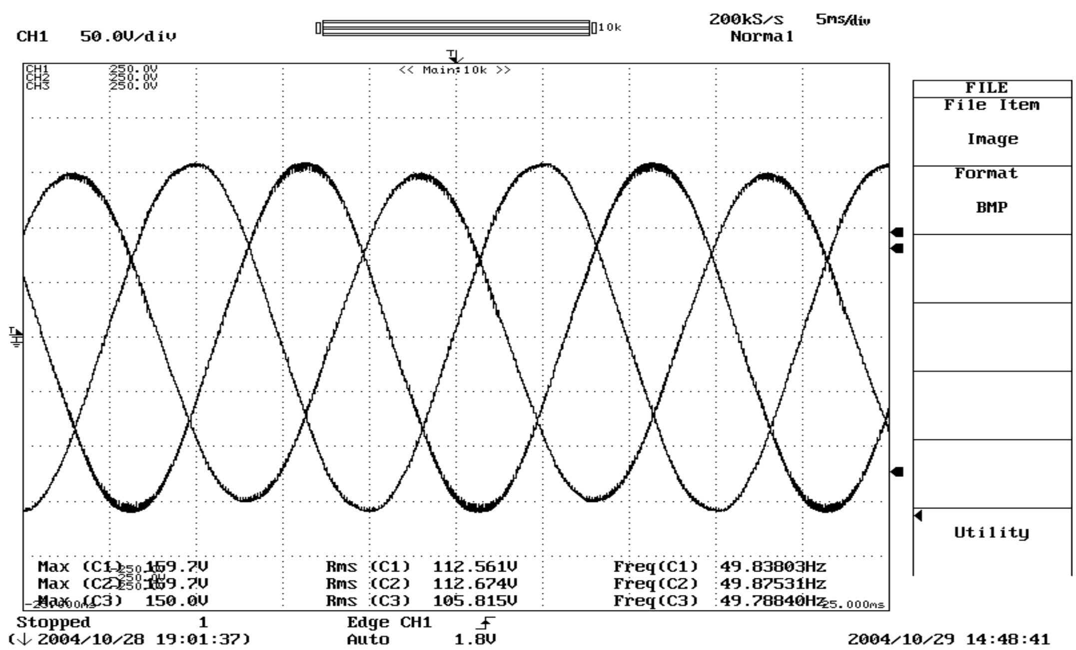


Figure 4.7.2.2-10: Three Phase Balance Load Voltage after Compensation

4.8 CONCLUSION:

Simulation results as well as experimentation results show that this new technique detect online negative & zero sequence component and compensate the same using voltage source inverter whose output changes through PWM techniques. Main advantage of detecting the negative & zero sequence component using instantaneous α-β-0 theory is that this can be easily implemented using DSP or analog circuit and no complex algebra is required which is the case using symmetrical components even with instantaneous symmetrical components as discussed in section 4.4.3 of this chapter. Also online instantaneous negative & zero sequence parameter of the bus voltage can be detected using this method. Negative sequence compensation is shown to be desirable for improved performance of 3-ϕ induction motor. The losses reduces due to compensation of negative sequence & zero sequence for 10 kW, 3 phase, 415 Volts, 50 Hz induction motor are given in following table:

<table>
<tr><th></th><th>Input power in Watts</th><th>output Power in Watts</th><th>Total Losses in Watts</th><th>% Effi. of motor</th><th>Total Saving in Watts</th><th>% saving of input power</th></tr>
<tr><td>with balance voltage</td><td>11312</td><td>9822</td><td>1038</td><td>87</td><td rowspan="2">2785</td><td rowspan="2">28%</td></tr>
<tr><td>with Unbalance voltage</td><td>10986</td><td>6858</td><td>3823</td><td>62</td></tr>
</table>

CHAPTER-V

COMBINED NEGATIVE SEQUENCE, ZERO SEQUENCE & HARMONICS SERIES COMPENSATOR

TABLE OF CONTENTS

5.1 INTRODUCTION TO CHAPTER

This chapter presents a novel and online method, which is also a novel contribution of this research work for detecting negative, zero sequence and harmonic compensation in voltages. It is based on instantaneous Active Reactive Power Theory. Using this concept, negative sequence, zero sequence & harmonics in voltage are computed online. The theory is verified through simulation results.

5.2 INTRODUCTION

In chapter-III series active filter was discussed and in chapter-IV negative & zero sequence compensator was discussed. In both the cases, to detect online & instantaneous voltage harmonic as well as negative sequence voltage harmonics Instantaneous Active Reactive Power Theory was used. After extracting the desired parameter (either voltage harmonic or negative sequence & zero sequence) voltage source inverter was controlled using above parameter so that desired compensation will be achieved. In simple word, the extracted parameter is used as reference along with PI control to control the voltage source inverter to achieve the desired compensation.

Extraction of parameter to be compensated can be done with different method as discussed in chapter-II i.e. IARP Theory, Synchronous Reference Frame Theory, and Sine Multiplication Theory. In chapter-III "Series Active Filter-Solution to Voltage Harmonics & Distortion and chapter-IV "Negative Sequence & Zero Sequence Compensator", Instantaneous Active & Reactive Power Theory is used for extraction of voltage harmonics and for extraction of negative sequence & zero sequence voltage.

In real application, both voltage harmonics and unbalance in voltage are present. The unbalance & harmonics voltage are present in the bus because of the unbalance and/or single phase, nonlinear having high content of harmonics load present in the same bus. This phenomenon is explained in detail in Chapter IV. In such cases these equipment may not compensate for both parameters. In such cases it is desirable to use a control scheme which can extract both the parameter online instantaneous and compensate for the same.

In this chapter, a novel control method with slight modification was introduced which can extract harmonics as well as negative & zero sequence parameter using instantaneous active reactive power theory. Hence extracted signal will be used as reference signals along with PI control to control the voltage source inverter. So that harmonics as well as negative & zero sequence voltages in the sensitive bus will be reduced

5.3 CONTROLLING OF NEGATIVE SEQUENCE COMPENSATOR

To control voltage source inverter and to generate voltages for compensating harmonics, negative & zero sequences a reference signal is required. This reference signal will control the Voltage Source Converter (VSC).

For extracting harmonics, negative sequence and zero sequence from the load bus voltage a novel technique using instantaneous active reactive power theory is used. When signal (reference and actual) are transformed into α-β-0 reference frame, the signals are converted into two phase equivalent. If both reference signal as well as actual signal in α-β-0 reference frame are put through cross & dot products the output leads to all information required to improve the power quality. By selecting proper filter, order of filter and its bandwidth desired signal can be extracted from actual signal. This signal will be used as reference signal to control VSC for desired output

5.4 GENERATION OF REFERENCE SIGNALS FOR COMBINED HARMONICS, NEGATIVE SEQUENCE & ZERO SEQUENCE VOLTAGE

To implement this theory, digitally generates reference unit rms amplitude three phase positive sequence voltages in synchronization with supply voltage v_a, v_b and v_c as follows.

$$v_{ar} = \sqrt{2}\ \sin(\omega t)$$

$$v_{br} = \sqrt{2}\ \sin\left(\omega t - {}^{2\pi}/_{3}\right)$$

$$v_{cr} = \sqrt{2}\ \sin\left(\omega t + {}^{2\pi}/_{3}\right) \tag{4.4-1}$$

Convert these reference voltages to α-β-0 reference [J16]-[J19] frame through the transformation matrix as shown below.

$$\begin{bmatrix} v_{0r} \\ v_{\alpha r} \\ v_{\beta r} \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{ar} \\ v_{br} \\ v_{cr} \end{bmatrix} \tag{4.4-2}$$

Let v_a, v_b and v_c be the phase to neutral unbalanced & distorted three phase voltages of the input power supply. These voltages are sensed through voltage sensor and converted into α-β-0 component through the transformation matrix as below.

$$\begin{bmatrix} v_{0} \\ v_{\alpha} \\ v_{\beta} \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} v_{a} \\ v_{b} \\ v_{c} \end{bmatrix} \tag{4.4-3}$$

Now the following calculation is performed using these α-β-0 components.

$$\begin{bmatrix} p_0 \\ p \\ q \end{bmatrix} = \begin{bmatrix} v_{0r} & 0 & 0 \\ 0 & v_{\alpha r} & v_{\beta r} \\ 0 & -v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} v_0 \\ v_\alpha \\ v_\beta \end{bmatrix} \tag{4.4-4}$$

p_o is zero sequence power and p & q can have dc and ac components as function of time.

In the instantaneous p-q theory DC components in p & q are responsible for the fundamental positive sequence or negative sequence power & this depends upon the reference voltage signal. If reference voltage v_{ra}, v_{rb}, v_{rc} are the positive sequence then DC components in the power p & q reflects the positive sequence power otherwise if reference voltage v_{ra}, v_{rb}, v_{rc} are of negative sequence, then DC components in p & q reflects the fundamental negative sequence power.

In Equation (4.4-4), the DC components are due to positive sequence voltages in the bus.

Similarly p_{ac} and q_{ac} are responsible for other than active power components i.e. harmonics, reactive and unbalance in the systems

AC components are filtered out using high pass filter to get the p_{ac} and $q_{ac.}$ The parameters defined for high pass filter are as follows:

Design method:	Elliptic
Filter type:	High pass
Filter Order:	2
Passband edge frequency :	6.14 rads/sec:
Passband ripple in dB:	0.1

Stopband ripple in dB: 60

From p_{ac} and q_{ac} compensating reference signal are computed [J15] using the following transformation.

$$\begin{bmatrix} v_{\alpha c} \\ v_{\beta c} \end{bmatrix} = \begin{bmatrix} v_{\alpha r} & v_{\beta r} \\ -v_{\beta r} & v_{\alpha r} \end{bmatrix}^{-1} \begin{bmatrix} -p_{ac} \\ -q_{ac} \end{bmatrix}$$

$$\begin{bmatrix} v_{\alpha c} \\ v_{\beta c} \end{bmatrix} = \frac{1}{(v_{\alpha r}{}^2 + v_{\beta r}{}^2)} \begin{bmatrix} v_{\alpha r} & -v_{\beta r} \\ v_{\beta r} & v_{\alpha r} \end{bmatrix} \begin{bmatrix} -p_{ac} \\ -q_{ac} \end{bmatrix} \tag{3.5.1-6}$$

$v_{\alpha c}$ and $v_{\beta c}$ are the α, β components of compensating voltage of the supply, which can be transformed back to three-reference voltages for PWM voltage source inverter.

$$\begin{bmatrix} v_{ac} \\ v_{bc} \\ v_{cc} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1 & 0 \\ 1/\sqrt{2} & -1/2 & \sqrt{3}/2 \\ 1/\sqrt{2} & -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} -v_0 \\ v_{\alpha c} \\ v_{\beta c} \end{bmatrix} \tag{3.5.1-7}$$

Where v_o is the zero sequence voltage component of the three phase bus voltages.

In the above manner, the instantaneous values of compensating voltages (including harmonic, negative sequence & zero sequence components) of phase 'a' phase 'b' and phase 'c' are computed.

When once these three reference signals are obtained they are fed to the voltage source inverter as reference signals to generate corresponding voltages at the output of voltage source inverter which will completely eliminate the harmonics, negative sequence and zero sequence component voltages in the mains supply and will not affect its positive sequence component voltages

Further there will be no additional braking torque on the rotor due to the positive sequence currents flowing through negative sequence compensator, as was the case with the other compensator. So, the mechanical output will not be reduced. The active energy input to harmonics, negative & zero sequence compensator is reduced Only very little amount of power loss takes place in the inverter and transformer. Further more positive sequence voltage applied to the motor can be improved to its rated voltage by adding a portion of the fundamental positive sequence mains voltage to the reference signals of the harmonics, negative & zero sequence compensator schemes.

5.5 SIMULATION

The simulation studies are carried out to predict the performance of the proposed new theory using MATLAB-SIMULINK-POWER SYSTEM BLOCK. For simulation of voltage harmonics and negative & zero sequences compensator a voltage source inverter whose output is changed through sinusoidal pulse width modulated (PWM) voltage source inverter. PWM method utilizes a reference signal (one for each phase), which is compared with a high frequency triangular wave, and the points of intersection of both these signals decide the time of switching on and off the devices connected to that particular phase. Block diagram for simulation is shown in Figure 4.5-1 to Figure 4.5-4.

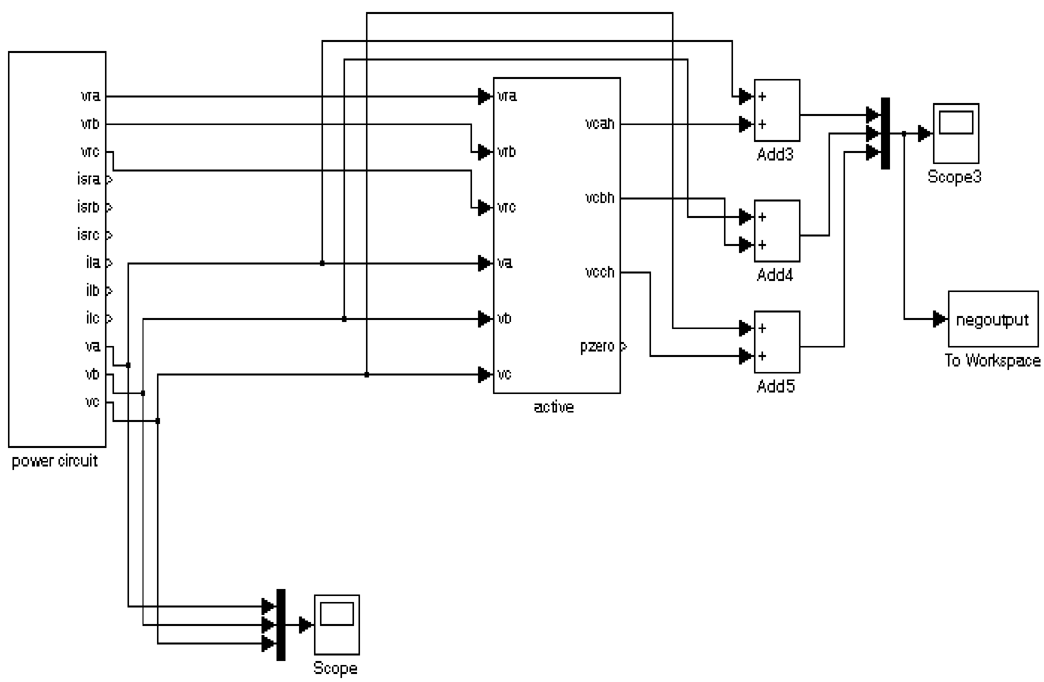


Figure 4.5-1-A: Block Diagram for Simulation

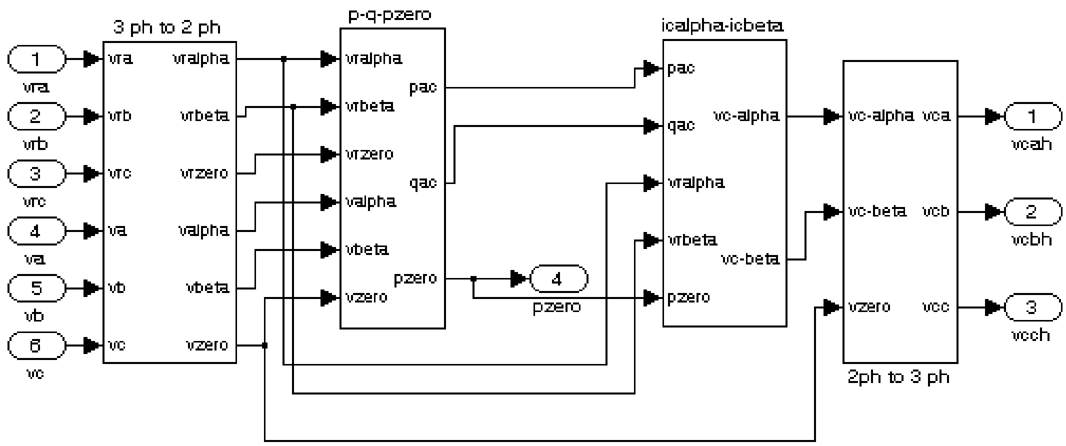


Figure 4.5-1-B: Control Block Diagram for Simulation

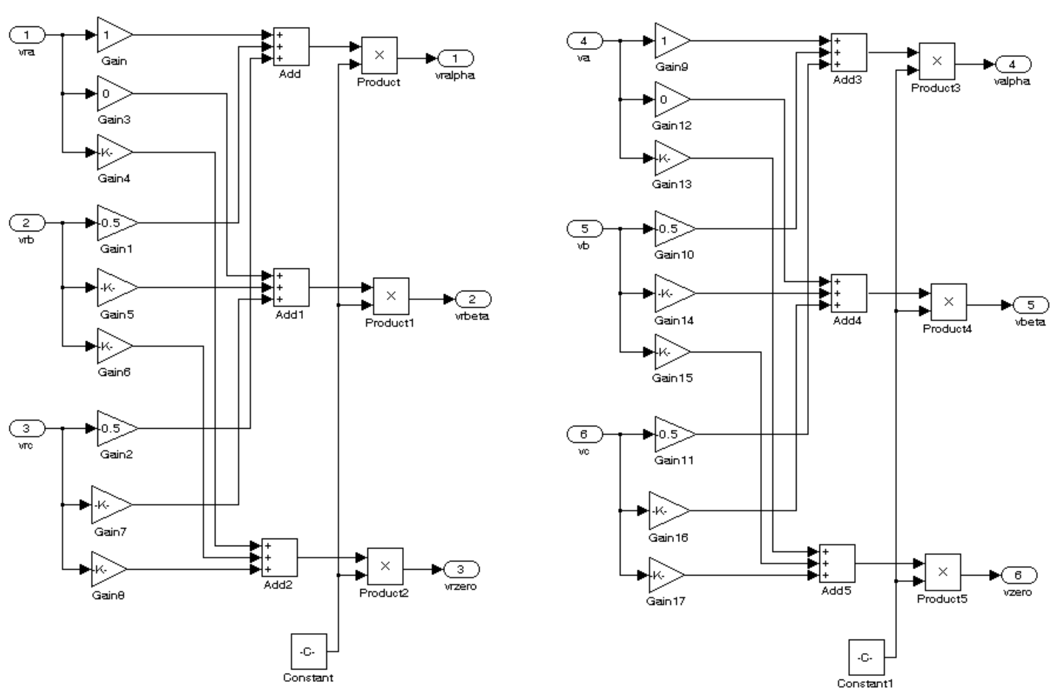


Figure 4.5-2: a-b-c to α-β-0 frame

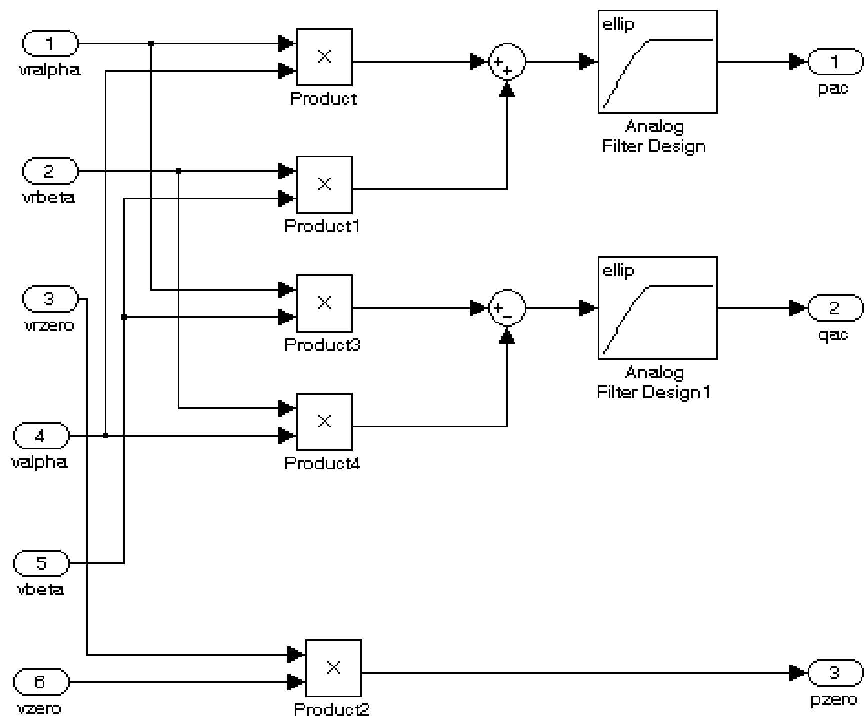


Figure 4.5-3: Cross Product & Dot Product For p_{ac} & q_{ac} of The Signal In α-β-0 Frame

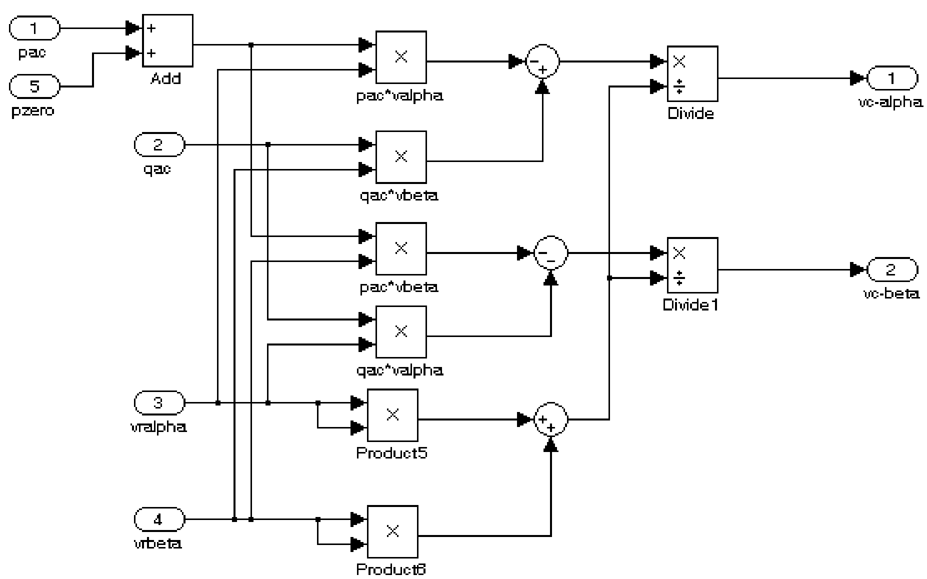


Figure 4.5-4: Compensating Reference Voltage Signal $V_{\alpha C}$ & $V_{\beta C}$ In α-β-0 Frame

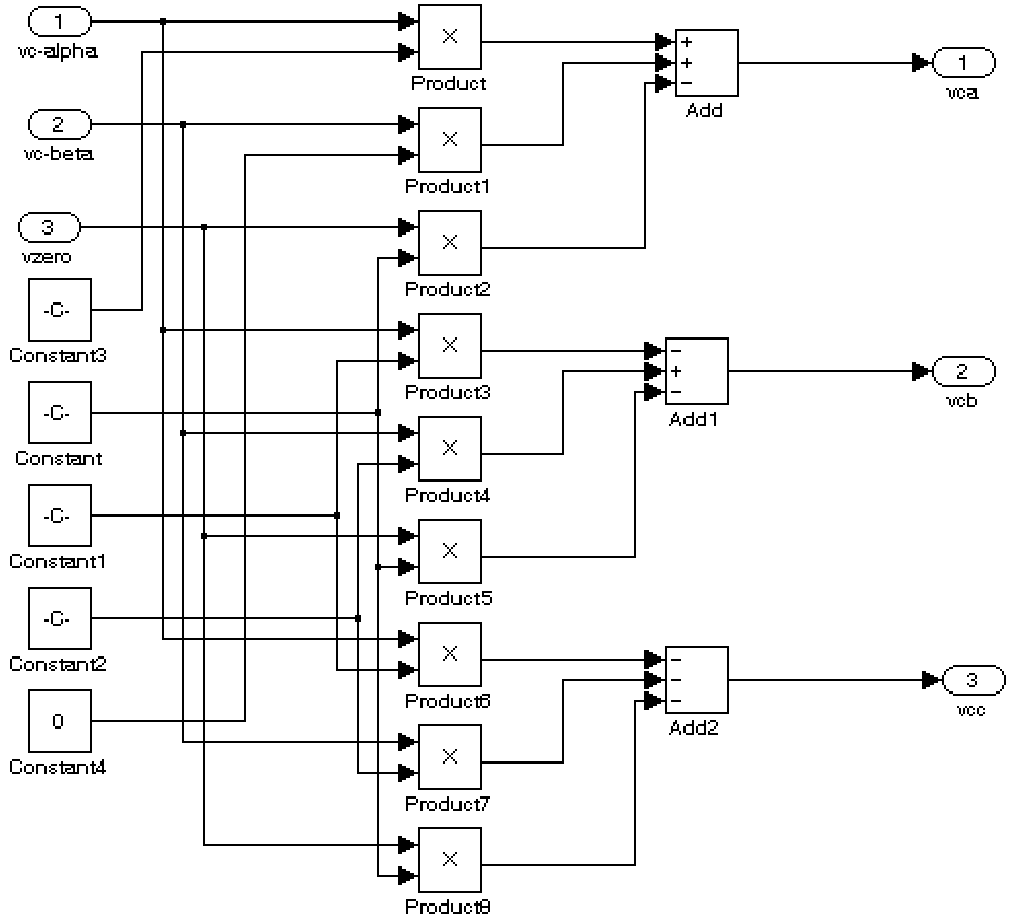


Figure 4.5-4: Compensating Reference Voltage Signal V_{ac}-V_{bc}-V_{cc} In a-b-c Frame

5.6 SIMULATION RESULTS & WAVEFORMS

The simulation results for above method are shown in Figure 4.6-1 to Figure 4.6-6.

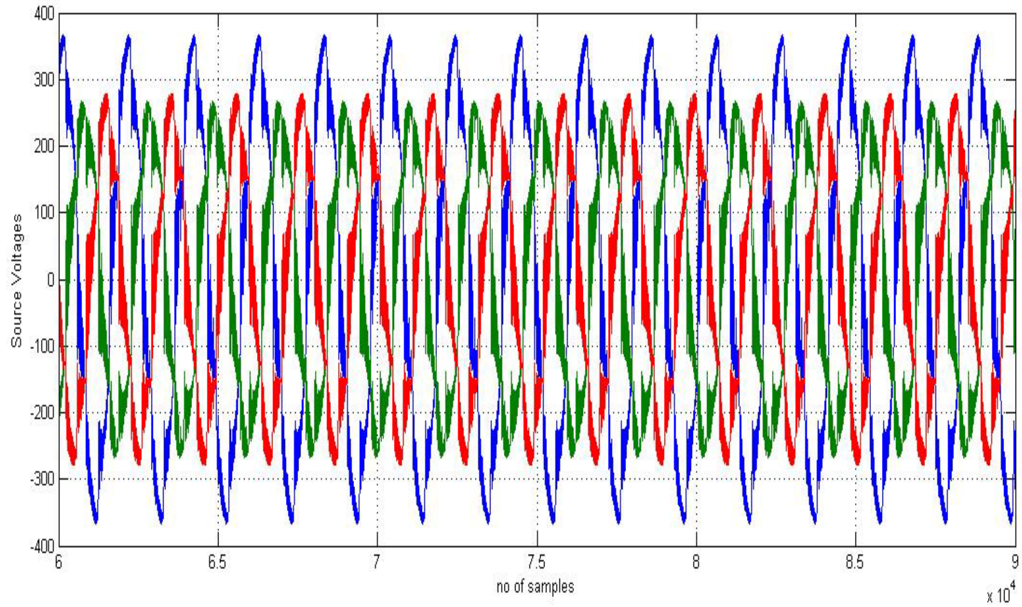


Figure 4.6-1: Input Side Source Voltage Having Distortion & Unbalance

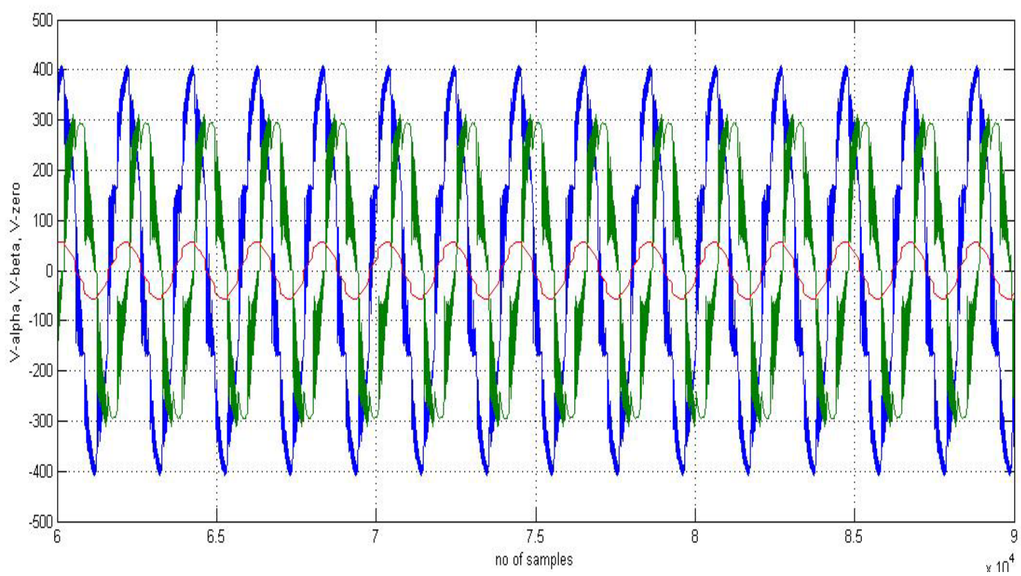


Figure 4.6-2: a-b-c to α-β-0 Conversion (Voltage Signal V_α & V_β in α-β-0 frame)

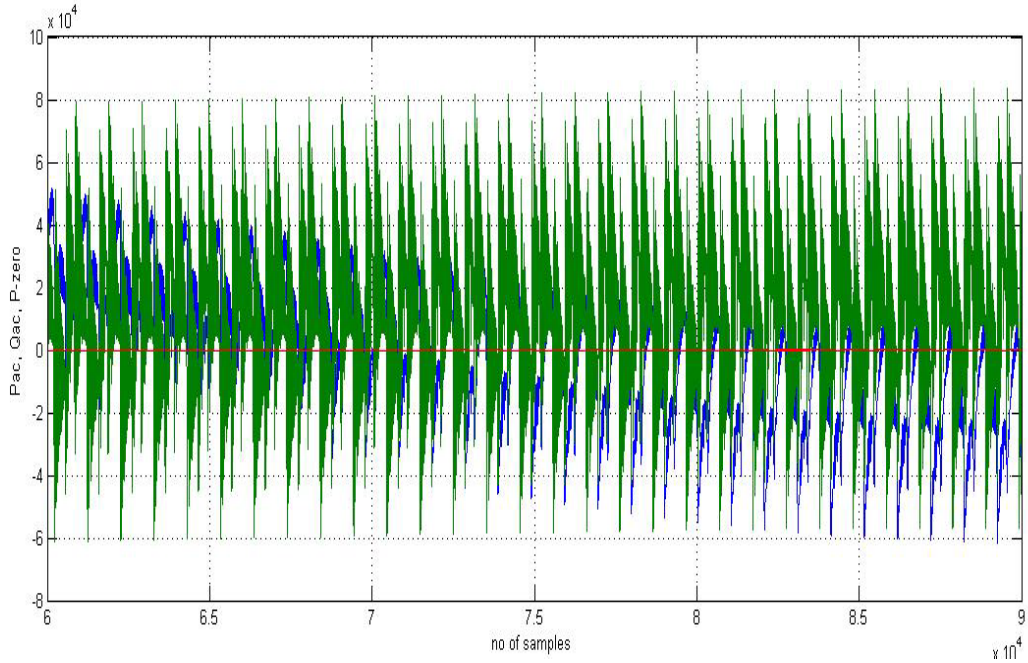


Figure 4.6-3: p_{ac} & q_{ac} of the signal in α-β-0 frame

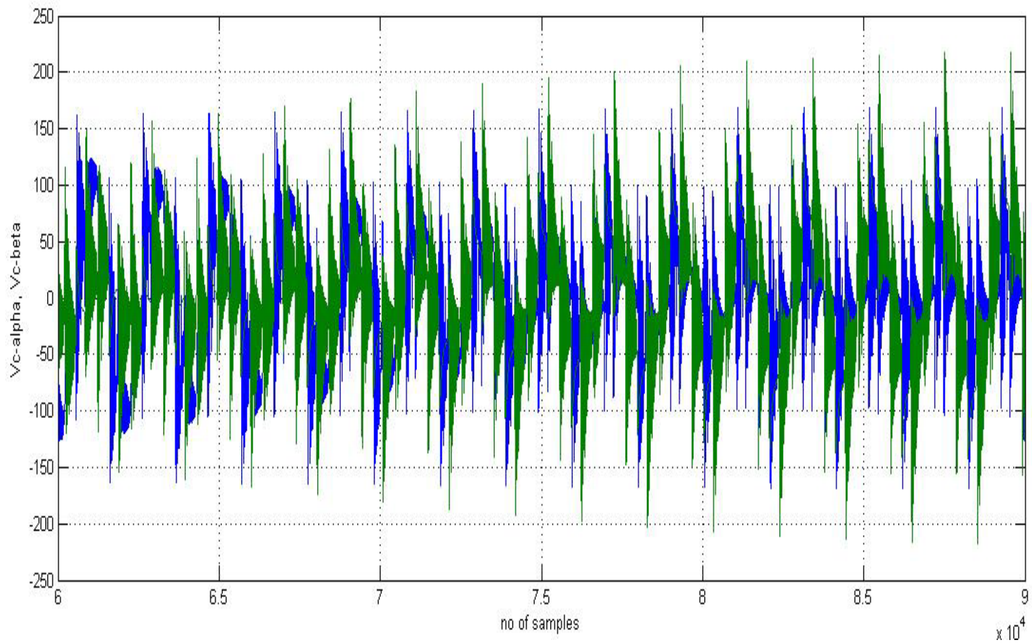


Figure 4.6-4: Compensating Reference Voltage Signal $V_{\alpha C}$ & $V_{\beta C}$ In α-β-0 Frame

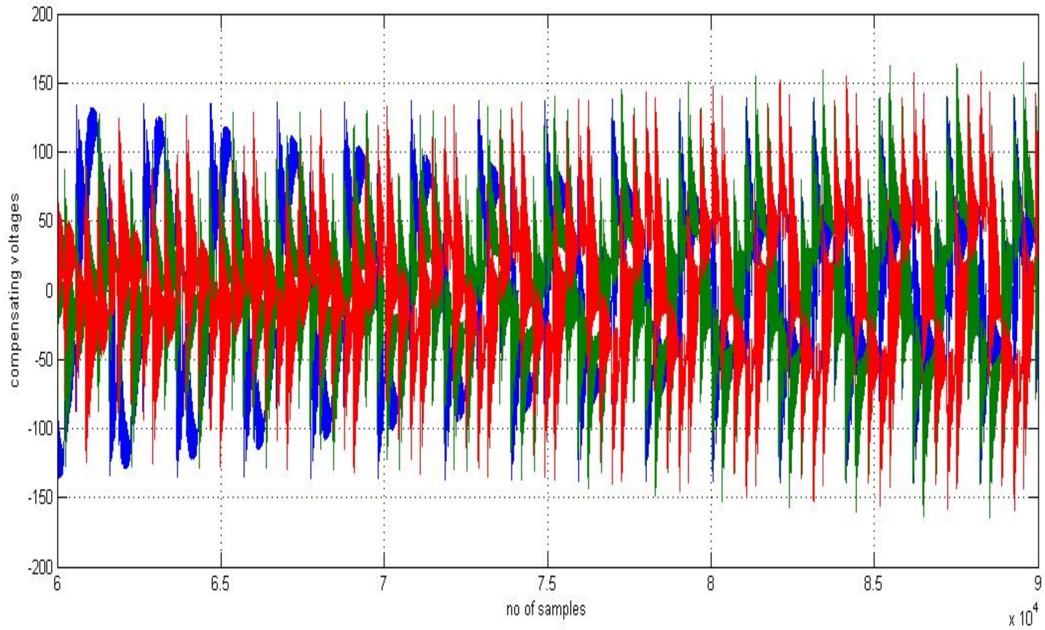


Figure 4.6-5: Compensating Voltage to Compensate Harmonics & Unbalance

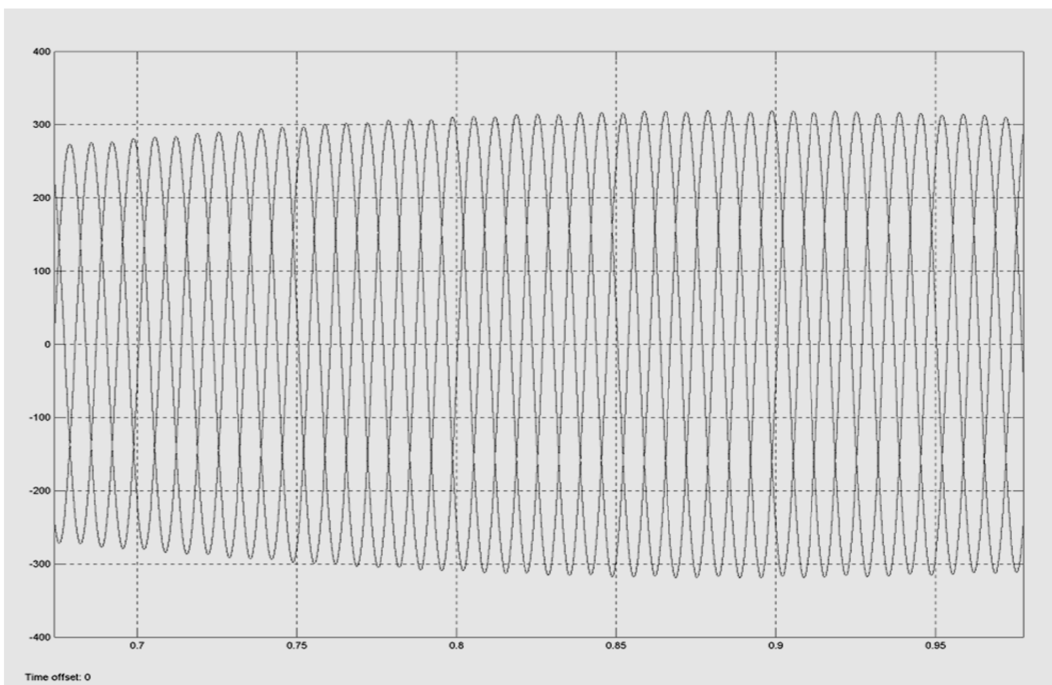


Figure 4.6-6: Load Bus Voltage After Compensation

5.7 CONCLUSION

Simulation results show that this new technique detect online harmonics, negative sequence & zero sequence components and compensate the same using voltage source inverter whose output changes through PWM techniques. Main advantage of detecting the negative sequence component using instantaneous α-β-0 theory is that this can be easily implemented using DSP or analog circuit and no complex algebra is required which is the case using symmetrical components. Implementing the same through DSP hardware is the future scope of work. Now a day's FAST DSP processors with built in ADC & DAC are available which can compute the signal online instantaneous without much delay.

CHAPTER-VI

CONCLUSION

6.1 CONCLUSION

This chapter consolidates the important findings and brief review of the work performed and reported in the thesis. Further the future scope / investigation in this area is indicated. This is followed by Reference used in the entire development work.

Power quality and reliability are essential for proper operation of industrial processes, which involve critical and sensitive loads. With the growing usage of solid-state converters in application such as adjustable speed drives and computer supplies, there has been a considerable increase in the voltage/current harmonic congestion and distortion on the power network. Short-duration power disturbances, such as voltage sags, swells and short interruptions, are major concern for industrial customer. Due to wide usage of sensitive electronic equipment in process automation, even voltage sags which last for only few tenths of second may cause production stops with considerable associated costs; these costs include production losses, equipment restarting, damaged or lower-quality product and customer satisfaction.

The high costs associated with these disturbances explain the increasing interest towards power quality & reliability problems mitigation techniques. The cost of mitigation intervention has to be compared with the loss of revenue and takes into account all economic factors involved.

Various power-conditioning devices such as series voltage compensator (e.g. DVR), shunt current compensator (e.g. STATCOM, SVC) and hybrid series and shunt filters are widely used to eliminate current harmonics and voltage distortion and thus improve the power quality.

Shunt passive filters (single tuned & double tuned filters) have traditionally been used to absorb the harmonics generated by large harmonics loads and provide the reactive supports below the tuned frequency. The supply and line impedance however strongly influence the compensation characteristics of the filter. Shunt filters are also highly susceptible to series & parallel resonances. Hence shunt filters are generally off-tuned with respects to dominant harmonics, which defeats their very purpose of installation.

Shunt active filters offer some improvement over passive filtering by way of smooth power factor and harmonic current control and non-susceptibility to resonance at dominant harmonic frequencies. But active current filtering is viable solution only if the peak harmonic distortion is limited. It is difficult to implement a large rated shunt PWM inverter with a high current bandwidth and a high switching frequency would elevate the switching losses.

Series compensators (e.g. DVR) have been effective voltage compensators. However it is difficult for them to control harmonics and zero sequence components in the absence of a passive shunt filter.

In this thesis novel techniques was develop using IARP theory to improve the power quality for sensitive load. Active series filter to eliminate the voltage harmonic for sensitive load was developed which reduced the voltage harmonic from approximately 17 % to 6%.

Similarly, using same IARP theory with slight modification it is possible to compensate negative as well as zero sequence components. Simulation results as well as experimentation results show that this new technique detect online negative & zero sequence component and compensate the same using voltage source inverter whose output changes through PWM techniques. Main advantage of detecting the negative & zero sequence component using instantaneous α-β-0 theory is that this can be easily implemented using DSP or analog circuit and no complex algebra is required which is the case using symmetrical components. Also online instantaneous negative & zero sequence parameter of the bus voltage can be detected using this method. Negative sequence compensation is shown to be desirable for improved performance of 3-ϕ induction motor.

Also novel control method was introduced which can extract harmonics as well as negative & zero sequence parameter using instantaneous active reactive power theory. Hence extracted signal will be used as reference signals along with PI control to control the voltage source inverter. These extracted signals are instantaneous & online so that harmonics as well as negative & zero sequence in the sensitive bus voltage will be reduced / compensated.

Following are the main benefits of these techniques

- This techniques is very simple & easy to implement
- On line compensation of power quality problem which lead to high productivity, less down time of machine and man power
- Losses in the power system components (i.e. transformer, transmission line, motor, cable etc) reduce.
- Improved Motor performance in terms of torque and efficiency
- Improve the performance of interconnected power system and the quality of the power.
- The techniques developed in this thesis covers to meets the following objectives:
- Compensation of voltage harmonics and negative & zero sequence voltage to the sensitive load bus.

A simple Control scheme to compensate both, voltage harmonics as well as negative & zero sequence voltage using IARP theory was also proposed along with simulation results.

With this it is clear that the hybrid topology (series active + shunt passive) has been shown to provide cost-effective solution to voltage distortion and current harmonics suppression. By extending the functionality of the hybrid compensator a wider range of power quality problems can be tackled.

REFERENCES

Reference A: Power Quality

[A1] Richard Redl, "Power Electronics and Electromagnetic Compatibility," Conference Record of IEEE-PESC 1996, pp. 15-21

[A2] Thomas S. Key and Jih-Sheng Lai, " IEEE and International Harmonic Standards Impacts on Power Electronic Equipment Design," Conference Record of IEEE-IECON 1997, pp. 430-436

[A3] Serge Bernard and Stephane Papoz, "Active Filter Design and Specification for Control of Harmonics in Industrial and Commercial Facilities," Conference Proceedings of Power Quality Solutions 1995, pp. 197-208

[A4] Tristan A. Kneschke, "Control of Utility System Unbalance Caused by Single-Phase Electric Traction," IEEE Transactions on Industry Application, Vol. IA-21, No. 6, Nov./Dec. 1985

[A5] Prem P. Khera, "Application of ZIGZAG Transformers for Reducing Harmonics in the Neutral Conductor of Low Voltage Distribution System," Conference Record of IEEE-IAS 1990, p.1092

[A6] J. E. Miller, "Reactive Power Control in Electrical Systems," New York: Wiley, 1982

[A7] A. Kusko and S. M. Peeran, "Voltage Distortion of Rotary UPS and Standby Engine Generator Sets," Conference Record of IEEE-IAS 1990, pp. 1757-1760

[A8] Karl Johnson and Robert Zavadil, "Assessing the Impacts of Nonlinear Loads on Power Quality in Commercial Buildings – An Overview," Conference Record of IEEEIAS 1991, pp. 1863-1869

[A9] IEEE-519, “IEEE Recommended Practices and Requirements for Harmonic Control in Electrical Power System,” 1992

[A10] IEC 1000-3-2, Electromagnetic Compatibility Part 3: Limits, Section 2: Limits for Harmonic Current Emissions (Equipment Input Current £ 16A per phase), Mar. 1995

[A11] IEC 1000-3-4, “Harmonic Limits for Connection of Equipment with Input Current Exceding 16A to the Public Low Voltage Supply System,” Mar. 1995

[A12] CBIP 251-1996 “ Guide For Limiting Voltage Harmonics ”. ISBN: 81-7336-264-5, CBIP Central Board of Irrigation & Power, Malcha Marg, Chanakyapuri, New Delhi-110021

[A13] NEMA standard MG1-14.34, 1980

[A14] “International Electrotechnical Vocabulary, Group 05, Fundamental Definitions,” International Electrotechnical Commission (IEC) Publication 50 (05), 1954

[A15] IEEE-Std 446, “IEEE Recommended Practice for Emergency and Standby Power Systems for Industrial and Commercial Applications,” 1995

[A16] ERDA’s report on “evaluation of parallel operation charges”.

[A17] IEC 61000-2-1(1990-05), “Description of the Environment—Electromagnetic Environment for Low Frequency Conducted Disturbances and Signaling in Public Power Supply Systems,” *Electromagnetic Compatibility (EMC)*—Part 2, Environment, Section 1, 1990.

[A18] IEEE Standard 100-1992, *IEEE Standard Dictionary of Electrical and Electronic Terms.*

[A19 IEC 61000-4-30 77A/356/CDV, *Power Quality Measurement Methods,* 2001.

[A20] IEC 61000-4-15, *Flicker Meter—Functional and Design Specifications,* 1997.

[A21] G. W. Mack and S. Santoso, “Understanding power system harmonics,” *IEEE Power Eng. Rev.*, vol. 21, no. 11, pp. 8–11, Nov. 2001.

Reference B: Pulse Width Modulation

[B1] Joachim Holtz, “Pulsewidth Modulation – A Survey,” IEEE Transactions on Industrial Electronics, Vol. 39, No. 5, Dec. 1992

[B2] Alfred Busse and Joachim Holtz, “Multiloop Control of a Unity Power Factor Fast Switching AC to DC Converter, Conference Record of IEEE-PESC 1982, pp. 171-179

[B3] Gerhard Pfaff, Alois Weschta, and Albert F. Wick, “Design and Experimental Results of a Brushless AC Servo Drive,” IEEE Transactions on Industry Applications, Vol. IA-20, No. 4, pp. 814-821, July/August, 1984

[B4] Joachim Holtz, Peter Lammert and Wolfgang Lotzkat, “High-Speed Drive System with Ultrasonic MOSFET-PWM-Inverter and Single-Chip-Microprocessor Control,” IEEE-IAS Annual Meeting Conference Record, 1986, pp. 12-17

[B5] Heinz Willi Van Der Broeck, Hans-Christoph Skudelny, and Georg Viktor Stanke, “Analysis and Realization of a Pulsewidth Modulator Based on Voltage Space Vectors,” IEEE Transactions on Industry Applications, Vol. 24, No. 1, January/February 1988

[B6] Yoshihiro Murai, Kazuharu Ohashi, and Isamu Hosono, “New PWM Method for Fully Digitized Inverters,” IEEE Transactions on Industry Applications, Vol. IA-23, No. 5, September/October 1987

[B7] Shoji Fujuda, Yoshitaka Iwaji, and Hirokazu Hasegawa, “PWM Technique for Inverter with sinusoidal Output Current,” IEEE Transactions on Power Electronics, Vol. 5, No.1, January 1990

[B8] Thomas G. Habetler, “A Space Vector-Based Rectifier Regulator for AC/DC/AC Converters,” Conference Record of EPE 1991, pp. 2.101–107

[B9] Hoachim Holtz, Wolfgang Lotzkat and Ashwin M. Khambadkone, “On Continuous Control of PWM Inverters in the Overmodulation Range Including the Six-Step Mode,” IEEE Transactions on Power Electronics, Vol. 8, No. 4, October 1993

[B10] Johann W. Kolar, Hans Ertl, and Franz C. Zach, “Influence of the Modulation Method on the Conduction and Switching Losses of a PWM Converter System,” IEEE Transactions on Industry Applications, Vol. 27, No. 6, November/December 1991

[B11] Victor R. Stefannovic and Slobodan N. Vukosavic, “Space-Vector PWM Voltage Control with Optimized Switching Strategy,” IEEE-IAS Annual Meeting 1992, pp. 1025-1033

[B12] Andrzej M. Trzynadlowski,and Stanislaw Legowski, “Minimum-Loss Vector PWM Strategy for Three-Phase Inverters,” IEEE Transactions on Power Electronics, Vol. 9, No.1, January 1994

[B13] Yoshihiro Murai, Yoichi Gohshi, Keiju Matsui, and Isamu Hosono, “High-Frequency Split Zero-Vector PWM with Harmonic Reduction for Induction Motor Drive,” IEEE Transactions on Industry Applications, Vol. 28, No. 1, January/February 1992

[B14] P. Enjeti and B. Xie, “A New Real Time Space Vector PWM Strategy for High Performance Converters,” IEEE-IAS Annual Meeting Conference Record 1992, pp. 1018-1024

[B15] Satoshi Ogasawara, Hirofumi Akagi, and Akira Nabae, “A Novel PWM Scheme of Voltage Source Inverters Based on Space Vector Theory,” Conference Record of EPE 1989, pp. 1197-1202

Reference C: Soft-Switching Power Converters

[C1] Carlos Caudros, Dusan Borojevic, Slobodan Gataric, Vlatko Vlatkovic, "Space Vector Modulated, Zero-Voltage Transition Three-Phase to DC Bidirectional Converter," Conference Record of IEEE-PESC 1994, pp. 16-23

[C2] Vlatko Vlatkovic, Dusan Borojevic, Fred C. Lee, Carlos Caudros, and Slobodan Gataric, "A New Zero-Voltage Transition, Three-Phase PWM Rectifier/Inverter Circuit," Conference Record of IEEE-PESC 1993, pp. 868-873

[C2] Hengchun Mao and Fred C. Lee, "An Improved Zero-Voltage-Transition Three- Phase Rectifier/Inverter," Conference Proceedings of IPEC 1995, pp. 853-858

[C3] Vlatko Vlatkovic, Dusan Borojevic, and Fred C. Lee, "Soft-Transition Three-Phase PWM Conversion Technology," Conference Record of IEEE-PESC 1994, pp. 79-84

[C4] Vlatko Vlatkovic, "Three-Phase Power Conversion Using Soft-Switching PWM Techniques," Ph.D. Dissertation 1994

[C5] Richard Zhang, Fred C. Lee and Dushan Boroyevich, "A Three-Phase ZVT Boost Rectifier with Improved Analog Controller," Conference Record of IEEE-PESC 1996, pp.469-474

[C6] Slobodan Gataric, Dusan Borojevic, and Fred C. Lee, "Soft-Switched, Single- Switch Three-Phase Rectifier with Power Factor Correction," Conference Record of IEEE-APEC 1994, pp. 738-744

[C7] V. Vlatkovic, D. Borojevic, X. Zhuang, and F. C. Lee, "Analysis and Design of a Zero-Voltage Switched, Three-Phase PWM Rectifier with Power Factor Correction," Conference Record of IEEE-PESC 1992, pp. 1352-1360

[C8] D. M. Divan, “The Resonant DC Link Converter – A New Concept in Static Power Conversion,” Conference Record of IEEE-IAS 1986, pp. 648-656

[C9] Jih-Sheng Lai and B. K. Bose, “An Improved Resonant DC Link Inverter for Induction Motor Drives,” Conference Record of IEEE-IAS 1988, pp. 742-748

[C10] Jinhe and Mohan, “Zero-Voltage Switching PWM Inverter for High-Frequency DC-AC Power Conversion,” Conference Record of IEEE-IAS 1990, pp. 1215-1221

[C11] A. Cheriti, et al. “A Rugged Soft Commutated PWM Inverter for AC Drives,” Conference Record of IEEE-IAS, 1990, pp. 656-662

[C12] P. K. Sood and T. A. Lipo, “Power Conversion Distribution System Using a Resonant High Frequency AC Link,” Conference Record of IEEE-IAS 1986, pp. 533-541

[C13] G. Hua, C. S. Leu, F. C. Lee, “Novel Zero-Voltage-Transition PWM Converters,” Conference Record of IEEE-PESC 1992, pp. 55-61

[C14] G. Hua, X. Yang, Y. Jiang, F. C. Lee, “Novel Zero-Current-Transition PWM Converters,” Conference Record of IEEE-PESC 1993, pp. 538-544

[C15] Hengchun Mao, “Soft-Switching Techniques for High-Power PWM Converters,” Ph.D. Dissertation, Virginia Polytechnic Institute and State University, Dec. 1996

[C16] Ned Mohan, Tore M. Undeland, William P. Robbins; "power electronics: converters applications and design, John willey and sons.

Reference D: Power Converter System

[D1] Kun Xing, Sudip Mazumder, Zhihong Ye, Fred C. Lee, and Dusan Borojevic, "The Circulating Current in Paralleled Three-Phase Boost PFC Rectifiers," Conference Record of IEEE-PESC 1998, pp. 783-789

[D2] Richard Zhang, Hengchun Mao, Fred C. Lee, and Dushan Boroyevich, "Towards Versatile Power Electronic Systems," Proceeding of VPEC Annual Seminar 1996

[D3] Richard Zhang, Fred C. Lee, Dushan Boroyevich, and Hengchun Mao, "New High Power, High Performance Power Converter System," Conference Record of IEEE-PESC 1998, pp. 8-14

[D4] Richard Zhang, Changrong Liu, Lijia Chen, Dushan Boroyevich, and Fred C. Lee, "An Utility Power Supply for Nonlinear and Unbalanced Load in a PEBB Based DC Distributed Power System," Proceeding of VPEC Annual Seminar 1998

[D5] Satoshi Ogasawara, Jin Takagaki, Hirofumi Akagi, and Akira Nabae "A Novel Control Scheme of a Parallel Current-Controlled PWM Inverter," IEEE Transactions on Industry Applications, Vol. 28, No.5, Sep./Oct. 1992

[D6] Yoshihiro Komatsuzaki, "Cross Current Control for Parallel Operating Three Phase Inverter," Conference Record of IEEE-PESC 1994, pp. 943-950

[D7] Takao Kawabata and Shigenori Higashino, "Parallel Operation of Voltage Source Inverters," IEEE Transactions on Industry Applications, Vol. 24, No.2, Mar./Apr. 1988

[D8] Keiju Matsui, "A Pulsewidth Modulated Inverter with Parallel-Connected Transisitors by Using Current Sharing Reactors," Conference Record of IEEE-IAS, 1985, pp. 1015-1019

[D9] Mitsuyuki Honbu, Yasuo Matsuda, Kouichi Miyazaki, and Yorito Jifuku, "Parallel Operation Techniques of GTO Inverter Sets for Large AC Motor Drives," IEEE Transactions on Industry Applications, Vol. IA-19, No. 2, March/April 1983

[D10] David J. Perreault, Hohn G. Kassakian, and Henrik Martin, "A Soft-Switched Parallel Inverter Architecture with Minimal Output Magnetics," Conference Record of IEEE-PESC 1994, pp. 970-977

[D11] A. Pina Martins, Adriano S. Carvalho, and Armando S. Araujo, "Design and Implementation of a Current Controller for the Parallel Operation of Standard UPSs," Conference Record of IEEE-IECON 1995, pp. 584-588

[D12] Kuniomi Oguchi, Eiichi Ikawa, and Yoshitaka Tsukiori, "A Three-Phase Sine Wave Inverter System Using Multiple Phase-Shifted Single-Phase Resonant Inverters," Conference Record of IEEE-IAS 1990, pp. 1125-1131

[D13] Earl Rhyne and Dave Bratton, "Scott-T Connected 3-Phase Inverters for Telecommunications Applications," Conference Record of IEEE-INTELEC 1986, pp. 461-468

[D14] John Reed and Naresh Sharma, "Large Parallel UPS Systems Utilizing PWM Technology," Conference Record of IEEE-INTELEC 1984, pp. 282-289

[D15] Allen Windhorn, "A Hybrid Static/Rotary UPS System," IEEE Transactions on Industry Applications, Vol. 28, No. 3, May/June 1992

[D16] Kun Xing, Fred C. Lee, J. S. Lai, Thandi Gurjit, and Dusan Borojevic, "Adjustable

Speed Drive Neutral Voltage Shift and Grounding Issues in a DC Distributed System," Conference Record of IEEE-IAS 1997, pp. 517-524

[D17] C. M. Liaw, S. J. Chiang, and S. C. Huang, “A Three-Phase Multi-Functional Battery Energy Storage System,” Conference Record of IEEE-IECON 1994, pp. 458-463

[D18] H. Jin, Y. Wang, and G. Joos, “A Hybrid Structure Using Phase-Controlled Rectifiers and High-Frequency Converters for Magnet-Load Power Supplies,” IEEE Transactions on Industrial Electronics, Vol. 43, No. 1, February 1996

[D19] Luis Moran, Luciano Fernandez, Juan Dixon, and Rogel Wallace, “A Simple and Low Cost Control Strategy for Active Power Filters Connected in Cascade,” Conference Record of IEEE-IECON 1995, pp. 463-468

[D20] N. R. Raju, S. S. Venkata, R. A. Kagalwala, and V. V. Sastry, “An Active Power Quality Conditioner for Reactive Power and Harmonics Compensation,” Conference Record of IEEE-PESC 1995, pp. 209-214

Reference E: Inverters

[E1] Thomas M. Jahns, Rik W. A. A. De Doncker, Arthur V. Radun, Paul M. Szczesny, and Fred G. Turnbull, "System Design Considerations for a High-Power Aerospace Resonant Link Converter," IEEE Transactions on Power Electronics, Vol. 8, No. 4, Oct. 1993

[E2] Robert L. Steigerwald, "Characteristics of a Current-Fed Inverter with Commutation Applied Through Load Neutral Point," IEEE Transactions on Industry Applications, Vol.IA-15, No. 5, Sep./Oct. 1979

[E3] A. L. Julian, T. A. Lipo, G. Oriti, "Elimination of Common Mode Voltage in Three Phase Sinusoidal Power Converters," Conference Record of IEEE PESC 1996, pp. 1968-1972

[E4] Richard Zhang, Himamshu Prasad, Dushan Boroyevich, and Fred C. Lee, "Analysis and Design of a Three-Phase Inverter with a Neutral Leg," Conference Record of EPE 1997, pp. 1.170-175

[E5] Yehia Tadros, Gunter Junge, and Samir Salama, "Design Aspects of High Power PWM Inverters with IGBT," Conference Record of EPE 1991, pp. 2-083-088

[E6] Isao Takahashi, Kenji Amei, and Youichi Itoh, "High Performance and Long Life Uninterruptible Power Source Using a Flywheel Energy Storage Unit," Conference Record of IEEE-IAS 1990, pp. 1049-1055

[E7] A. Kurnia, H. Cherradi, D. Divan, "Impact of IGBT Behavior on Design Optimization of Soft Switching Inverter Topologies," Conference Record of IEEE-IAS 1993, pp. 807-813

[E8] G. L. Skibinski and D. M. Divan, "Design Integration of a 200 kW GTO RDCL Converter," Conference Record of IEEE-IAS 1993, pp. 1029-1040

Reference F: Rectifiers

[F1] Prasad N. Enjeti, Phoivos D. Ziogas, and Mehrdad Ehsani, "Unbalanced PWM Converter Analysis and Corrective Measures," Conference Record of IEEE-IAS 1989, pp. 861-870

[F2] Prasad N. Enjeti and Shamin A. Choudhury, "A New Control Strategy to Improve the Performance of a PWM AC to DC Converter under Unbalanced Operating Conditions," IEEE Transactions on Power Electronics, Vol. 8, No. 4, Oct. 1993

[F3] Donato Vincenti and Hua Jin, "A Three-Phase Regulated PWM Rectifier with On-Line Feedforward Input Unbalance Correction," IEEE Transactions on Industrial Electronics, Vol. 41, No. 5, Oct. 1994

[F4] Richard Zhang, Fred C. Lee and Dushan Boroyevich, "A Three-Phase ZVT Boost Rectifier with Improved Analog Controller," Conference Record of IEEE-PESC 1996, pp. 469-474

[F5] P. Rioual and H. Pouliquen, "Regulation of a PWM Rectifier in the Unbalanced Network State," Conference Record of IEEE-PESC 1993, pp. 641-646

[F6] A. R. Prasad, P.D. Ziogas, and S. Manias, "An Active Power Factor Correction Technique for Three-Phase Diode Rectifiers," Conference Record of IEEE-PESC 1989, pp. 58-66

[F7] Vlatko Vlatkovic, Dusan Borojevic, and Fred C. Lee, "A Zero-Voltage Switched, Three-Phase Isolated PWM Buck Rectifier," IEEE Transactions on Power Electronics, Vol. 10, No. 2, March 1995

[F8] Silva Hiti, Vlatko Vlatkovic, Dusan Borojevic, and Fred C. Lee, "A New Control Algorithm for Three-Phase PWM Buck Rectifier with Input

Displacement Factor Compensation," IEEE Transactions on Power Electronics, Vol. 9, No. 2, March 1994

[F9] Lawrence J. Borle and Chem V. Nayar, "ZACE Current Controlled Power Flow for AC-DC Power Converters," Conference Record of IEEE-PESC 1994, pp. 539-545

[F10] Laszlo Huber and Dusan Borojevic, "Digital Modulator for Forced Commutated Cycloconverters with Input Power Factor Correction," Conference Record of IEEE-PESC 1992, pp. 518-523

Reference G: Active Filter

[G1] V. B. Bhavaraju and Prasad N. Enjeti, “Analysis and Design of an Active Power Filter for Balancing Unbalanced Loads,” IEEE Transactions on Power Electronics, Vol. 8, No. 4, Oct. 1993

[G2] Conor A. Quinn, Ned Mohan, “Active Filtering of Harmonic Currents in Three-Phase, Four-Wire Systems with Three-Phase and Single-Phase Non-Linear Loads,” Conference Record of IEEE-APEC, 1992, pp. 829-836

[G3] M. Bou-Rabee, D. Sutanto, F. Barone, “A New Technique for Three-Phase Active Power Filter,” Conference Record of IEEE-APEC, 1992, pp. 837-843

[G4] Kevork Haddad, Thierry Thomas, Geza Joos, and Alain Jaafari, “Dynamic Performance of Three Phase Four Wire Active Filters,” Conference Record of IEEEAPEC, 1997, pp. 206-212

[G5] H. Sasaki and T. Machida, “A New Method to Eliminate AC Harmonic Currents by Magnetic Compensation – Considerations on Basic Design,” IEEE Transactions on Power Application Systems, Vol. 90, No. 5, pp. 2009-2019, 1971

[G6] L. Gyugyi and E. C. Strycula, “Active AC Power Filters,” Conference Record of

IEEE-IAS, 1976, pp. 529-535

[G7] Hirofumi Akagi, “Trends in Active Power Line Conditioners,” Conference Record of IEEE-IECON 1992, pp. 19-24

[G8] Fang Zheng Peng, “Application Issues of Active Power Filters,” IEEE Industry Applications Magazine, September/October 1998

[G9] Masatoshi Takeda, Kazuo Ikeda, Akiyoshi Teramoto and Tomohiko Aritsuka, "Harmonic Current and Reactive Power Compensation with an Active Filter," Conference Record of IEEE-PESC 1988, pp. 1174-1179

[G10] Luis Moran, Marcelo Diaz, Victor Higuera, Rogel Wallace, and Juan Dixon, "A Three-Phase Active Power Filter Operating with Fixed Switching Frequency for Reactive Power and Current Harmonic Compensation," Conference Record of IEEE-IECON 1992, pp. 362-367

[G11] Fang Zheng Peng, Hirofumi Akagi, and Akira Nabae, "A New Approach to Harmonic Compensation in Power Systems – A Combined System of Shunt Passive and Series Active Filters," IEEE Transactions on Industry Applications, Vol. 26, No. 6, November/December 1990

[G12] Fang Zheng Peng, Hirofumi Akagi, and Akira Nabae, "Compensation Characteristics of the Combined System of Shunt Passive and Series Active Filters," Conference Record of IEEE-IAS 1989, pp. 959-966

[G13] Farrukh Kamran and Thomas G. Habetler, "Combined Deadbeat Control of a Series-Parallel Converter Combination used as a Universal Power Filter," Conference Record of IEEE-PESC 1995, pp. 196-201

[G14] Luis Moran, Phoivos Ziogas, and Geza Joos, "Analysis and Design of a Novel 3-f Solid-State Power Factor Compensator and Harmonic Suppressor System," Conference Record of IEEE-APEC 1988, pp. 267-277

[G15] Hirofumi Akagi, Yukifumi Tsukamoto, and Akira Nabae, "Analysis and Design of an Active Power Filter Using Quad-Series Voltage Source PWM Converters," IEEE Transactions on Industry Applications, Vol. 26, No. 1, January/February, 1990

[G16] Luis Moran, Pedro Godoy, Rogel Wallace, and Juan Dixon, “A New Current Control Strategy for Actove Power Filters Using Three PWM Voltage Source Inverters,” Conference Record of IEEE-PESC 1993, pp. 3-9

[G17] Beena Acharya, Deepakraj M. Divan, and Randal W. Gascoigne, “Active Power Filters using Resonant Pole Inverters,” IEEE Transactions on Industry Applications, Vol. 28, No. 6, November/December, 1992

[G18] Janko Nastran, Rafael Cajhen, Matija Seliger, and Peter Jereb, “Active Power Filter for Nonlinear AC Loads,” IEEE Transactions on Power Electronics, Vol. 9, No. 1, January 1994

[G19] Hirofumi Akagi, Yoshihira Kanazawa, and Akira Nabae, “Generalized Theory of the Instantaneous Reactive Power in Three-Phase Circuits,” Conference Record of IPEC 1983, pp. 1375-1386

[G20] C. E. Lin, C. L. Chen, and C. L. Huang, “Calculating Approach and Implementation for Active Filters in Unbalanced Three-Phase System using Synchronous Detection Method,” Conference Record of IEEE-IECON 1992, pp. 374-380

[G21] Andrea Cavallini and Gian Carlo Montanari, “Compensation Strategies for Shunt Active Filter Control,” IEEE Transactions on Power Electronics, Vol. 9, No. 6, November 1994

[G22] Johan H. R. Enslin and Jacobus D. Van Wyk, “A New Control Philosophy for Power Electronic Converters as Fictitious Power Compensators,” IEEE Transactions on Power Electronics, Vol. 5, No. 1, January 1990

[G23] P. Enjeti and S. Kim, “A New DC-Side Active Filter for Inverter Power Supplies Compensates for Unbalanced and Nonlinear Loads,” Conference Record of IEEE-IAS 1991, pp. 1023-1031

[G24] H. Akagi, Y. Kanazawa, and A. Nabae, "Instantaneous reactive power compensators comprising switching devices without energy storage components," IEEE Trans. Ind. Appl., vol. IA-20, pp, 625, 1984.

[G25] H. Sasaki and T. Machida, "A new method to eliminate ac harmonic currents by magnetic compensation-Considerations on basic design," *IEEE Trans. Power Appl. Syst.,* vol. 90, no. *5,* p. 2009, 1971.

[G26] L. Gyugyi and E. C. Strycula, "Active ac power filters," is presented at *IEEEIIAS Ann. Meeting,* 1976, pp. 529.

[G27] N. Mohan *et al.,* "Active filters for ac harmonic suppression," presented at IEEE/PES Winter Meeting, A77 026-8, 1977.

[G28] H. Kawahira, T. Nakamura, and S. Nakazawa, "Active power filters," in *Proc. JIEE IPEC-Tokyo,* 1983, pp. 981.

[G29] F. Z. Peng, H. Akagi, and A. Nabae, "Some considerations on an active power filter using quad-series voltage source PWM converters," JIEE PE-86.154, 1986, (in Japanese).

[G30] K. Komatsugi and T. Imura, "Harmonic current compensator composed of static power converter," presented at ***IEEE Power Electronics Specialist Conf.,*** 1986, pp. 283.

[G31] M. Takeda ***et*** al., "Harmonic Current Compensation with Active Filter," presented at ***IEEEIIAS Ann. Meeting,*** pp. 808, 1987.

[G32] A. Nakajima ***et al.,*** "Development of Active Filter with Series Resonant Circuit" presented at ***IEEE Power Electronics Specialist Conf.,*** pp. 1168, 1988.

[G33] M. Kohata ***et al.,*** "Compensator for harmonics and reactive power using static induction thyristors," presented at European Conf. on Power Electronics and Applications, 1987.

[G34] F. Z. Peng, H. Akagi, and A. Nabae, “A study of active power filter using quad-series voltage-source PWM converters for harmonic compensation,” presented at ***IEEE Power Electronics Specialist Conf.,***

[G35] Y . Tanoue, F. Aoyama, and K. Terashima, “5-MVA flicker suppression equipment-A hybrid system of a passive filter and an active filter using GTO thyristors,” ***Toshiba Rev.,*** vol. 43, no. 4, 1988.

[G36] H. Akagi, Y. Kanazawa, and A. Nabae, “Instantaneous reactive power compensators comprising switching devices without energy storage components,” ***IEEE*** Trans. ***Ind. Appl.,*** vol. IA-20, pp, 625, 1984.

[G37] H. Akagi, A. Nabae, and S. Atoh, “Control strategy of active power filters using multiple-voltage source PWM converters,” *IEEE Trans. Ind. Applicat.,* vol. IA-20, pp. 460–465, May/June 1986.

[G38] J. Nastran, R. Cajhen, M. Seliger, and P. Jereb, “Active power filter for Non-linear AC loads,” *IEEE Trans. Power Electron.,* vol. 9, pp. 92–96, Jan. 1994.

[G39] J. W. Dixon, J. J. Garc´ıa, and L. A. Mor´an, “Control system for three-phase active power filter which simultaneously compensates power factor and unbalanced loads,” *IEEE Trans. Ind. Electron.,* vol. 42, pp. 636–641, Dec. 1995.

[G40] F. Z. Peng, H. Akagi, and A. Nabae, “A new approach to harmonic compensation in power systems: A combined system of shunt passive and series active filters,” *IEEE Trans. Ind. Applicat.,* vol. 26, pp. 983–990, Nov./Dec. 1990.

[G41] , “Compensation characteristics of a combined system of shunt passive filters and series active filters,” *IEEE Trans. Ind. Applicat.,* vol. 29, pp. 144–152, Jan./Feb. 1993.

[G42] H. Akagi, Y. Kanazawa, and A. Nabae, “Instantaneous reactive power compensators comprising switching devices without energy storage components,” *IEEE Trans. Ind. Applicat.,* vol. IA-20, pp. 625–630, May/June 1984.

[G43] J. Jerzy and F. Ralph, “Voltage waveshape improvement by means of hybrid active power filter,” in *Proc. IEEE ICHPS VI,* Bologna, Italy, Sept. 21–23, 1994, pp. 250–255.

[G44] J. Nastran, R. Cajhen, M. Seliger, and P. Jereb, “Active power filter for Non-linear AC loads,” *IEEE Trans. Power Electron.,* vol. 9, pp. 92–96, Jan. 1994.

[G45] S. Tepper, J. Dixon, G. Venegas, and L. Mor´an, “A simple frequency independent method for calculating the reactive and harmonic current in a Non-linear load,” *IEEE Trans. Ind. Electron.,* vol. 43, pp. 647–654, Dec. 1996.

[G46] B.M.Bird, J.F.Marsh and P.R.Mclellan, "Harmonic reduction in multiple converters by triple frequency current injection”

[G47] H.Sasaki, and T.Machida,"A new method to eliminate AC harmonic currents by magnetic compensation consederation on basic design," IEEE Trans. Power Appl. System, vol PAS-90.

[G48] A. Ametani,"Harmonic reduction in thyristor converters by harmonic current injection,"

[G49] H.Akagi,Y. Kanazawa and A.Nabae," instantaneous reactive power compensators comprising switching devices without energy storage components",.

[G50] M.Aredes,Jhafner and K.Heunmann,"Three phase four wire shunt active filter control strategies", IEEE Trans. Power Electronics.

[G51] L.Moran, J.Dixon and R.Wallance, "A three phase active power filter operating with fixed switching frequency for reactive power and current harmonic compensation",

[G52] H.Akagi, A.Nabae and S.Atoh" Control stratagy of active power filters using multiple voltage source PWM converters",

[G53] H.Akagi,Y.Kanazawa and A. Nabae," Analysis and design of an active power filter using quad series voltage source PWM converters.

[G54] F.Z.Peng,H.Akagi, and A. Nabae," A study of active power filter using quad series voltage source PWM converters.

[G55] F.Z. Peng, H. Akagi, and A. Nabae," A new approach to harmonic compensation in power system - A combined system of shunt passive and series active filters.

Reference H: Modeling and Control

[H1] Silva Hiti, "Modeling and Control of Three-Phase PWM Converters," Ph.D. Dissertation, Virginia Polytechnic Institute and State University, 1995

[H2] Silva Hiti and Dushan Boroyevich, "Small-Signal Modeling and Control of Threephase PWM Converters," Conference Record of IEEE-IAS, 1994

[H3] J. Mark Noworolski and Seth R. Sanders, "Generalized In-Place Circuit Averaging," Conference Record of IEEE-APEC 1991, pp. 445-450

[H4] Luigi Malesani and Paolo Tomasin, "PWM Current Control Techniques of Voltage Source Converters – A Survey," Conference Record of IEEE-IECON 1993, pp. 670-675

[H5] Kalyan P. Gokhale, Atsuo Kawamura and Richard G. Hoft, "Dead Beat Microprocessor Control of PWM Inverter for Sinusoidal Output Waveform Synthesis," Conference Record of IEEE-IAS 1985, pp. 28-36

[H6] Atsuo Kawamura, Toshimasa Haneyoshi, and Richard G. Hoft, "Deadbeat Controlled PWM Inverter with Parameter Estimation Using Only Voltage Sensor," Conference Record of IEEE-IAS 1986, pp. 576-583

[H7] T. Kawabata, T. Miyashita, and Y. Yamamoto, "Dead Beat Control of Three Phase PWM Inverter," Conference Record of IEEE-IAS 1987, pp. 473-481

[H8] Atsuo Kawamura and Kiyoshi Ishihara, "High Frequency Deadbeat Control of Three Phase PWM Inverter Used for Uninterruptible Power Supply," Conference Record of IEEE-PESC 1988, pp. 644-649

[H9] Annette von Jouanne, Prasad N. Enjeti, and Donald J. Lucas, "DSP Control of High-Power UPS Systems Feeding Nonlinear Loads," IEEE Transactions on Industrial Electronics, Vol. 43, No. 1, February 1996

[H10] Osman Kukrer, "Discrete-Time Current Control of Voltage-Fed Three-Phase PWM Inverters," IEEE Transactions on Power Electronics, Vol. 11, No. 2, March 1996

[H11] Takao Kawabata, Takeshi Miyashita, and Yushin Yamamoto, "Digital Control of Three Phase PWM Inverter with L-C Filter," Conference Record of IEEE-PESC 1988, pp. 634-643

[H12] Seung-Gi Jeong and Myung-Ho Woo, "DSP Based Active Power Filter with Predictive Current Control," Conference Record of IEEE-IECON 1995, pp. 645-650

[H13] Takayuki Matsui, Toshiaki Okuyama, Junichi Takahashi, Takashi Sukegawa, and Kenzo Kamiyama, "A High Accuracy Current Component Detection Method for Fully Digital Vector-Controlled PWM VSI-Fed AC Drives," IEEE Transactions on Power Electronics, Vol. 5, No. 1, January 1990

[H14] Jong-Woo Choi, and Seung-Ki Sul, "Inverter Output Voltage Synthesis Using Novel Dead Time Compensation," IEEE Transactions on Power Electronics, Vol. 11, No. 2, March 1996

[H15] Juan W. Dixon, Jaime J. Garcia, and Luis Moran, "Control System for Three-Phase Active Power Filter Which Simultaneously Compensates Power Factor and Unbalanced Loads," IEEE Transactions on Industrial Electronics, Vol. 42, No. 6, December 1995

[H16] Jong-Woo Choi and Seung-Ki Sul, "New Current Control Concept – Minimum Time Current Control in 3-Phase PWM Converter," Conference Record of IEEE-IECON 1995, pp. 332-338

[H17] Akir Nabae, Satoshi Ogasawara, and Hirofumi Akagi, "A Novel Control Scheme for Current-Controlled PWM Inverters," IEEE Transactions on Industry Applications, Vol. IA-22, No. 4, July/August 1986

[H18] Michael J. Ryan and Robert D. Lorenz, “A High Performance Sine Wave Inverter Controller with Capacitor Current Feedback and ‘Back-EMF’ Decoupling,” Conference Record of IEEE-IAS 1995, pp. 507-513

[H19] Farrukh Kamran and Thomas G. Habetler, “Combined Deadbeat Control of a Series-Parallel Converter Combination Used as a Universal Power Filter,” Conference Record of IEEE-IAS 1995, pp. 196-201

[H20] M. C. Chandorkar, D. M. Divan, and B. Banerjee, “Control of Distributed UPS Systems,” Conference Record of IEEE-IAS 1994, pp. 197-204

[H21] Yukihiko Sato and Teruo Kataoka, “A Current Type PWM Rectifier with Active Damping Function,” Conference Record of IEEE-IAS 1995, pp. 2333-2340

[H22] Yan Guo, Xiao Wang, Howard C. Lee, and Boon-Teck Ooi, “Pole-Placement Control of Voltage-Regulated PWM Rectifiers Through Real-Time Multiprocessing,” IEEE Transactions on Industrial Engineering, Vol. 41, No. 2, April 1994

[H23] Silva Hiti and Dushan Boroyevich, “Control of Front-End Three-Phase Boost Rectifier,” Conference Record of IEEE-PESC 1994, pp. 927-933

[H24] Silva Hiti, Dusan Borojevic, Ravindra Ambatipudi, Richard Zhang, and Yimin Jiang, “Average Current Control of Three-Phase PWM Boost Rectifier,” Conference Record of IEEE-APEC 1995, pp. 131-137

[H25] Yasuhiko Miguchi, Atsuo Kawamura, and Richard Hoft, “Decoupling Servo- Control of Three-Phase PWM Inverter for UPS Application,” Conference Record of IEEE-IAS 1987, pp. 724-733

Reference I: Passive Filters

[I1] S. Srivastva, R. Koul, A. K. Tripathi “Application and design trends of filters”

[I2] H.Sasaki, and T.Machida,"A new method to eliminate AC harmonic currents by magnetic compensation consederation on basic design," IEEE Trans. Power Appl. System, vol PAS-90.

Reference J: Negative Sequence Compensator

[J1] TR Mukundan & P Kundu, “Method of Reducing the effect of supply unbalance on induction motor” IEEE Power Engineering Review May 2000.

[J2] SS choi, BH Li and DM Vilath Gamuvia, “Dynamic Voltage Restoration with minimum energy injection”. IEEE Transactions on Power Systems Vol. 15 Feb. 2000

[J3] A Campos,G Joos, P.W. Ziogas and J F Lindsay, “Analysis and Design of a series voltage unbalance compensator based on 3 phase VSI operating with unbalanced switching functions”. IEEE Transactions on Power Electronics. Vol. 9 May 1994.

[J4] Mauricio Aredes, Jurgen Hafner and Klemens Heumann, “Three phase four wire shunt active filter control strategies”. IEEE transaction on power electronics. Vol. 12 No 2, March 1997.

[J5] A Campos, G Joos, P.W. Ziogas and J F Lindsay, “Analysis and Design of a series voltage unbalance compensator based on 3 phase VSI operating with unbalanced switching functions”. IEEE Transactions on Power Electronics. Vol. 9 May 1994.

[J6] S. Srivastva, R. Koul, A. K. Tripathi "Application and design trends of filters"

[J7] H.Sasaki, and T.Machida, "A new method to eliminate AC harmonic currents by magnetic compensation consederation on basic design," IEEE Trans. Power Appl. System, vol PAS-90.

[J8] T. J. E. Miller "Reactive Power Control In Electric Systems" 1st Edition 1982: A Wiley-Interscience Publication John Willey & Sons Newyork,

[J9] R Bruce Shipley, "Introduction To Matrices And Power Systems" 1st Edition 1976: The Commonwelth And International Library Of Science, Technology, Engineering And Liberal Studies, A Wiley-Interscience Publication John Willey & Sons Newyork

[J10] Colonel Wm. T. Mclyman, "Transformers And Inductors Design Handbook" A User's Guide To Practice & Specification, 2nd Edition , Marcel Dekker, Newyork

[J11] Blackburn, J.L., Symmetrical Components for Power Systems Engineering, Marcel Dekker, New York, 1993.

[J12] Grainger, J.J. and Stevenson, Jr., W.D., Power System Analysis, McGraw-Hill, Inc., New York, 1994.

[J13] Krause, P.C., Analysis of Electric Machinery, McGraw-Hill, New York, 1986.

[J14] Kundur, P., Power System Stability and Control, McGraw-Hill, Inc., New York, 1994.

[J15] K.F. Rilly, M.P. Hobson, S.J. Bence, "Mathematical Methods for Physics and Engineering", Cambridge University Press.

[J16] N. N. Hancock, "Matrix Analysis of Electrical Machinery" , 1st Edition 1964, Pergmon Press Ltd., Headington Hill Hall, Oxford 4&5 Fitzroy Square, London W.1. The Macmillan Company 60 Fifth Avenue, Newyork 11, N.Y.

[J17] Peter Vas, "Vector Control Of Ac Machine" 1990, Edition, Clarendon Press, Walton Street, Oxford Ox2 6dp And Oxford University Press , New York.

[J18] Dr. P. S. Bimbhra, "Generalized Theory Of Electrical Machines", 5th Edition 1995, Khanna Publishers, 2-B, Nath Market, Nai Sadak, New Delhi 110006

[J19] Pal K. Kovacs, "Transient Phenomena In Electrical Machines" 2nd Edition 1984 , 1st Edition 1907, Head Office: Elsevier Science Publishers B. V., 1 Molenwerf, P.O. Box 211, 1000 Ae Amsterdam, The Netherlands

[J20] L. S. Czarnecki, "On some misinterpretations of the instantaneous reactive power theory," *IEEE Trans. Power Electron.*, vol. 19, no. 3, pp. 828–836, May 2004.

[J21] A. Ghosh and A. Joshi, "A new approach to load balancing and power factor correction in power distribution system," *IEEE Trans. Power Del.*, vol. 15, no. 1, pp. 417–422, Jan. 2000.

[J22] G. Chicco, P. Postolache, and C. Toader, "Analysis of three-phase systems with neutral under distorted and unbalanced conditions in the symmetrical component-based framework," *IEEE Trans. Power Del.*, vol. 22, no. 1, pp. 674–683, Jan. 2007.

[J23] S. George and V. Agarwal, "A novel, DSP based algorithm for optimizing the harmonics and reactive power under non-sinusoidal supply voltage conditions," *IEEE Trans. Power Del.*, vol. 20, no. 4, pp. 2526–2534, Oct. 2005.

LIST OF PAPERS PUBLISHED

PAPERS PRESENTED AT INTERNATIONAL AND NATIONAL CONFERENCES

Sr. No	Name Of Conference	Organized By	Date of Events	Title of Paper
1	National Workshop on Reactive Power at ERDA Vadodara	ERDA Vadodara.	29-30, Sept. 2005	Series Active Filter- Elimination of Voltage Harmonics
2	National Seminar on "application of power electronics to the power system"	Society of Power Engineer (India) Vadodara Chapter	10-11 June 2006	Series Active Filter-Solution for Elimination of Voltage Harmonics for Sensitive Load
3	International conference on "Recent Advancements & Application of Computer in Electrical Engineering" RACE'07	Govt. Engineering College Bikaner. Sponsored by IEEE	24-25 March 2007	Shunt Active Power Filter to Improve the Power Quality Based on IARP Theory
4	INDICON 2007 & 16th annual symposium of IEEE Bangalore section, at CPRI, Bangalore	Organized By IEEE Bangalore Section & IEEE India Council	6-8, Sept. 2007	Novel Technique to compensate Negative Sequence & Zero Sequence Voltage Using Instantaneous Active Reactive Power Theory
5	IEI Journal 8, Gokhale Road, Calcutta 700 020	IEI Journal Volume 89, 8, Gokhale Road, Calcutta 700 020	Dec. 2008, Page No 31 to 36	Novel Technique to compensate Negative Sequence & Zero Sequence Voltage Using Instantaneous Active Reactive Power Theory
6	National Power Electronics Conference, NPEC 2007	IISc, Bangalore & IEEE Bangalore Section	17-18, Dec. 2007	Novel Single Phase Boost Converter with low THD & Unity PF for Power Quality Improvement in Domestic Application
7	National Power Systems Conference, NPSC-2010	Osmania University, Hyderabad	15-17 Dec. 2010	Single phase Active harmonic filters for Harmonic elimination and Power Factor correction for Distributed loads
8	National Power Systems Conference, NPSC-2010	Osmania University, Hyderabad	15-17 Dec. 2010	Comparison of Amorphous & CRGO Core Transformer Losses Under Nonlinear Load Condition

Annexure I

Transformer core loss evaluation under harmonics conditions

AI.1 INTRODUCTION

Transformer is very important static power system equipment in service for more than a century having highest efficiency. However, of late, the performance of distribution transformer has become critical due to mounting failures. The comforts of new era and super refined control of various power operations is the boon from advancements of power electronics. However, same has added the worst of distortions to the input voltage to transformer and huge harmonics of unprecedented magnitude in the load current supplied by the transformer. Therefore, transformers have become vulnerable; their losses have mounted manifold resulting in very poor efficiency and reduced life.

At the same time, advents of new material and improved core configurations have made it possible to find solutions to the problem. Amorphous magnetic material and wound delta core configuration are typical examples which have offered specific advantages.

It is well known that transformers with amorphous core material are having no load losses lower by 70-80% as compared to transformer with CRGO core at normal power frequency operation. Due to more luxuries and comforts being added through electronics and digital communications, the use of non-linear load world-wide has grown up from less than 5% in 1990 to greater than 50% in 2000. Situation in India is not much different at present. Therefore, utilities in Japan have noticed the increase in losses while supplying nonlinear load. However they have noticed that increase is much more in the case of CRGO core transformers. The increase in load losses due to non-linearity in load current can be evaluated using K-factor, which is well defined in IEEE C57.110-1998. However, the increase in no-load losses due to non-linearity in supply voltage has not been emphasized yet properly.

Based on previous experience through experiments, specific transformers were designed and got manufactured having identical coils losses with CRGO and Amorphous Metal cores with 25 kVA 11000/433 Volts, three phase transformer & 10 kVA, 11000/240 Volts rating. This helped to assign the increase in losses to core and eddy current losses. Authors have measured no-load losses of 10 kVA, 11000/240 Volts rating CRGO and Amorphous Core Transformer (AMT) having identical coil at different frequencies and flux densities. Dedicated and sophisticated equipment is identified having measurement accuracy of 0.03% over entire frequency range of interest. A new mathematical model is developed at author's laboratory for evaluation of transformer core losses in situation where supply voltage is distorted.

AI.2 MATHEMATICAL MODEL DEVELOPED AT ERDA LABORATORY

ERDA has developed a simple method to find the empirical constants required to evaluate core losses of a distorted Voltage Fed Transformer through experimental results taking only design value of flux density as input. The mathematical model so developed is:

$$W = K_1.B_s{}^p.f + K_2.B_s{}^q.f^m \tag{1}$$

Where W is the total iron loss,

` B_s is the specified flux density,

f is the applied power frequency

K_1, K_2, p, q and m are the constants found out experimentally.

For a given design, the empirical co-efficient so derived in above mathematical model through sets of experiments are valid for particular design & rating of transformer with any given situation having different combinations of voltage and current distortions.

The above mathematical model is independent of design variables like mass, specific resistance, density, thickness of lamination, coefficient of hysteresis loss, etc. used in theoretical formula.

AI.3 CORE LOSS ANALYSIS OF TRANSFORMER WHILE FED FROM NON-SINUSOIDAL SUPPLY VOLTAGE

The core loss is measured using 0.03% accurate Power Analyzer at 50, 60, 70, 80, 90 & 100 Hz by maintaining rated flux density and half the rated flux density.

The mathematical model used is:

$$W = K_1.{B_s}^P.f + K_2.{B_s}^Q.f^m \qquad (1)$$

Iron loss = Hysteresis loss + Eddy current loss

Where W is the total iron loss,

B_s is the specified flux density,

f is the applied frequency

K_1, K_2, P, Q and m are the constants found out experimentally.

Equation 1 can be rewritten as

$$\mathrm{W} = A.f + B.f^m \qquad (2)$$

Where

$$A = K_1.B_s^P \qquad (3)$$

$$B = K_2 . B_s^Q \tag{4}$$

Dividing equation (2) by f we get

$$\frac{W}{f} = A + B . f^{\,m-1} \tag{5}$$

So, if equation (5) is plotted and the constants A & B are found from proper curve fitting, the value of m is calculated by substituting the values of A and B in equation (5).

A different sets of experiment with half the rated flux densities were done at same set of frequencies as mentioned above to obtain two different values of constant A; A_1 (at rated flux density B_{s1}) *and* A_2 (at half the rated flux density B_{s2}) and constant B; B_1 (at rated flux density B_{s1}) and B_2 (at half the rated flux density B_{s2}), which is obtained from equations (3) and (4) as

$$A_1 = K_1 . B_{s1}^P \tag{6}$$

$$A_2 = K_1 . B_{s2}^P \tag{7}$$

$$B_1 = K_2 . B_{s1}^Q \tag{8}$$

$$B_2 = K_2 . B_{s2}^Q \tag{9}$$

Taking ratio of equations (6) and (7) we get

$$\frac{A_1}{A_2} = \left[\frac{B_{s1}}{B_{s2}}\right]^P$$

$$\therefore \; P = \left[\frac{Ln\left(\dfrac{A_1}{A_2}\right)}{Ln\left(\dfrac{B_{s1}}{B_{s2}}\right)}\right] \tag{10}$$

Taking ratio of equations (8) and (9) we get

$$\frac{B_1}{B_2} = \left[\frac{B_{s1}}{B_{s2}}\right]^Q$$

$$\therefore\ Q = \left[\frac{Ln\left(\frac{B_1}{B_2}\right)}{Ln\left(\frac{B_{s1}}{B_{s2}}\right)}\right] \qquad (11)$$

K_1 and K_2 are calculated using equations for A_1 equation no (6) & and B_1 equation no (6).

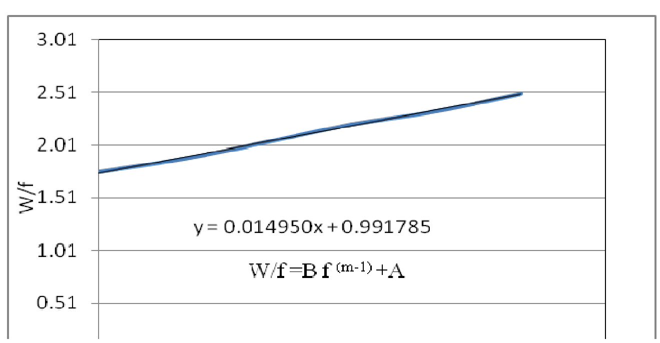
Now, various constants like K_1, K_2, P, Q & m have been found. Flux densities corresponding to each individual harmonic voltages and frequency are found out. Eddy current and hysteresis loss is found out by substituting the values of flux densities and above constants in the formula for each individual harmonic frequency. The total core loss is found out by summing up the obtained eddy current loss and hysteresis loss.

AI.3.1 RESULTS

The core loss is measured at different frequency by maintaining rated flux density at *V/f=4.8* & 2.4 for Three phase transformer and graph plotted between *W/f* and *f* is as shown below

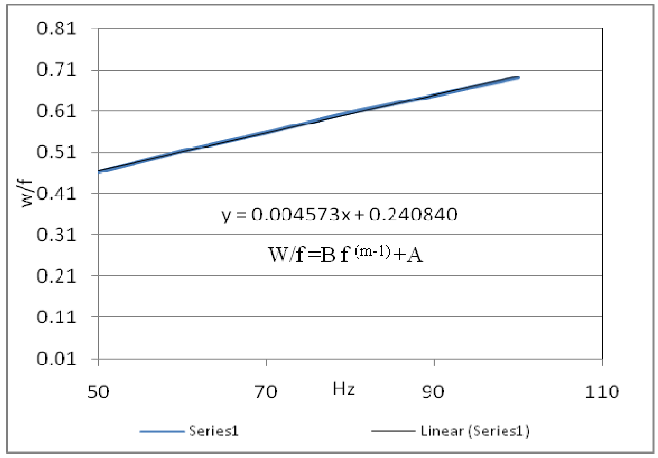
CRGO, 3 Ø, 25 kVA, 11000/433Volts

For V/F	8.66	
Freq in Hz	W/f	m1
50	1.75	2.00
60	1.87	2.00
70	2.04	2.00



80	2.20	2.00
90	2.32	2.00
100	2.49	2.00
	Avg	2.00

V/F	4.8	
Freq in Hz	W/f	m1
50	0.47	2.00
60	0.52	2.00
70	0.56	2.00
80	0.61	2.00
90	0.65	2.00
100	0.70	2.00
	Avg	2.00



Similarly From the above graph the values of A1, A2, B1, B2 and m is obtained for CRGO three phase transformer as explained earlier in the methodology.

	Flux Density		1.54	(as per design)					
A1	**A2**	**B1**	**B2**	**p**	**q**	**m1**	**m2**	**K1**	**K2**
0.9918	0.2408	0.0150	0.0046	2.3985	2.0074	2.0000	1.9999	0.0056	0.0002
								0.0855	0.0019

AI.4 LOAD LOSS ANALYSIS OF TRANSFORMER WHILE SUPPLYING NON-SINUSOIDAL LOAD CURRENT

The transformer is characterized to find out load loss and ohmic (I^2R) loss at linear load condition.

The stray loss is determined by subtracting I^2R loss from load loss. This stray loss comprises of winding eddy current loss and other stray losses. The winding eddy current loss is taken to be 33% of total stray loss & remaining being other stray loss.

Harmonic loss factor F_{HL-EC} for winding eddy current loss is determined as given below.

$$F_{HL-EC} = \frac{\sum_{h=1}^{h=h_{\max}} I_h^2 . h^2}{\sum_{h=1}^{h=h_{\max}} I_h^2}$$

Where I_h is the individual harmonic current and h is the harmonic order.

Harmonic loss factor F_{HL-STR} for other stray loss is determined as given below.

$$\mathrm{F_{HL-STR}} = \frac{\sum_{\mathrm{h=1}}^{\mathrm{h=h_{max}}} \mathrm{I_h^2 . h^{0.8}}}{\sum_{\mathrm{h=1}}^{\mathrm{h=h_{max}}} \mathrm{I_h^2}}$$

The loss density is given by

$P_{LL}(pu) = \sum_{h=h_1}^{h=h_{\max}} \left(\frac{I_h}{I_1} \right)^2 .X$, where X is the per unit loading.

The I^2R loss at linear load condition is multiplied with $P_{LL}(pu)$ to obtain I^2R loss at non-linear load condition.

The winding eddy current loss at linear load condition is multiplied with $P_{LL}(pu)$ and F_{HL-EC} to obtain eddy current stray loss at non-linear load condition.

The other stray loss at linear load condition is multiplied with $P_{LL}(pu)$ and F_{HL-STR} to obtain other stray loss at non-linear load condition

The addition of thus obtained I^2R loss, winding eddy current loss and other stray loss results in total increased load loss at non-linear load condition.

AI.4.1.1 CRGO TRANSFORMER 3 Ø 25 KVA, 11/0.433 KV, 50 HZ, OIL COOLED:

1. Linear losses at rated condition and 75 °C

PARAMETER	HV	LV	HV	LV	
VOLTAGE(Volts)	11000	433			
CURRENT (Amps)	1.31	33.33			
MEASURED RESISTANCE (Ω) AT	30.5°C		75°C		
1U-1V(HV)//2U-2V(LV)	107.36	0.12	125.39	0.14	Ω
1V-1W(HV)//2V-2W(LV)	106.92	0.12	124.87	0.14	Ω
1W-1U(HV)//1W-1U(LV)	107.62	0.12	125.69	0.14	Ω
MEASURED RESISTANCE PER PHASE IN Ω	160.95	0.058728	187.9775	0.06859	Ω
NO LOAD LOSSES			85.1		W
LOAD LOSSES AT 75°C			565		WATTS AT 75°C
TOTAL					
I^2R LOSSES	277.12	195.77	323.65	228.65	W
TEMPERATURE, °C	30.5		75		
TOTAL I2R LOSSES		472.89		552.30	WATTS AT 75°C
STRAY LOSSES & EDDY LOSSES				12.70	WATTS AT 75°C
EDDY CURRENT LOSSES				4.19	WATTS AT 75°C
OTHER STRAY LOSSES				8.51	WATTS AT 75°C
SUMMARY					
TEMPERATURE, °C			75 °C		
NO LOAD LOSSES				85.10	W
I^2R LOSSES				552.30	W
STRAY & EDDY LOSSES				12.70	W
EDDY CURRENT LOSSES				4.19	W
OTHER STRAY LOSSES				8.51	W
TOTAL LOSSES				650.10	W

2. Measured Data at Site

Input		Output	
U-1-Total	10198.00	U-3-Total	389.50
U-2-Total	10195.60	U-4-Total	390.60
I-1-Total	1.07	I-3-Total	25.97
I-2-Total	0.95	I-4-Total	26.40
P-1-Total	3898.20	P-3-Total	9535.30
P-2-Total	9168.80	P-4-Total	3130.20
S-1-Total	10901.60	S-3-Total	10114.20
S-2-Total	9714.30	S-4-Total	10312.30
Q-1-Total	10180.90	Q-3-Total	-3372.90
Q-2-Total	-3209.40	Q-4-Total	9825.70
PF-1-Total	0.36	PF-3-Total	0.94
PF-2-Total	0.94	PF-4-Total	0.30
P-SigmaA-Total	13066.90	P-SigmaB-Total	12665.50
S-SigmaA-Total	17853.90	S-SigmaB-Total	17689.80
Q-SigmaA-Total	6971.40	Q-SigmaB-Total	6452.80
FreqU-1-Total	49.88		

Losses at Site condition	
TEMPERATURE, °C	45
NLL at 390 Volts	62.82 W
I^2R	498.77 W
STRAY & EDDY LOSSES	14.06 W
EDDY CURRENT LOSSES	4.64 W
OTHER STRAY LOSSES	9.42 W
TOTAL LOSSES	575.65 W

Input Power	Output Power	Difference
13066.90	12665.50	401.40

3. Harmonic Analysis of Input & Output Voltage & Current

	Input				Output				Input				Output			
Harm order	volt Mag	% of funda	volt Mag	% of funda	volt Mag	% of funda	volt Mag	% of funda	current mag	% of funda	current mag	% of funda	current mag	% of funda	current mag	% of funda
1	10170.4	100.0	10166.9	100.0	385.7	100.0	387.5	100.0	1.02	100.0	0.90	100.0	24.73	100.0	25.1	100.0
2	4.0	0.0	4.8	0.0	0.4	0.1	0.3	0.1	0.0	0.6	0.0	1.0	0.1	0.3	0.1	0.5
3	63.0	0.6	24.3	0.2	1.2	0.3	3.7	1.0	0.1	6.4	0.0	2.4	1.3	5.2	1.6	6.5
4	1.4	0.0	0.9	0.0	0.1	0.0	0.1	0.0	0.0	0.2	0.0	0.3	0.0	0.1	0.0	0.2
5	675.6	6.6	654.6	6.4	39.5	10.2	37.1	9.6	0.2	22.3	0.2	24.8	5.9	23.8	5.7	22.8
6	6.5	0.1	2.7	0.0	0.2	0.1	0.3	0.1	0.0	0.2	0.0	0.2	0.0	0.2	0.0	0.2
7	199.9	2.0	234.8	2.3	21.7	5.6	20.7	5.3	0.1	11.4	0.1	14.7	3.1	12.6	3.2	12.7
8	3.1	0.0	5.1	0.1	0.3	0.1	0.3	0.1	0.0	0.1	0.0	0.2	0.0	0.0	0.0	0.1
9	44.8	0.4	10.3	0.1	1.5	0.4	2.3	0.6	0.0	1.6	0.0	0.7	0.3	1.4	0.4	1.5
10	0.8	0.0	2.7	0.0	0.2	0.1	0.1	0.0	0.0	0.1	0.0	0.1	0.0	0.1	0.0	0.0
11	223.4	2.2	202.1	2.0	16.7	4.3	15.2	3.9	0.1	5.8	0.1	6.1	1.5	6.1	1.4	5.6
12	3.0	0.0	2.3	0.0	0.1	0.0	0.3	0.1	0.0	0.1	0.0	0.0	0.0	0.0	0.0	0.1
13	139.3	1.4	181.6	1.8	14.2	3.7	13.3	3.4	0.0	3.3	0.0	4.8	1.0	3.8	0.9	3.6
14	1.4	0.0	2.6	0.0	0.2	0.1	0.2	0.1	0.0	0.0	0.0	0.1	0.0	0.0	0.0	0.0
15	42.1	0.4	10.7	0.1	1.3	0.3	2.3	0.6	0.0	0.9	0.0	0.3	0.2	0.7	0.2	0.8
16	0.4	0.0	1.3	0.0	0.2	0.1	0.1	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
17	121.1	1.2	109.1	1.1	9.9	2.6	8.2	2.1	0.0	2.3	0.0	2.4	0.6	2.5	0.5	2.1
18	2.6	0.0	2.9	0.0	0.1	0.0	0.2	0.1	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
19	94.5	0.9	141.9	1.4	10.6	2.7	9.4	2.4	0.0	1.3	0.0	2.4	0.4	1.7	0.4	1.5
20	2.1	0.0	3.4	0.0	0.2	0.1	0.2	0.1	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
21	44.8	0.4	14.1	0.1	1.4	0.4	2.4	0.6	0.0	0.6	0.0	0.3	0.1	0.5	0.2	0.6
22	1.3	0.0	1.4	0.0	0.2	0.1	0.1	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
23	75.7	0.7	72.0	0.7	6.9	1.8	5.0	1.3	0.0	1.2	0.0	1.3	0.3	1.3	0.2	1.0
24	2.5	0.0	2.2	0.0	0.1	0.0	0.2	0.1	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
25	51.3	0.5	105.1	1.0	7.9	2.0	6.1	1.6	0.0	0.6	0.0	1.3	0.2	0.9	0.1	0.6
%THD/Ampl	**10200.22**	**7.66**	**10196.70**	**7.66**	**389.36**	**13.82**	**390.66**	**12.80**	**1.06**	**26.93**	**0.94**	**30.22**	**25.73**	**28.60**	**26.07**	**27.92**

4. Calculation of K- Factor

CALCULATION OF HARMONIC LOSS FACTOR & K-FACTOR												
Harm.No	**Current (I)**	**(Ih/I1)**	**h^2**	**$(Ih/I1)^2$**	**$(Ih/I1)^{2h2}$**	**h0.8**	**$(Ih/I1)^{2h0.8}$**	**(Ih/IR)**	**$(Ih/IR)^2$**	**$(Ih/IR)^{2h2}$**	**$(Ih/IR)^{2h0.8}$**	**$(Ih)^2$**
h	**Amps**											
1	24.73	1.00	1.00	1.00	1.00	1.00	1.00	0.74	0.55	0.55	0.55	611.75
2	0.08	0.00	4.00	0.00	0.00	1.74	0.00	0.00	0.00	0.00	0.00	0.01
3	1.29	0.05	9.00	0.00	0.02	2.41	0.01	0.04	0.00	0.01	0.00	1.65
4	0.02	0.00	16.00	0.00	0.00	3.03	0.00	0.00	0.00	0.00	0.00	0.00
5	5.88	0.24	25.00	0.06	1.41	3.62	0.21	0.18	0.03	0.78	0.11	34.62
6	0.04	0.00	36.00	0.00	0.00	4.19	0.00	0.00	0.00	0.00	0.00	0.00
7	3.12	0.13	49.00	0.02	0.78	4.74	0.08	0.09	0.01	0.43	0.04	9.72
8	0.01	0.00	64.00	0.00	0.00	5.28	0.00	0.00	0.00	0.00	0.00	0.00
9	0.34	0.01	81.00	0.00	0.02	5.80	0.00	0.01	0.00	0.01	0.00	0.12
10	0.02	0.00	100.00	0.00	0.00	6.31	0.00	0.00	0.00	0.00	0.00	0.00
11	1.51	0.06	121.00	0.00	0.45	6.81	0.03	0.05	0.00	0.25	0.01	2.27
12	0.01	0.00	144.00	0.00	0.00	7.30	0.00	0.00	0.00	0.00	0.00	0.00
13	0.95	0.04	169.00	0.00	0.25	7.78	0.01	0.03	0.00	0.14	0.01	0.90
14	0.00	0.00	196.00	0.00	0.00	8.26	0.00	0.00	0.00	0.00	0.00	0.00
15	0.17	0.01	225.00	0.00	0.01	8.73	0.00	0.01	0.00	0.01	0.00	0.03
16	0.01	0.00	256.00	0.00	0.00	9.19	0.00	0.00	0.00	0.00	0.00	0.00
17	0.61	0.02	289.00	0.00	0.17	9.65	0.01	0.02	0.00	0.10	0.00	0.37
18	0.01	0.00	324.00	0.00	0.00	10.10	0.00	0.00	0.00	0.00	0.00	0.00
19	0.43	0.02	361.00	0.00	0.11	10.54	0.00	0.01	0.00	0.06	0.00	0.19
20	0.00	0.00	400.00	0.00	0.00	10.99	0.00	0.00	0.00	0.00	0.00	0.00
21	0.12	0.00	441.00	0.00	0.01	11.42	0.00	0.00	0.00	0.01	0.00	0.01
22	0.01	0.00	484.00	0.00	0.00	11.86	0.00	0.00	0.00	0.00	0.00	0.00
23	0.31	0.01	529.00	0.00	0.08	12.29	0.00	0.01	0.00	0.05	0.00	0.10
24	0.00	0.00	576.00	0.00	0.00	12.71	0.00	0.00	0.00	0.00	0.00	0.00
25	0.23	0.01	625.00	0.00	0.06	13.13	0.00	0.01	0.00	0.03	0.00	0.05
∑	25.73	1.61		1.08	4.38		1.34	0.77	0.60	2.41	0.74	661.79

5. Calculation of Core Loss Due to Distortion in Supply Voltage

V_{THD}	13.817	Bs=	1.54000	at V/f=	8.80000		
Vrms	389.36	k1=	0.00559	k2=	0.00020		
Frequency	49.995 Hz	m=	2.00003				
		p=	2.39854	q=	2.00736		
No.	**Voltage(V)**	**Frequency(Hz)**	**V/f**	**B_s**	**$k1B_s^Pf$**	**$k2B_s^qf^m$**	**Calculated Loss**
1	379.10	50.00	7.58	1.33	36.06	28.62	64.68
2	0.30	99.99	0.00	0.00	0.00	0.00	0.00
3	1.10	149.99	0.01	0.00	0.00	0.00	0.00
4	0.10	199.98	0.00	0.00	0.00	0.00	0.00
5	42.10	249.98	0.17	0.03	0.02	0.34	0.36
6	0.20	299.97	0.00	0.00	0.00	0.00	0.00
7	23.30	349.97	0.07	0.01	0.00	0.10	0.11
8	0.20	399.96	0.00	0.00	0.00	0.00	0.00
9	1.40	449.96	0.00	0.00	0.00	0.00	0.00
10	0.10	499.95	0.00	0.00	0.00	0.00	0.00
11	16.60	549.95	0.03	0.01	0.00	0.05	0.05
12	0.20	599.94	0.00	0.00	0.00	0.00	0.00
13	14.30	649.94	0.02	0.00	0.00	0.04	0.04
14	0.10	699.93	0.00	0.00	0.00	0.00	0.00
15	1.20	749.93	0.00	0.00	0.00	0.00	0.00
16	0.00	799.92	0.00	0.00	0.00	0.00	0.00
17	9.60	849.92	0.01	0.00	0.00	0.02	0.02
18	0.20	899.91	0.00	0.00	0.00	0.00	0.00
19	10.50	949.91	0.01	0.00	0.00	0.02	0.02
20	0.20	999.90	0.00	0.00	0.00	0.00	0.00
21	1.40	1049.90	0.00	0.00	0.00	0.00	0.00
22	0.10	1099.89	0.00	0.00	0.00	0.00	0.00
23	6.70	1149.89	0.01	0.00	0.00	0.01	0.01
24	0.20	1199.88	0.00	0.00	0.00	0.00	0.00
25	7.40	1249.88	0.01	0.00	0.00	0.01	0.01
Calculated Power (W)					36.08	29.22	65.30

6. Loss Calculation Summary

EDDY CURRENT HARMONIC LOSS FACTOR =			**4.05**			
STRAY LOSSES HARMONIC LOSS FACTOR =			**1.24**			
K-FACTOR FOR EDDY CURRENT=			**2.41**			
K-FACTOR FOR STRAY LOSSES=			**0.74**			
Loading	0.77					
Pll(pu)	1.04					
Pll(pu) at Particular Loading	0.62					
Temp Correction Factor						
	Rated Loss AT 75°C	**Linear Loss at Actual Loading at 45°C**	**Harmonic Multiplier**	**Computed Losses**	**Measured Losses**	**Deviation in %**
No Load Losses	85.10	62.82		65.30		8.28
I^2R Losses	552.30	308.97		308.97		
Winding Eddy Losses	4.19	2.87	4.05	11.63		
Other Stray Losses	8.51	5.84	1.24	7.22		
Total Losses	650.10	380.50		393.12	401.40	2.11
DIFFERENCE(Non linear-linear)					20.90	5.49

Annexure II-A: Filter Design Calculation For Electronics Ballast:

Source ind Zs in henery	Q of Zs	Source res. Rs	Reso. React. X_{Lr}	Reso. Ind. L_r	Reso. res	Q reso.	Reso. Cap X_{Cr}	Xcr	Tuned factor (N)	%THD	Fundamental	ballast current	Remarks
1.00	8.0	39.270	520.11	1.656	43.3	12.0	0.00000068	4681.0	3.0	104.2	0.1837	0.265	SELECT Resonance . Cap

Table 1 - Numerical results of impedances, voltages and currents at characteristeric harmonic orders of a series resonant circuit in a network with distorted supply voltage

h	X_s	R_s	X_{Cr}	X_{Lr}	R_r	X_L+(-X_C)	Impedance Z_f	Impedance Zs	I_h		I_f	I_s	I_s	U_B	U_B(%)	U-X_{lr}	U-X_{cr}
	Ohms	ohm	Ohms	Ohms		Ohms	Ohms	Ohms	amp	%	A	A	%	V	%	V	V
1	314.2	39.3	4681.0	520.1	43.3	-4160.9	4161.1	316.6	0.2	100.0	0.0	0.2	100.0	55.0	143.7	6.9	61.9
3	942.5	39.3	1560.3	1560.3	43.3	0.0	43.3	943.3	0.1	73.8	0.1	0.0	3.5	5.7	14.9	205.9	205.9
4	1256.6	39.3	1170.3	2080.5	43.3	910.2	911.2	1257.3	0.0	0.0	0.0	0.0	0.0	0.0		0.0	0.0
5	1570.8	39.3	936.2	2600.6	43.3	1664.4	1664.9	1571.3	0.1	46.0	0.0	0.0	25.5	69.5	181.6	108.6	39.1
6	1885.0	39.3	780.2	3120.7	43.3	2340.5	2340.9	1885.4	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
7	2199.1	39.3	668.7	3640.8	43.3	2972.1	2972.4	2199.5	0.1	35.8	0.0	0.0	22.2	84.7	221.3	103.7	19.1
9	2827.4	39.3	520.1	4681.0	43.3	4160.9	4161.1	2827.7	0.1	27.3	0.0	0.0	17.5	85.9	224.3	96.6	10.7
11	3455.8	39.3	425.5	5721.3	43.3	5295.7	5295.9	3456.0	0.0	27.3	0.0	0.0	7.0	41.8	109.3	45.2	3.4
13	4084.1	39.3	360.1	6761.5	43.3	6401.4	6401.6	4084.3	0.0	10.7	0.0	0.0	1.8	12.5	32.6	13.2	0.7
15	4712.4	39.3	312.1	7801.7	43.3	7489.6	7489.8	4712.6	0.0	8.0	0.0	0.0	5.3	43.4	113.3	45.2	1.8
17	5340.7	39.3	275.4	8841.9	43.3	8566.6	8566.7	5340.9	0.0	14.4	0.0	0.0	9.6	88.8	232.0	91.7	2.9
19	5969.0	39.3	246.4	9882.2	43.3	9635.8	9635.9	5969.2	0.0	11.8	0.0	0.0	7.8	81.1	211.8	83.2	2.1
										DF(A) %=	104.2		DF(B) %=	41.1	201.7	507.0	

Annexure II-B

P-Q THEORY WITH ZERO SEQUENCE CURRENT

It is possible to extend the instantaneous reactive power theory developed for three phase circuit including zero phase sequence components. The instantaneous space vectors ea, eb, and ec are transformed to 0-α-β co-ordinates as follows:

$$\begin{bmatrix} eo \\ e\alpha \\ e\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ 1 & \frac{-1}{2} & \frac{-1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} ea \\ eb \\ ec \end{bmatrix} \qquad \text{(AIIB-1)}$$

like wise, the instantaneous space vectors i_o, i_α, and i_β on the 0-α-β co-ordinates are given as follows:

$$\begin{bmatrix} io \\ i\alpha \\ i\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ 1 & \frac{-1}{2} & \frac{-1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} \qquad \text{(AIIB-2)}$$

another instantaneous power p_o which is defined as the product of instantaneous space vectors, e_o and i_o on the zero axis:

$$p0 = e0 * i0 \qquad \text{(AIIB-3)}$$

$$\begin{bmatrix} p0 \\ p \\ q \end{bmatrix} = \begin{bmatrix} eo & 0 & 0 \\ 0 & e\alpha & e\beta \\ 0 & -e\beta & e\alpha \end{bmatrix} \begin{bmatrix} i0 \\ i\alpha \\ i\beta \end{bmatrix} \qquad \text{(AIIB-4)}$$

$$\begin{bmatrix} i0 \\ i\alpha \\ i\beta \end{bmatrix} = \begin{bmatrix} eo & 0 & 0 \\ 0 & e\alpha & e\beta \\ 0 & -e\beta & e\alpha \end{bmatrix}^{-1} \begin{bmatrix} p0 \\ 0 \\ 0 \end{bmatrix} + \begin{bmatrix} eo & 0 & 0 \\ 0 & e\alpha & e\beta \\ 0 & -e\beta & e\alpha \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ p \\ 0 \end{bmatrix} +$$

$$\begin{bmatrix} eo & 0 & 0 \\ 0 & e\alpha & e\beta \\ 0 & -e\beta & e\alpha \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ 0 \\ q \end{bmatrix}$$

$$\begin{bmatrix} i0 \\ i\alpha \\ i\beta \end{bmatrix} = \begin{bmatrix} i0 \\ 0 \\ 0 \end{bmatrix} + \begin{bmatrix} 0 \\ i\alpha p \\ i\beta p \end{bmatrix} + \begin{bmatrix} 0 \\ i\alpha q \\ i\beta q \end{bmatrix} \qquad \text{(AIIB-5)}$$

from equation (e) the instantaneous currents on the a-b-c coordinates are divided in the following three components, respectively:

$$\begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & 1 & 0 \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i0 \\ 0 \\ 0 \end{bmatrix}$$

$$+ \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & 1 & 0 \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} 0 \\ i\alpha p \\ i\beta p \end{bmatrix}$$

$$+ \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & 1 & 0 \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{-1}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} 0 \\ i\alpha q \\ i\beta q \end{bmatrix} \qquad \text{((AIIB-6))}$$

$$= \begin{bmatrix} ia0 \\ ib0 \\ ic0 \end{bmatrix} + \begin{bmatrix} iap \\ ibp \\ icp \end{bmatrix} + \begin{bmatrix} iaq \\ ibq \\ icq \end{bmatrix}$$

where

$$ia0 = ib0 = ic0 = i0/\sqrt{3}$$

let the a, b, c phase instantaneous powers be p_a, p_b, and p_c respectively. by applying the equation (f), following is obtained:

$$\begin{bmatrix} pa \\ pb \\ pc \end{bmatrix} = \begin{bmatrix} ea * ia0 \\ eb * ib0 \\ ec * ic0 \end{bmatrix} + \begin{bmatrix} ea * iap \\ eb * ibp \\ ec * icp \end{bmatrix} + \begin{bmatrix} ea * iaq \\ eb * ibq \\ ec * icq \end{bmatrix}$$

$$\begin{bmatrix} pa \\ pb \\ pc \end{bmatrix} = \begin{bmatrix} pa0 \\ pb0 \\ pc0 \end{bmatrix} + \begin{bmatrix} pap \\ pbp \\ pcp \end{bmatrix} + \begin{bmatrix} paq \\ pbq \\ pcq \end{bmatrix}$$

the instantaneous reactive power in each phase P_{aq}, p_{bq} and p_{cq} make no contribution to the instantaneous power flow in the three phase circuit which is represented by p_o and p, because the sum of the instantaneous power is always zero.

Annexure II-C
Results of Shunt Active Power Filter

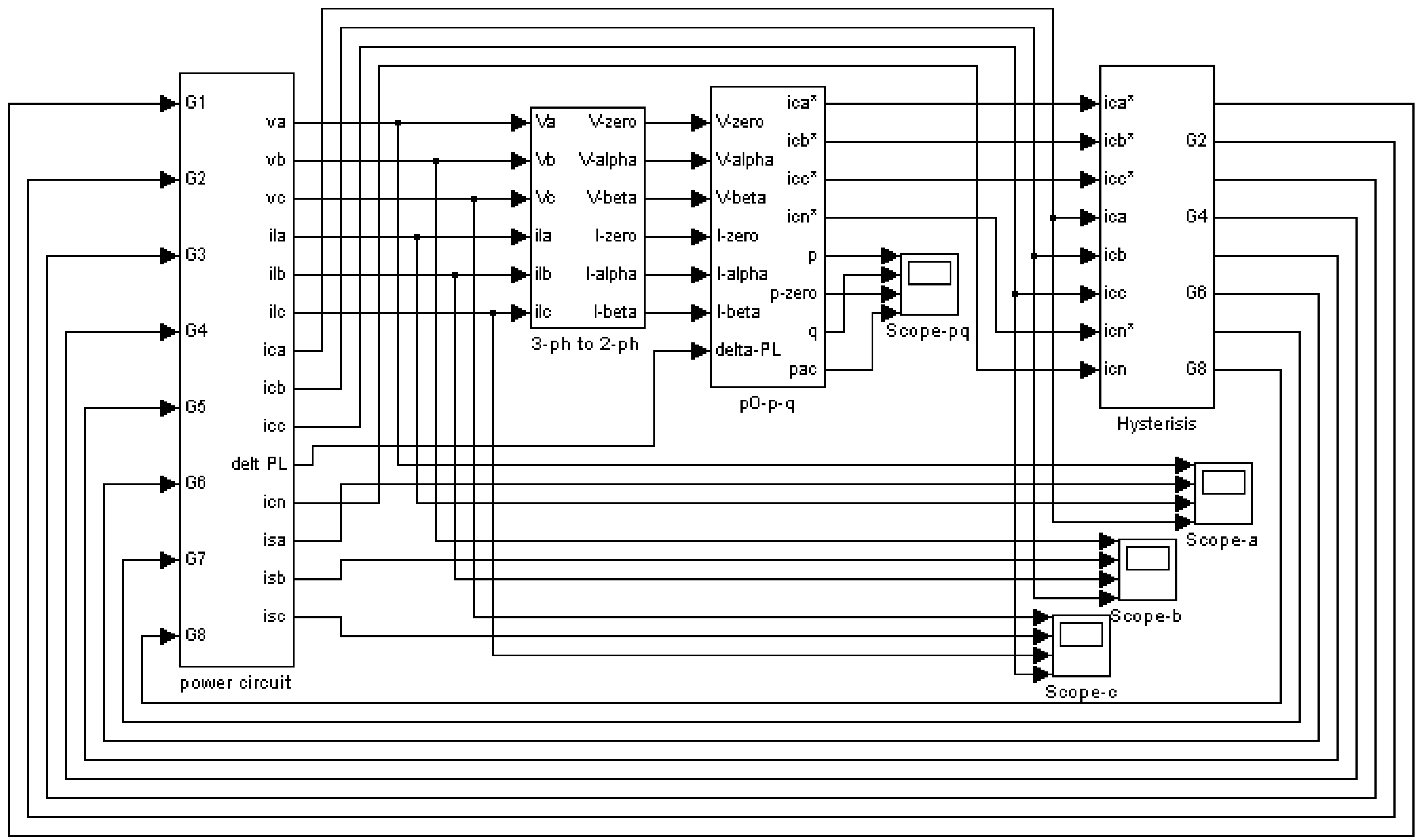
RESULT

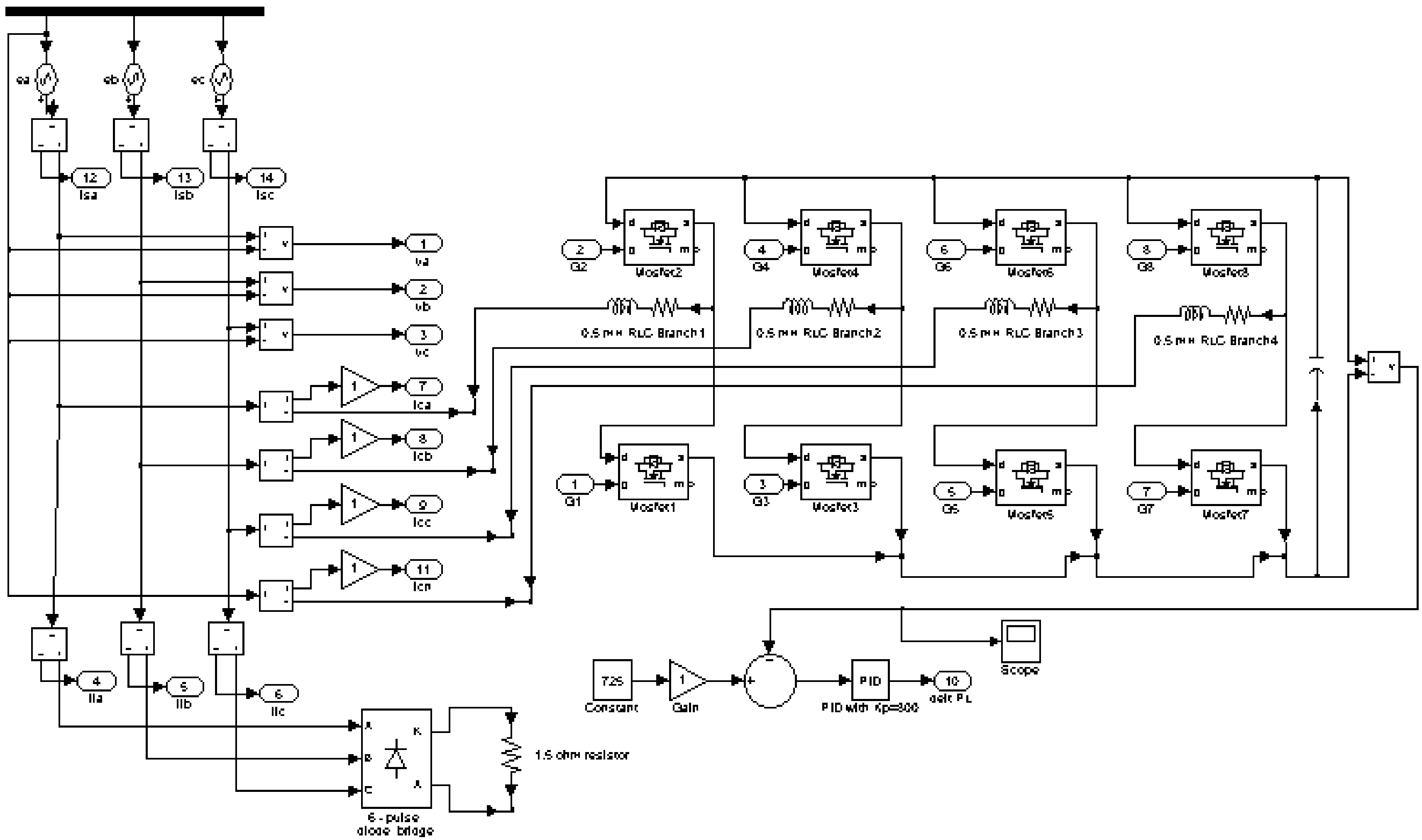
The simulations as well as experimental results are given here.

Simulation results

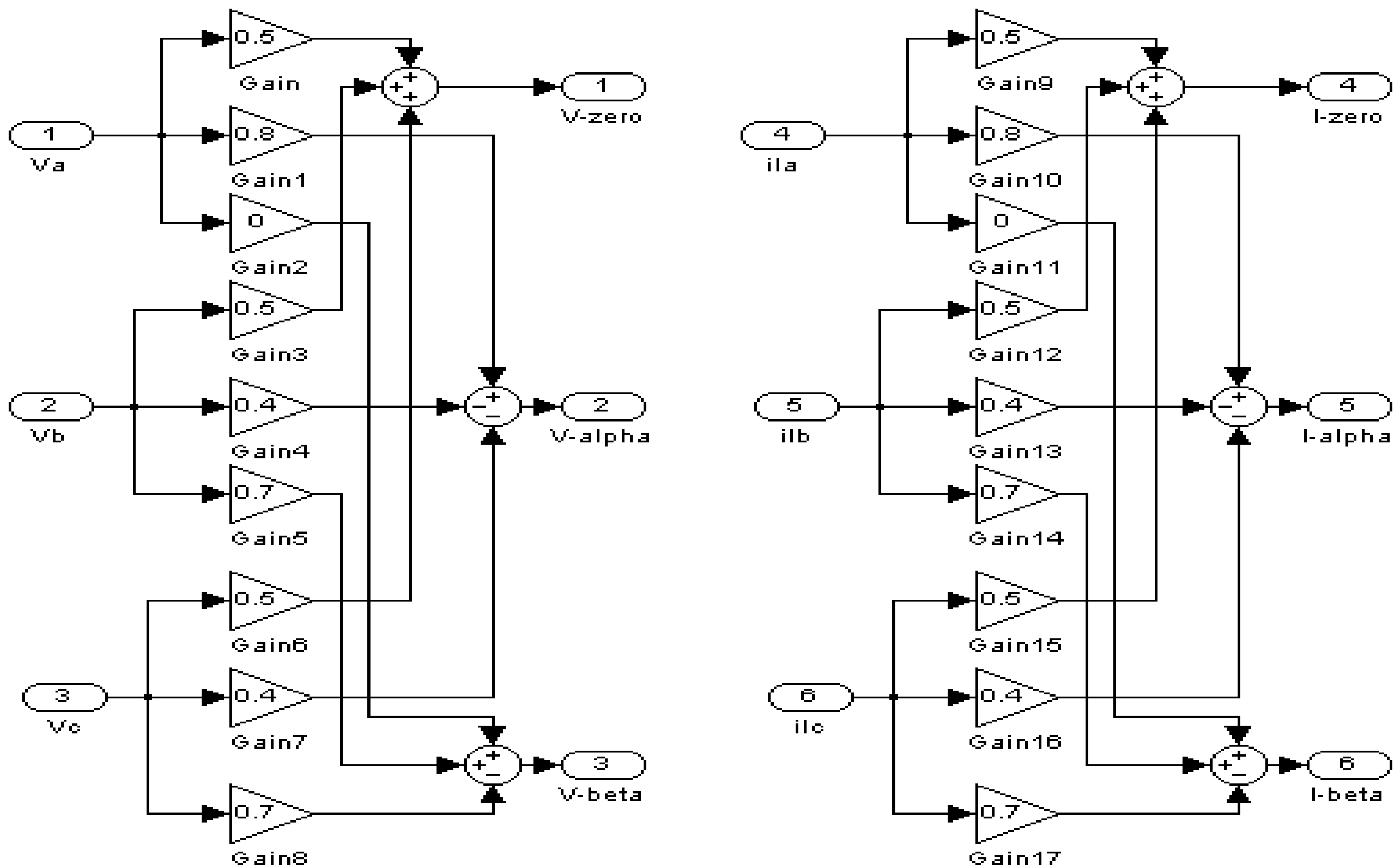
A complete model of the shunt active filter for 5 amps, 800V DC with input voltage of 440V and 5-amp load was implemented MATLAB (SIMULINK) and the most important results will be presented to compare actual and simulated results. The fundamental frequency of the system is 50 Hz. The wave-form of source voltage, source current, load current and compensating current is observed for a-phase, b-phase, & c-phase using scope available in the simulation tools. These results show that total harmonic current required by the load is supplied by the shunt active filter and absent from the source. Result are shown at the end of this chapter

ACTIVE FILTER SIMULATION BLOCK DIAGRAM

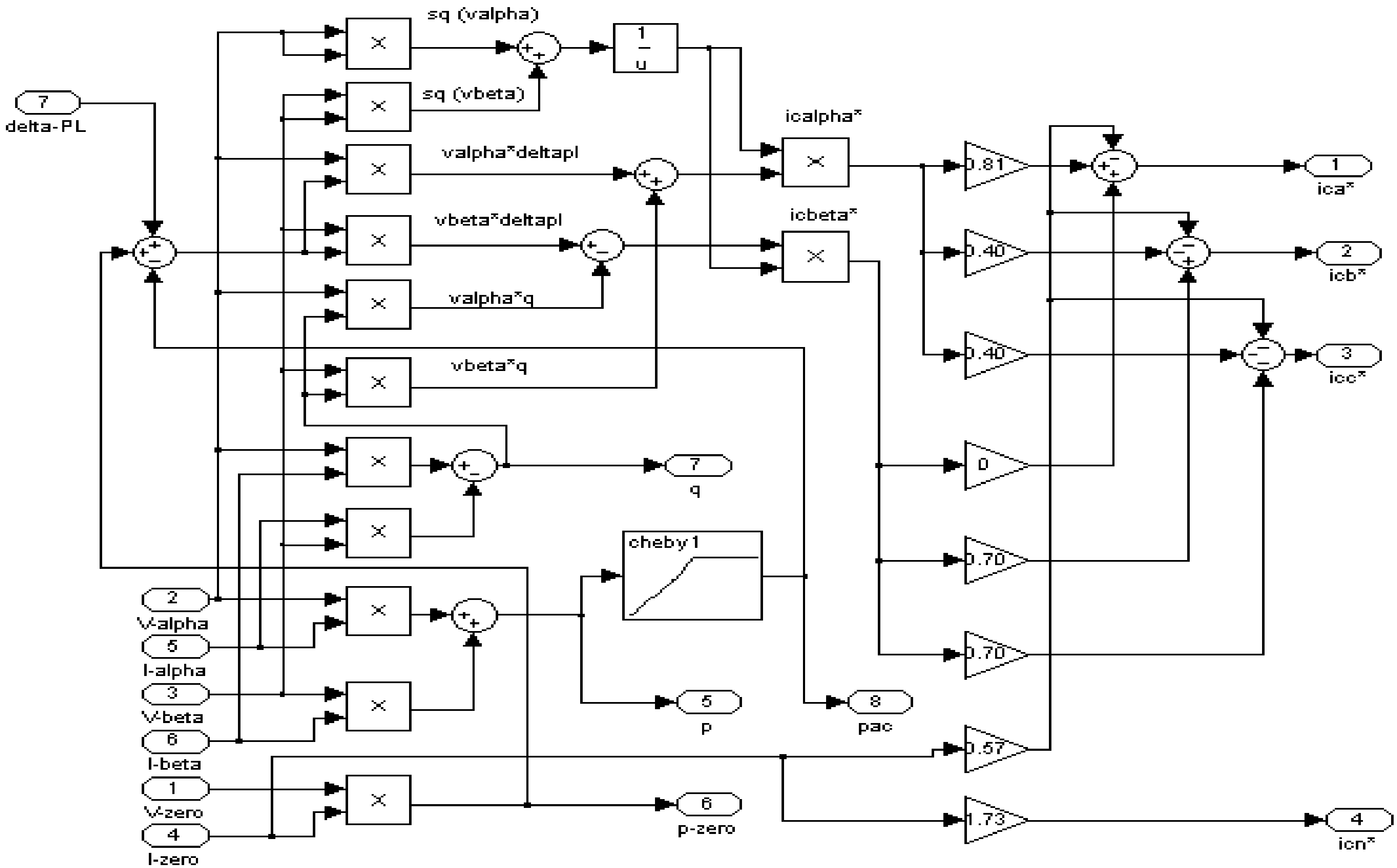




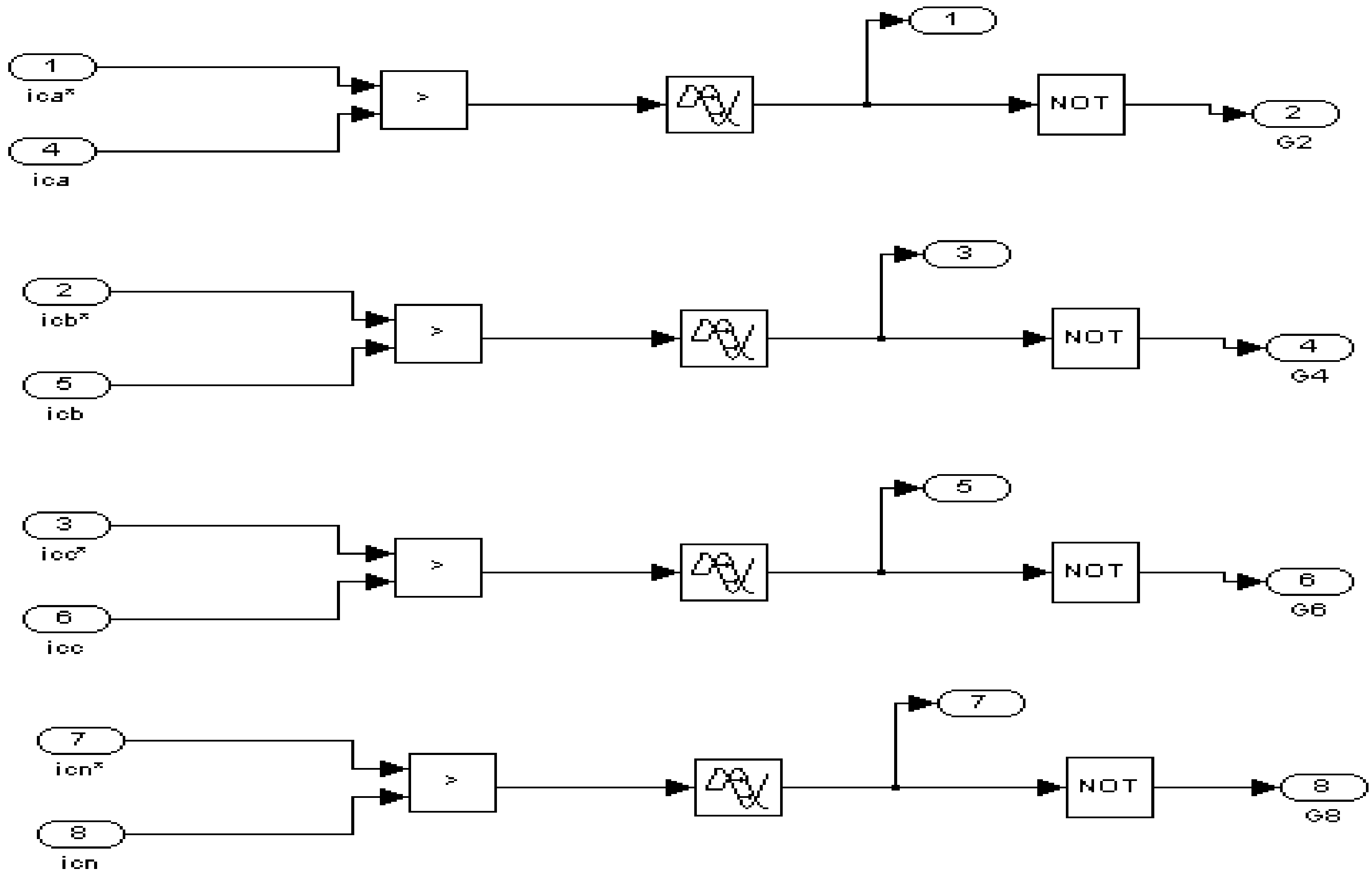
ACTIVE FILTER POWER CIRCUIT DIAGRAM FOR SIMULATION



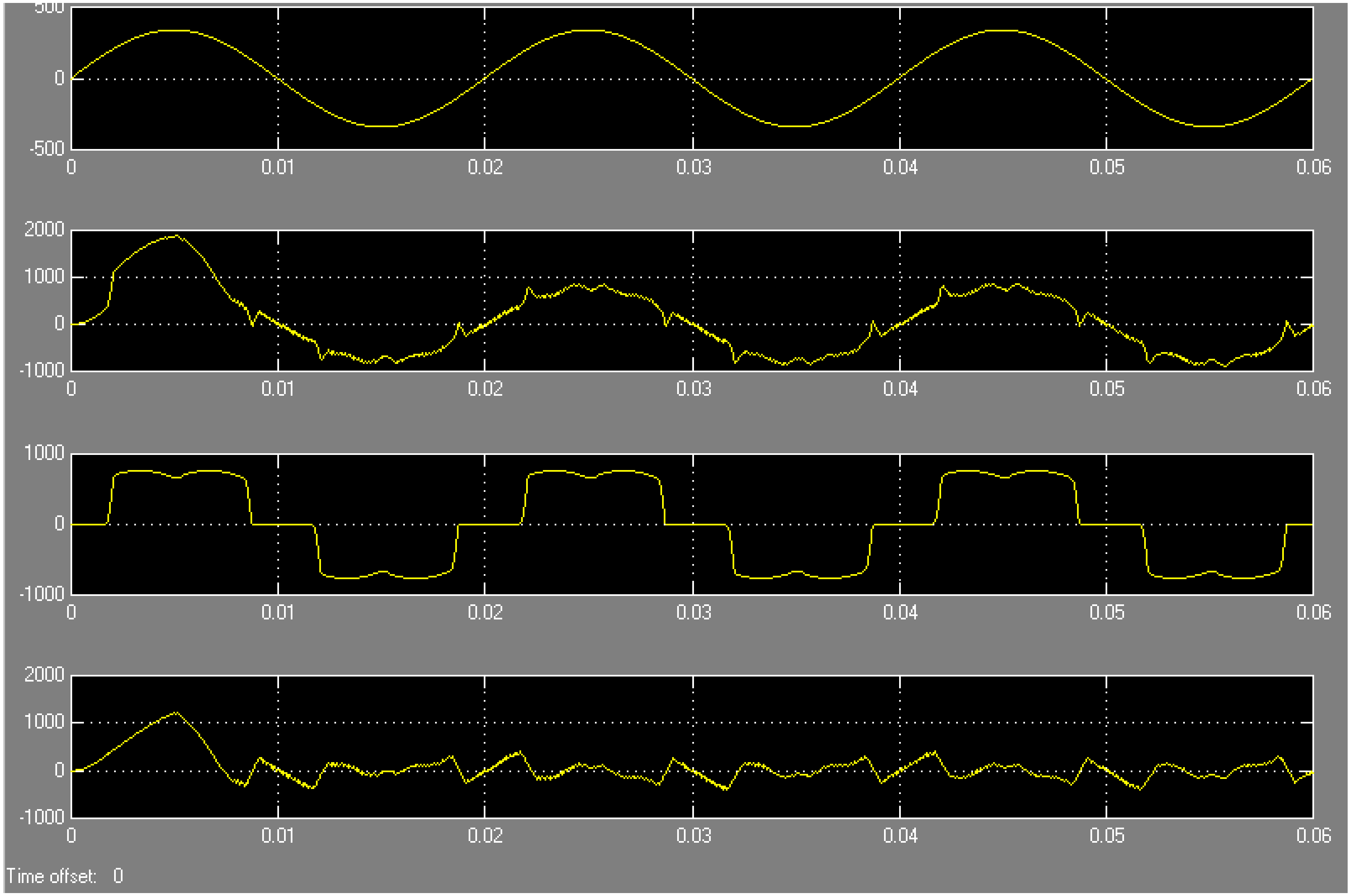
THREE PHASE TO ALPHA-BETA TRANSFORMATION



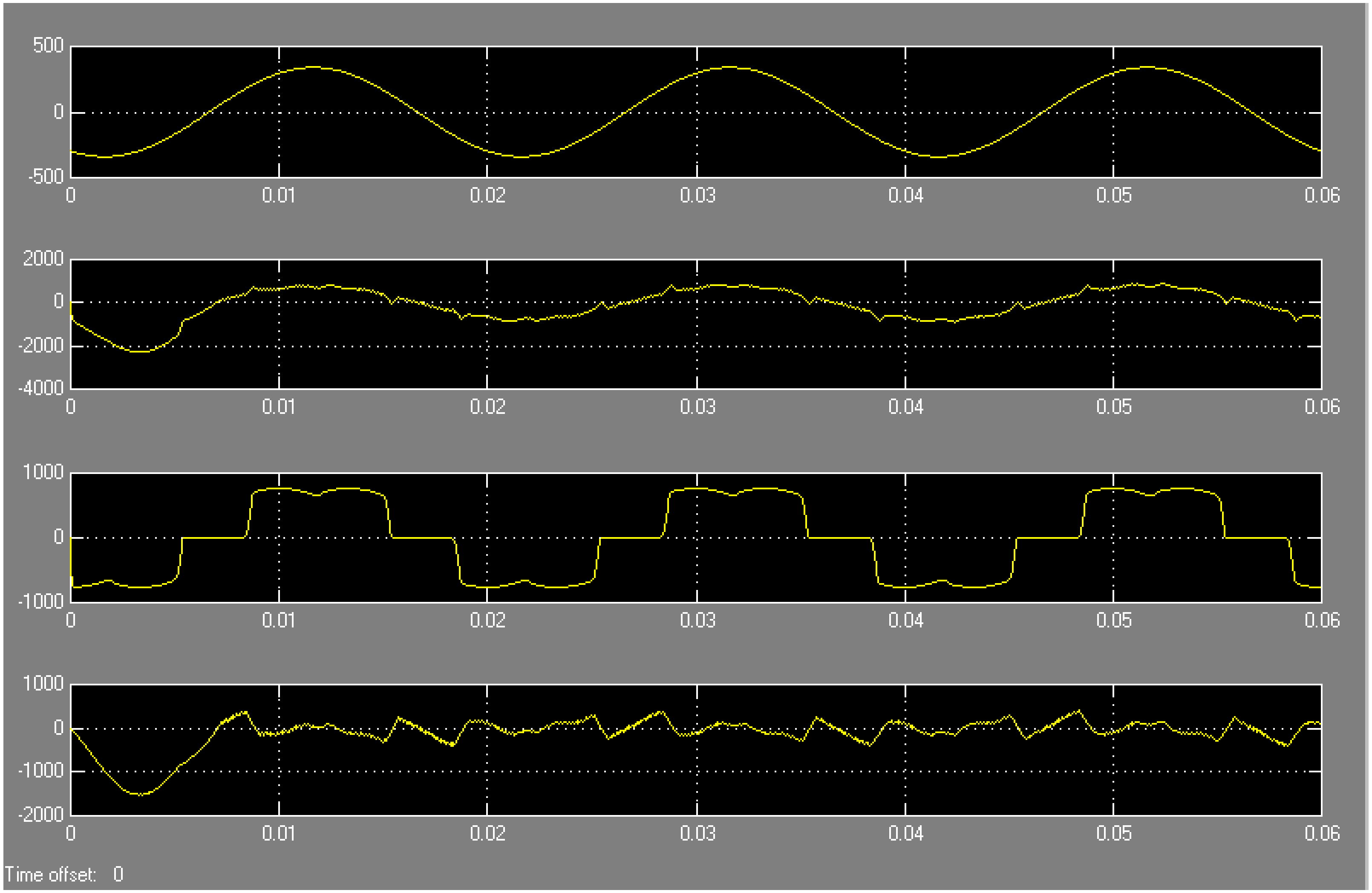
POWER CALCULATION AND COMPENSATING REFERANCE CURRENT GENERATION



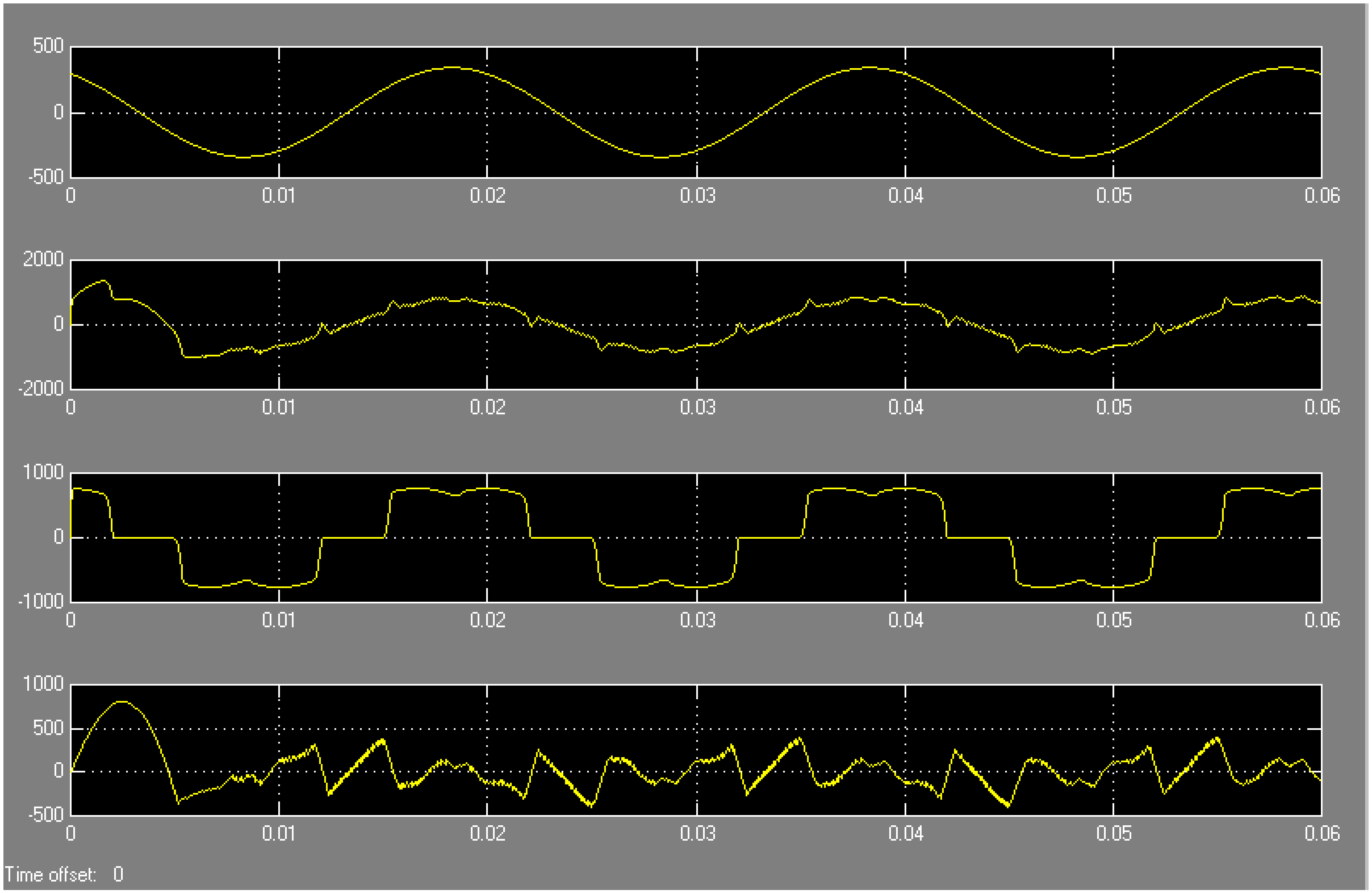
HYSTERISIS CURRENT CONTROLLER



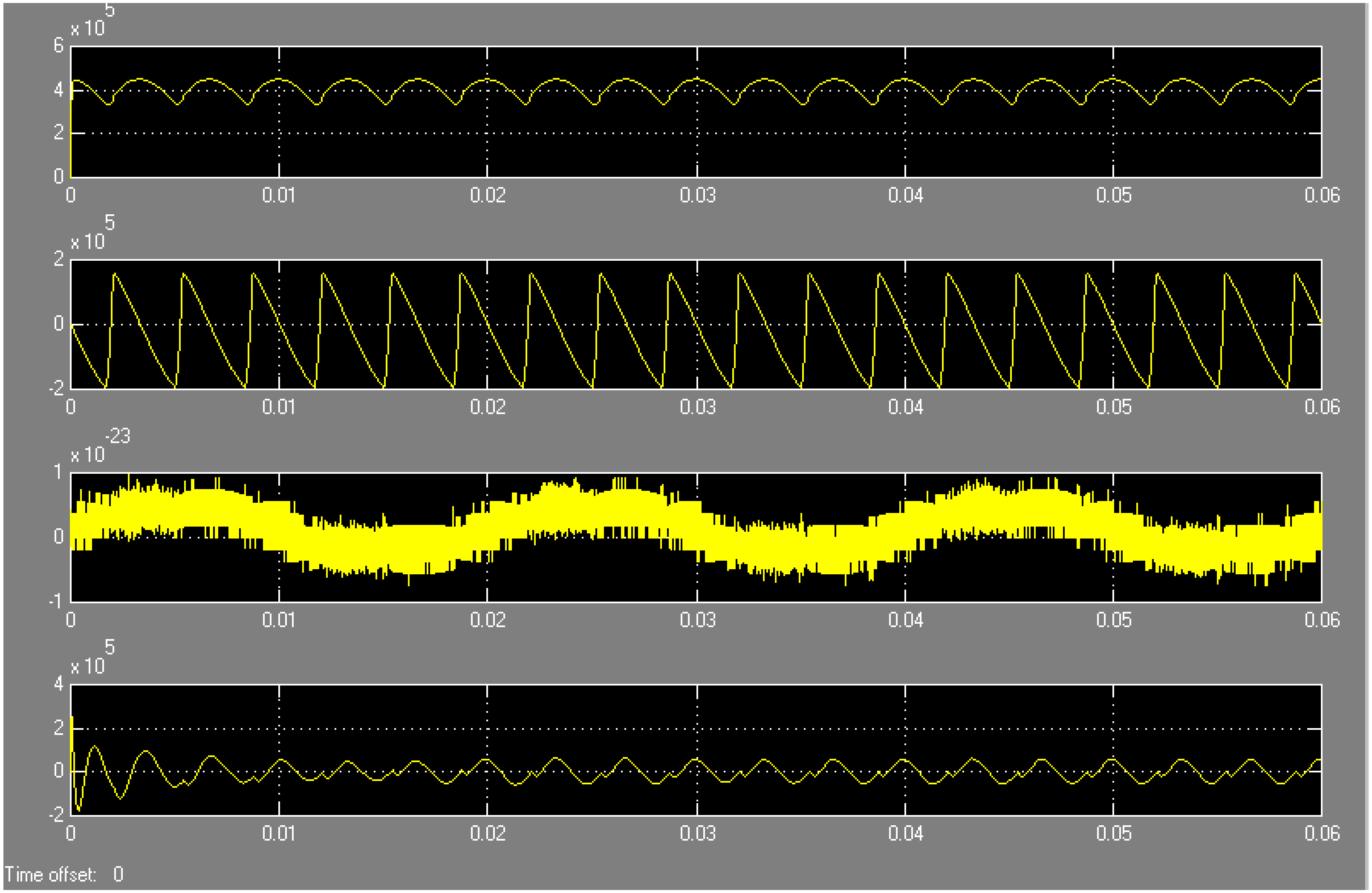
SOURCE -VOLTAGE, SOURCE –CURRENT, LOAD-CURRENT AND COMPENSATING-CURRENT FOR A-PHASE



SOURCE -VOLTAGE, SOURCE –CURRENT, LOAD-CURRENT AND COMPENSATING-CURRENT FOR B-PHASE

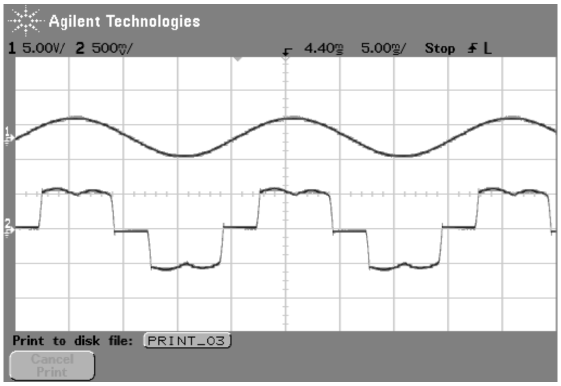


SOURCE -VOLTAGE, SOURCE –CURRENT, LOAD-CURRENT AND COMPENSATING-CURRENT FOR C-PHASE

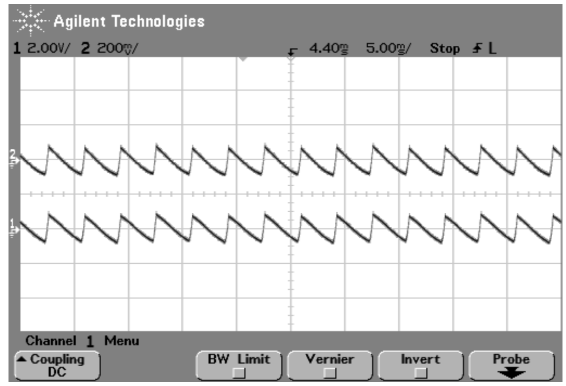


p-ACTIVE POWER, q-REACTIVE POWER, p_0 AND P_{dc} FOR ACTIVE POWER FILTER

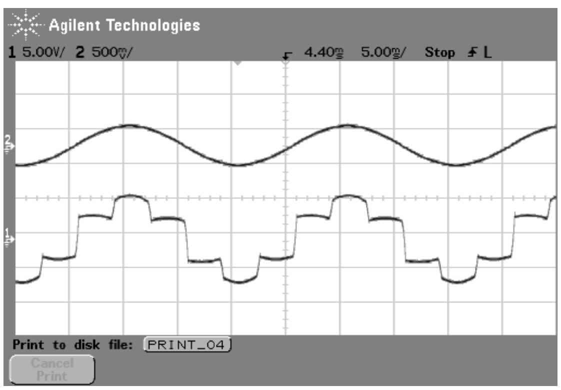
EXPERIMENTAL RESULTS:



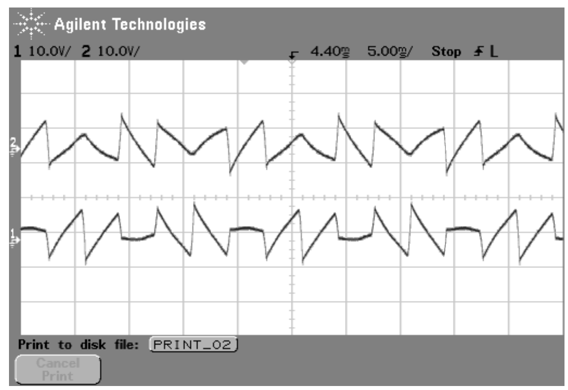
CONTROL WAVEFORM FOR v-ALPHA AND i-ALPHA



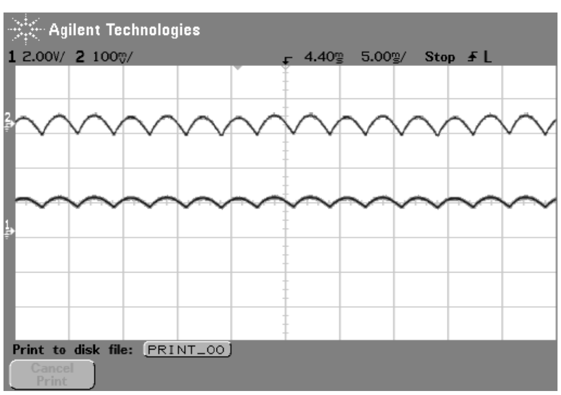
CONTROL WAVEFORM FOR qac AND q (REACTIVE POWER)



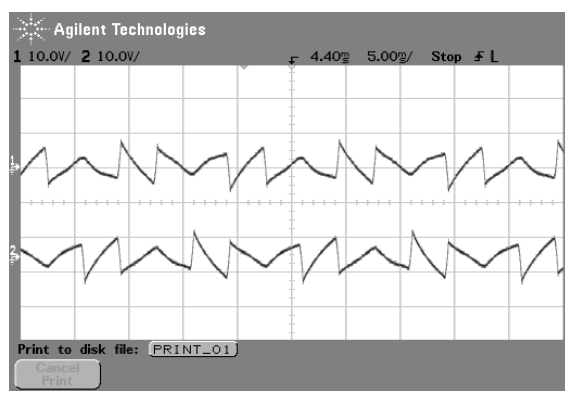
CONTROL WAVEFORM FOR v-BETA AND i-BETA



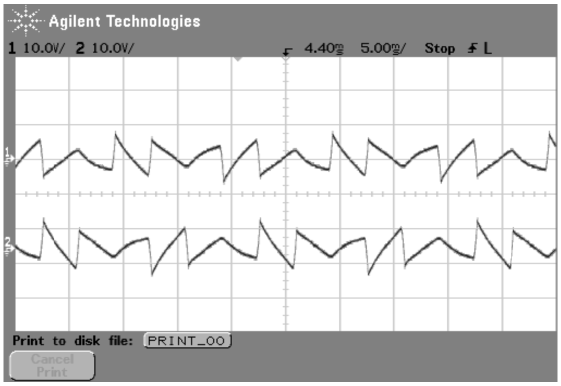
CONTROL WAVEFORM FOR ic-ALPHA & ic-BETA



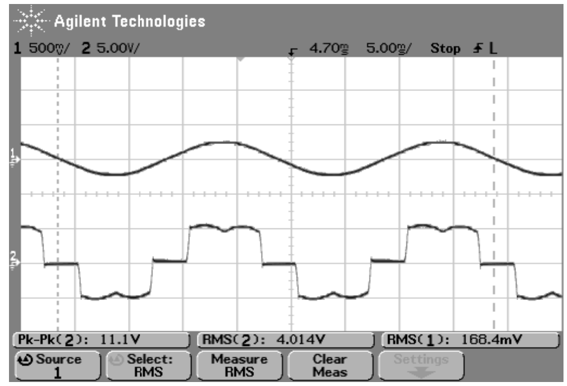
CONTROL WAVEFORM FOR pac AND p (POWER)



CONTROL WAVEFORM FOR ica, & icb

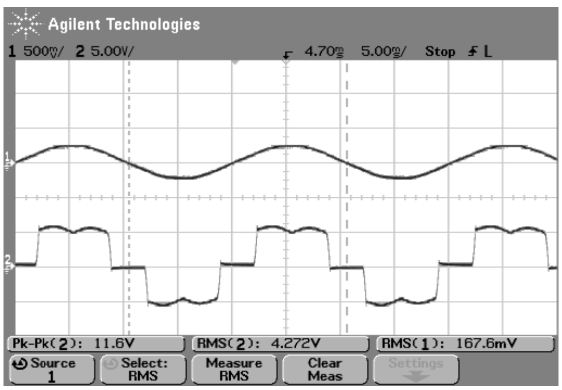


CONTROL WAVEFORM FOR ica, AND icc



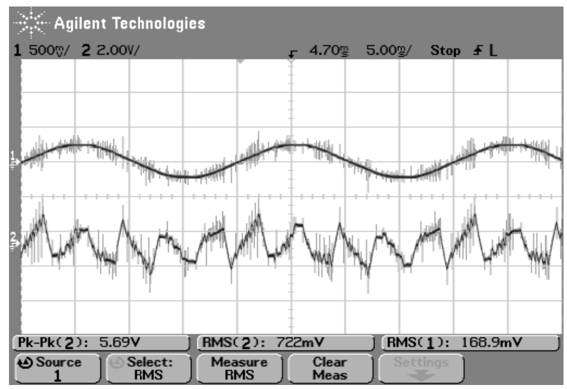
ACTUAL WAVEFORM FOR SOURCE VOLTAGE

AND LOAD CURRENT FOR C-PHASE



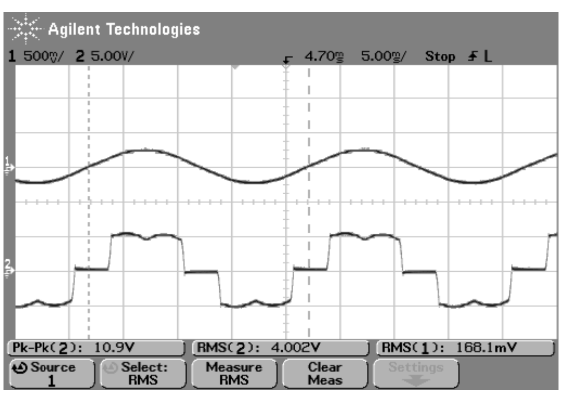
ACTUAL WAVEFORM FOR SOURCE VOLTAGE

AND LOAD CURRENT FOR A-PHASE



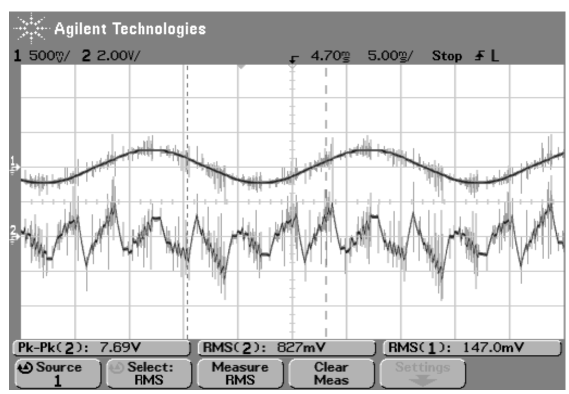
ACTUAL WAVEFORM FOR SOURCE VOLTAGE

AND COMPENSATING CURRENT FOR A-PHASE



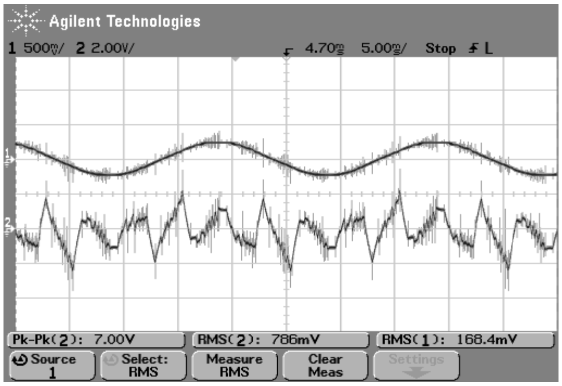
ACTUAL WAVEFORM FOR SOURCE VOLTAGE

AND LOAD CURRENT FOR B-PHASE

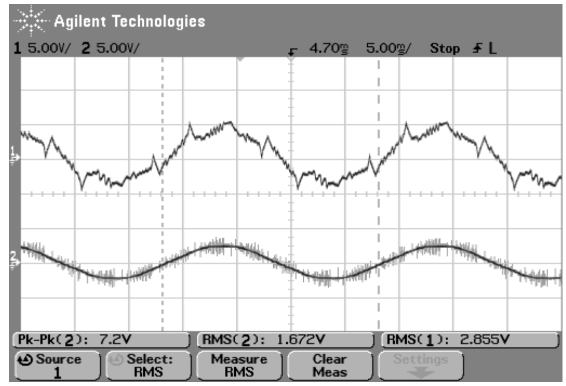


ACTUAL WAVEFORM FOR SOURCE VOLTAGE

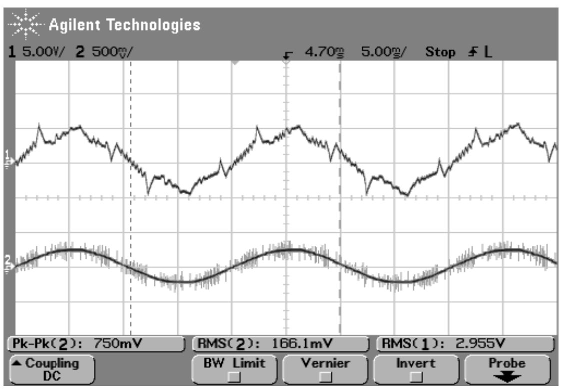
AND COMPENSATING CURRENT FOR B-PHASE



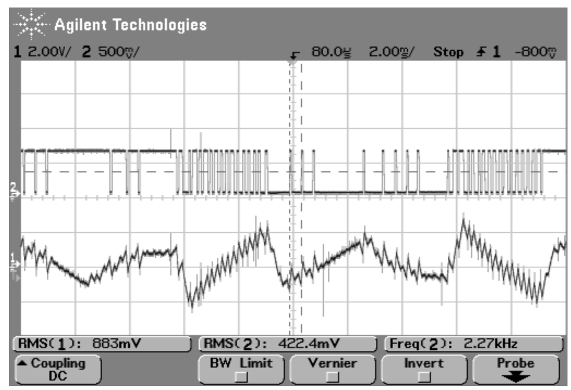
ACTUAL WAVEFORM FOR SOURCE VOLTAGE AND COMPENSATING CURRENT FOR C-PHASE



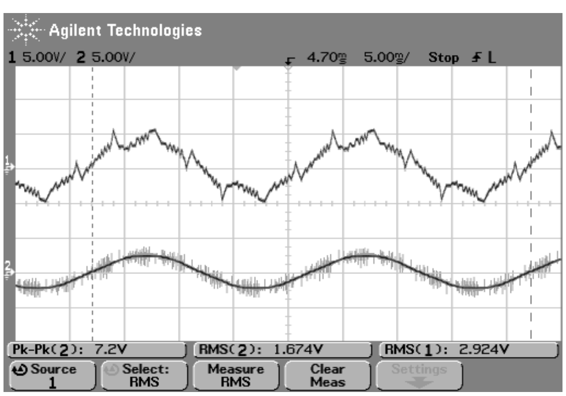
ACTUAL WAVEFORM FOR SOURCE CURRENT AND SOURCE VOLTAGE FOR PHASE- C



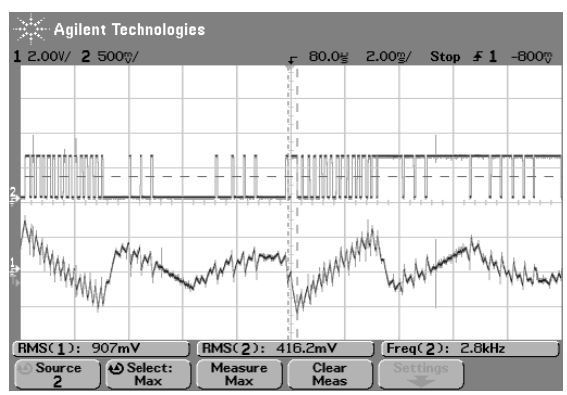
ACTUAL WAVEFORM FOR SOURCE CURRENT AND SOURCE VOLTAGE FOR PHASE-A



ACTUAL WAVEFORM FOR Vce AND COMPENSATING CURRENT FOR A-PHASE UPPER IGBT



ACTUAL WAVEFORM FOR SOURCE CURRENT AND SOURCE VOLTAGE FOR PHASE-B

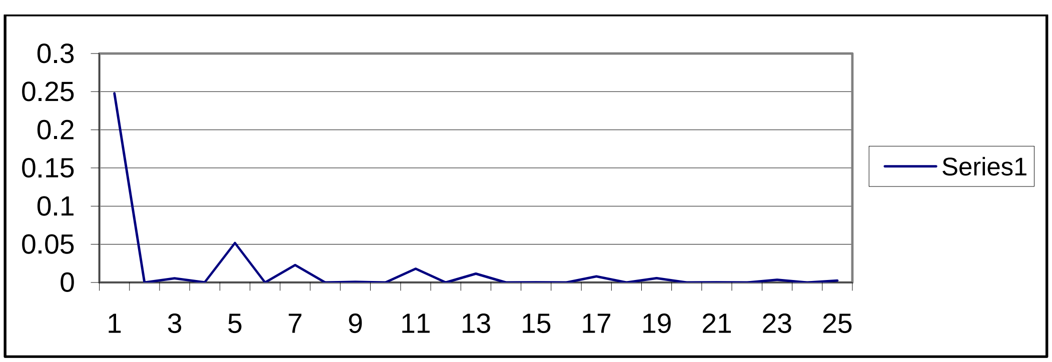


ACTUAL WAVEFORM FOR Vce AND COMPENSATING CURRENT FOR A-PHASE LOWER IGBT

FFT ANALYSIS OF SOURCE CURRENT WITHOUT COMPENSATING

(CT PROBE RATIO 10)

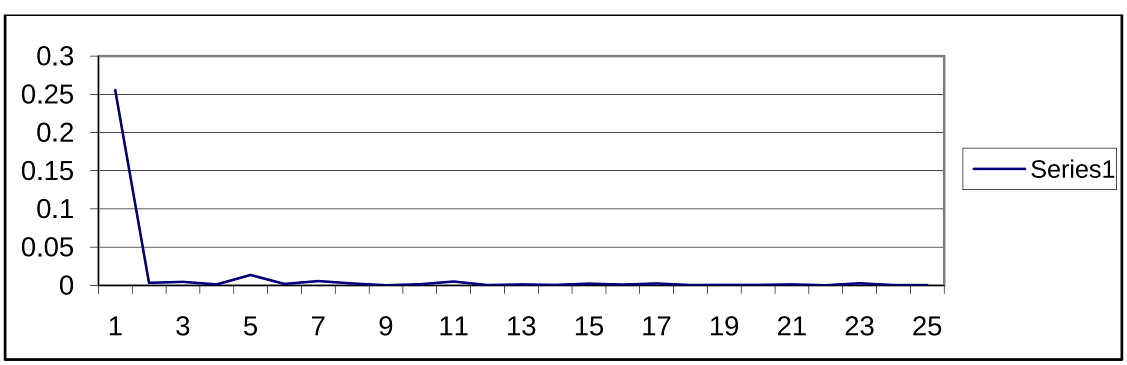
Fundamental	50Hz	
Harm no.	Voltage magnitude Volts	Voltage %
1	0.2478	100.00
2	0.0001	0.04
3	0.0054	2.18
4	0.0002	0.08
5	0.0517	20.86
6	0.0001	0.04
7	0.0228	9.20
8	0.0001	0.04
9	0.0008	0.32
10	0	0.00
11	0.018	7.26
12	0	0.00
13	0.0115	4.64
14	0	0.00
15	0.0003	0.12
16	0.0001	0.04
17	0.0079	3.19
18	0	0.00
19	0.0056	2.26
20	0	0.00
21	0.0002	0.08
22	0.0001	0.04
23	0.0035	1.41
24	0	0.00
25	0.0026	1.05
	THD%=	24.85



FFT ANALYSIS OF SOURCE CURRENT WITH COMPENSATION.

(CT PROBE RATIO 10)

Fundamental Frequency= 50 Hz		
Harm no.	Voltage magnitude Volts	Voltage %
1	0.2554	100.00
2	0.0035	1.37
3	0.0046	1.80
4	0.0014	0.55
5	0.0136	5.32
6	0.0019	0.74
7	0.0058	2.27
8	0.0025	0.98
9	0.0004	0.16
10	0.0015	0.59
11	0.0051	2.00
12	0.0006	0.23
13	0.0014	0.55
14	0.0008	0.31
15	0.0024	0.94
16	0.0011	0.43
17	0.0027	1.06
18	0.0005	0.20
19	0.0008	0.31
20	0.0007	0.27
21	0.0013	0.51
22	0.0003	0.12
23	0.0028	1.10
24	0.0005	0.20
25	0.0006	0.23
	THD%=	7.02



Annexure III

PHASE SEQUENCE CALCULATION FOR INDUCTION MOTOR

Equivalent circuit for 10 kW, 50 Hz, 415 Volts, thee phase delta connected Induction motor is given below to calculate the effect of unbalance voltage on Induction motor. The value are obtained by conducting the No Load Test, Load Test & Block Rotor Test at ERDA Laboratory,

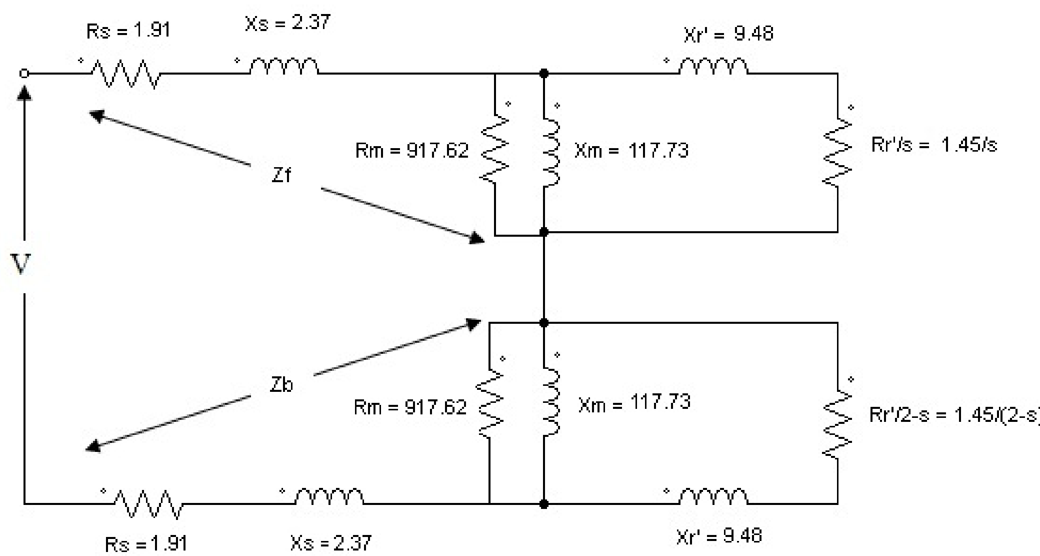


Figure AIII-1 Per phase equivalent circuit of Induction motor for calculation of speed torque characteristics under unbalance stator voltage condition.

The 10 kW motor performance characteristics was calculated with input rms voltage having unbalance as below in Table AIII-1:

Table AIII-1: table of unbalance voltage used for calculation

V_a	V_a angle	V_b	V_b angle	V_c	V_c angle
415.00	0.00	415.00	-120.00	215.00	120.00

Sequence voltage was calculated from the input voltage as mentioned in Table AIII-1 which is given below in Table AIII-2 for Zero Sequence, Positive Sequence and Negative Sequence voltage.

Table AIII-2: table of calculated sequence voltage & its percentage w.r.t positive sequence

V_o	V_+	V_-
33.33-57.73j	348.33-3.31E-14j	33.33+57.73j
V_{omag}	V_{+mag}	V_{-mag}
66.67	348.33	66.67
$\%V_o/V_p$		$\%V_n/V_p$
19.14		19.13875598

With sequence voltage & equivalent circuit of Induction motor for Positive Sequence, Negative Sequence & Zero Sequence which is given below, the speed torque characteristics of Induction Motor was obtained.

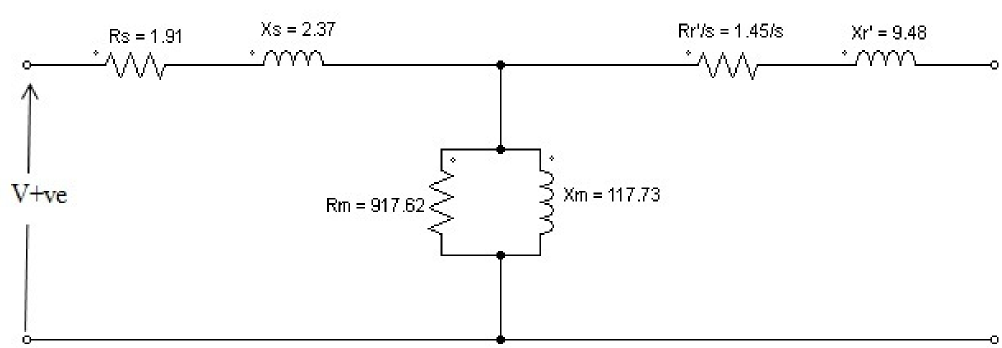


Figure III-2 Per phase equivalent circuit of Induction motor for calculation of speed torque characteristics under +ve sequence voltage.

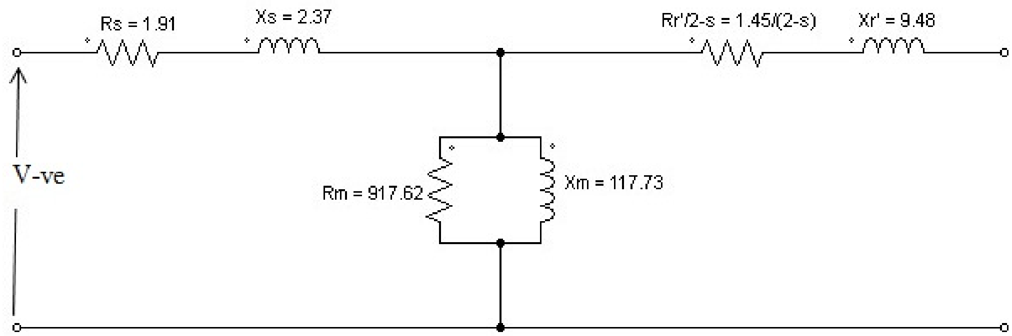


Figure III-3 Per phase equivalent circuit of Induction motor for calculation of speed torque characteristics under -ve sequence voltage.

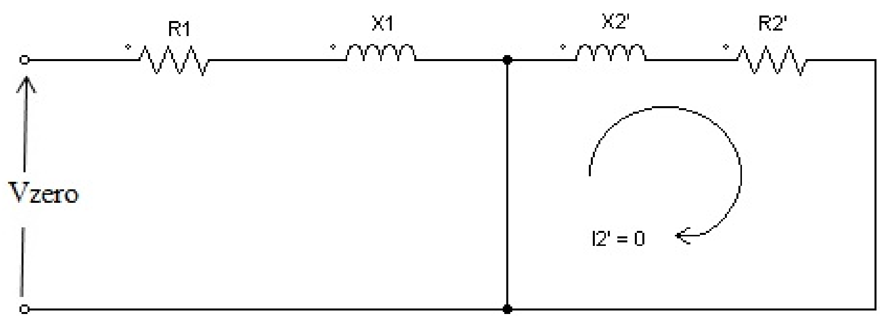


Figure III-4 Per phase equivalent circuit of Induction motor for calculation of speed torque characteristics under Zero sequence voltage.

The detail of calculation speed & torque at different slips with resolution of 0.005 are given below for balance voltage of 415 Volts

1	2	3	4		6	7	8	9	10	11	12	13	14	15	16	17
Voltage	**Avg. Cur.**	**Input W**	**N50**	**Slip**	**Rotor output**	**Eff.**	**P.F.**	**Eff.*p.f.**	**Stator Cu. Loss**	**Rotor Cu loss**	**Rotor output**	**Stray loss**	**Losses**	**Wm**	**Shaft power**	**Shaft Torque (Balance voltage)**
415.0	6.0	630.8	1499.9	0.000	22.7	3.6	0.1	0.5	69.4	0.0	22.7	31.5	100.9	157.1	22.7	0.1
415.0	6.8	2300.3	1492.5	0.005	1672.0	72.7	0.5	34.3	88.3	8.4	1672.0	31.5	128.2	156.3	1672.0	10.7
415.0	8.3	3958.8	1485.0	0.010	3269.7	82.6	0.7	54.9	131.8	33.0	3269.7	31.5	196.3	155.5	3269.7	21.0
415.0	10.2	5569.0	1477.5	0.015	4781.6	85.9	0.8	65.5	198.2	72.8	4781.6	31.5	302.5	154.7	4781.6	30.9
415.0	12.2	7118.7	1470.0	0.020	6198.4	87.1	0.8	70.7	285.5	126.5	6198.4	31.5	443.5	153.9	6198.4	40.3
415.0	14.3	8597.5	1462.5	0.025	7513.2	87.4	0.8	73.2	391.5	192.6	7513.2	31.5	615.6	153.2	7513.2	49.1
415.0	16.4	9997.3	1455.0	0.030	8721.9	87.2	0.9	74.2	513.8	269.7	8721.9	31.5	815.1	152.4	8721.9	57.2
415.0	**18.4**	**11312.4**	**1447.5**	**0.035**	**9822.3**	**86.8**	**0.9**	**74.3**	**650.2**	**356.2**	**9822.3**	**31.5**	**1037.9**	**151.6**	**9822.3**	**64.8**
415.0	**20.4**	**12539.0**	**1440.0**	**0.040**	**10814.7**	**86.2**	**0.9**	**73.8**	**798.2**	**450.6**	**10814.7**	**31.5**	**1280.3**	**150.8**	**10814.7**	**71.7**
415.0	32.2	18663.2	1387.5	0.075	15031.1	80.5	0.8	65.0	1988.6	1218.7	15031.1	31.5	3238.8	145.3	15031.1	103.4
415.0	38.4	20839.2	1350.0	0.100	15857.2	76.1	0.8	57.6	2822.6	1761.9	15857.2	31.5	4615.9	141.4	15857.2	112.2
415.0	**42.2**	**21694.5**	**1320.0**	**0.120**	**15752.6**	**72.6**	**0.7**	**52.0**	**3412.9**	**2148.1**	**15752.6**	**31.5**	**5592.5**	**138.2**	**15752.6**	**114.0**
415.0	51.3	21501.2	1200.0	0.200	12888.5	59.9	0.6	35.0	5045.5	3222.1	12888.5	31.5	8299.1	125.7	12888.5	102.6
415.0	56.2	19340.7	1050.0	0.300	9071.8	46.9	0.5	22.5	6050.0	3887.9	9071.8	31.5	9969.4	110.0	9071.8	82.5
415.0	58.5	17441.0	900.0	0.400	6335.9	36.3	0.4	15.1	6554.1	4223.9	6335.9	31.5	10809.5	94.2	6335.9	67.2
415.0	59.7	16000.8	750.0	0.500	4416.4	27.6	0.4	10.3	6841.6	4416.4	4416.4	31.5	11289.5	78.5	4416.4	56.2
415.0	60.5	14912.4	600.0	0.600	3025.3	20.3	0.3	7.0	7022.5	4538.0	3025.3	31.5	11592.0	62.8	3025.3	48.1
415.0	61.0	14072.7	450.0	0.700	1980.2	14.1	0.3	4.5	7144.9	4620.5	1980.2	31.5	11796.8	47.1	1980.2	42.0
415.0	61.4	13409.5	300.0	0.800	1169.9	8.7	0.3	2.7	7232.3	4679.6	1169.9	31.5	11943.3	31.4	1169.9	37.2
415.0	61.7	12874.3	150.0	0.900	524.9	4.1	0.3	1.2	7297.4	4723.7	524.9	31.5	12052.5	15.7	524.9	33.4
415.0	61.9	12434.3	0.0	1.000	0.0	0.0	0.3	0.0	7347.6	4757.7	0.0	31.5	12136.8	0.0	0.0	30.3

As mentioned above same calculation was done with unbalance voltage as mention in Table AIII-1: table of unbalance voltage used for calculation.

2			3	4	5.000	6	7	8	9	10	11	12	13	14	15	16	17
Current Pos Seq.	**Current Neg. Seq.**	**Current Zero Seq.**	**Input power**	**Speed**	**Slip**	**Rotor Output**	**eff.**	**P.F.**	**Eff.*p.f.**	**Stator Cu. Loss**	**Rotor Cu loss**	**Rotor output**	**Stray loss**	**Losses**	**Wm**	**Shaft power**	**Shaft Torque (unbalance voltage)**
5.1	10.1	37.9	3459.5	1499.9	0.000	-47.2	-1.4	0.3	-0.5	2992.3	126.3	-47.2	44.7	3163.3	157.0	-47.2	-0.3
5.7	10.1	37.9	4635.9	1492.5	0.005	1114.9	24.0	0.4	10.7	3005.6	132.2	1114.9	44.7	3182.5	156.3	1114.9	7.1
7.0	10.1	37.9	5804.4	1485.0	0.010	2240.8	38.6	0.5	19.6	3036.2	149.6	2240.8	44.7	3230.5	155.5	2240.8	14.4
8.5	10.1	37.9	6939.1	1477.5	0.015	3306.1	47.6	0.5	25.8	3083.0	177.6	3306.1	44.7	3305.3	154.7	3306.1	21.4
10.2	10.1	37.9	8031.0	1470.0	0.020	4304.4	53.6	0.6	29.9	3144.5	215.4	4304.4	44.7	3404.6	153.9	4304.4	28.0
12.0	10.1	37.9	9072.9	1462.5	0.025	5230.9	57.7	0.6	32.6	3219.1	262.0	5230.9	44.7	3525.8	153.2	5230.9	34.2
13.7	10.1	37.9	10059.3	1455.0	0.030	6082.5	60.5	0.6	34.5	3305.3	316.3	6082.5	44.7	3666.3	152.4	6082.5	39.9
15.5	**10.1**	**37.9**	**10986.0**	**1447.5**	**0.035**	**6858.0**	**62.4**	**0.6**	**35.7**	**3401.4**	**377.3**	**6858.0**	**44.7**	**3823.3**	**151.6**	**6858.0**	**45.2**
17.1	**10.1**	**37.9**	**11850.2**	**1440.0**	**0.040**	**7557.3**	**63.8**	**0.6**	**36.5**	**3505.6**	**443.7**	**7557.3**	**44.7**	**3994.0**	**150.8**	**7557.3**	**50.1**
27.0	10.1	37.9	16165.9	1387.5	0.075	10529.0	65.1	0.6	36.2	4344.2	984.8	10529.0	44.7	5373.7	145.3	10529.0	72.5
32.2	10.1	37.9	17699.7	1350.0	0.100	11112.0	62.8	0.5	33.9	4931.7	1367.5	11112.0	44.7	6343.8	141.4	11112.0	78.6
35.4	**10.1**	**37.9**	**18303.0**	**1320.0**	**0.120**	**11039.0**	**60.3**	**0.5**	**31.7**	**5347.6**	**1639.5**	**11039.0**	**44.7**	**7031.7**	**138.2**	**11039.0**	**79.9**
43.1	10.1	37.9	18169.5	1200.0	0.200	9024.2	49.7	0.5	24.0	6497.5	2396.0	9024.2	44.7	8938.2	125.7	9024.2	71.8
47.2	10.1	37.9	16651.0	1050.0	0.300	6339.5	38.1	0.4	17.1	7204.9	2864.9	6339.5	44.7	10114.4	110.0	6339.5	57.7
49.1	10.0	37.9	15316.7	900.0	0.400	4416.7	28.8	0.4	12.4	7559.6	3101.3	4416.7	44.7	10705.7	94.2	4416.7	46.9
50.1	10.0	37.9	14306.7	750.0	0.500	3069.7	21.5	0.4	8.9	7761.8	3236.7	3069.7	44.7	11043.2	78.5	3069.7	39.1
50.8	10.0	37.9	13545.1	600.0	0.600	2095.7	15.5	0.4	6.3	7888.8	3322.0	2095.7	44.7	11255.5	62.8	2095.7	33.4
51.2	10.0	37.9	12959.5	450.0	0.700	1366.4	10.5	0.4	4.2	7974.4	3379.7	1366.4	44.7	11398.8	47.1	1366.4	29.0
51.6	10.0	37.9	12499.2	300.0	0.800	803.5	6.4	0.4	2.6	8035.3	3420.9	803.5	44.7	11500.9	31.4	803.5	25.6
51.8	10.0	37.9	12130.2	150.0	0.900	358.5	3.0	0.4	1.2	8080.3	3451.4	358.5	44.7	11576.4	15.7	358.5	22.8
52.0	9.9	37.9	11829.7	0.0	1.000	0.0	0.0	0.4	0.0	8114.7	3474.7	0.0	44.7	11634.0	0.0	0.0	20.6

Speed v/s Torque curve for balance voltage & unbalance voltage is shown below

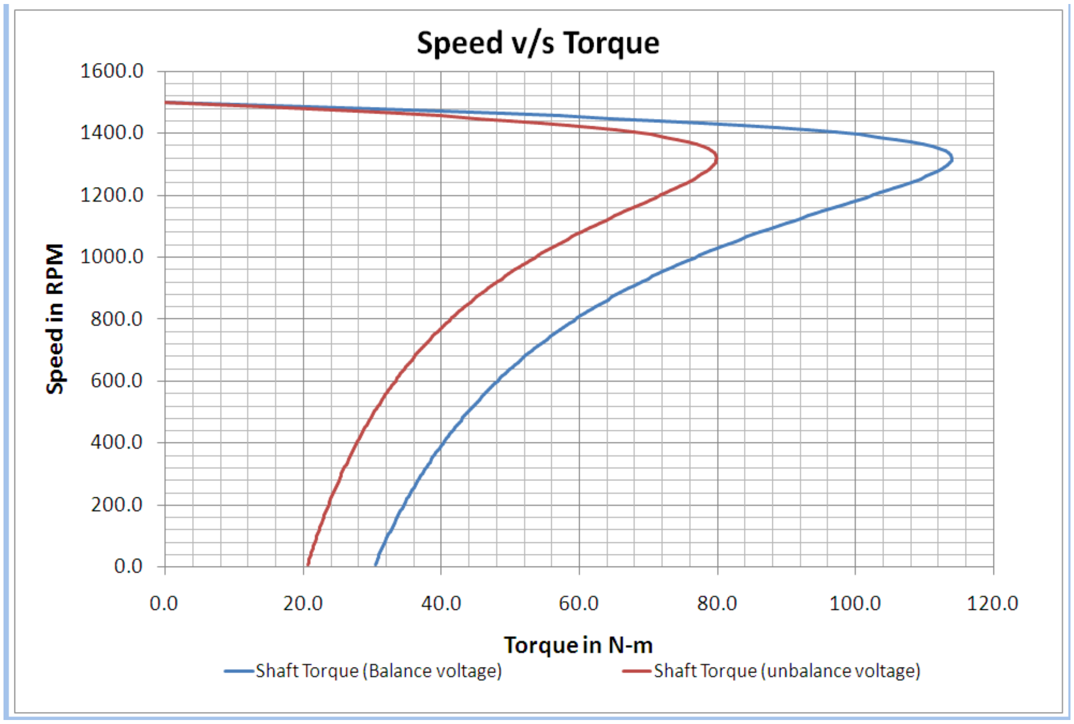


Figure III-5 Speed Torque Characteristics of Induction motor under balance voltage as well as unbalance voltage condition.

From above data it is clear that with unbalance voltages, losses in the motor are high as compared to if motor is supplied with balance voltage. For 10 kW, 4 poles, 3 phases, delta connected, 415 volts, three phase motor losses with balance voltage & unbalance voltage having negative sequence voltage of 20% & zero sequence voltage of 20 %, the losses calculated are as follows:

	Input power Watts	output Power Watts	Total Losses Watts	% Effi.	Total Saving Watts	% saving of input power
with balance voltage	11312	9822	1038	87	2785	28
with Unbalance voltage	10986	6858	3823	62		