PERFORMANCE INVESTIGATIONS ON NON-SINUSOIDAL SUPPLY FED INDUCTION MACHINE DRIVE

A DISSERTATION

Submitted in partial fulfilment of the requirements for the award of the degree

of

MASTER OF ENGINEERING

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ELECTRICAL ENGINEERING

(With Specialization in Power Apparatus and Electric Drives)

By

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CANDIDATE'S DECLARATION

I hereby declare that the work which is being presented in the dissertation entitled, **"PERFORMANCE INVESTIGATIONS ON NON-SINUSOIDAL SUPPLY FED INDUCTION MACHINE DRIVE"**, in partial fulfillment of the requirement for the award of degree of **MASTER OF ENGINEERING** in ELECTRICAL ENGINEERING with specialization in POWER APPARATUS AND ELECTRIC DRIVES, submitted in the Department of Electrical Engineering, **University of Roorkee, Roorkee (India)** is an authentic record of my own work carried out from August 1999 to March 2000 under the supervision of **Dr. S.P. Srivastava, Associate Professor and Dr. G.K. Singh, Assistant Professor** in the Department of Electrical Engineering, University of Roorkee, Roorkee.

The matter embodied in this dissertation has not been submitted by me for the award of any other degree or diploma.

Date : March 31 2000

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CERTIFICATE

This is to certify that the above statement made by the candidate is correct to the best of our knowledge.

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ABSTRACT

Static converters have lead to a revolutionary change in the field of the induction motor control. These static converter devices produce such types of output which are rich in harmonics. The effect of these harmonics on motor performances be not favourable. The motor currents and losses increase in significant amount. The induction motor performance is deteriorated due to the increase in the losses. Then it becomes essential to investigate the performance of the induction motor with these supplies. In the present work, the induction motor performance is calculated with both sinusoidal and non-sinusoidal. In the non-sinusoidal supplies the stepped voltage source inverters and different PWM techniques are used. A comparison between the performances obtained by both type of supplies is investigated.

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INTRODUCTION

1.1 Introduction

Developments in digital and power semiconductors have lead to a rapid increase in the use of nonlinear devices including adjustable speed drives. For this reason harmonics have become a common problem in many industrial electrical systems. In power system harmonic pollution problems have long been bothering power companies, manufacturers and customers. The originations of harmonics can be classified as follows:

1. Power electronics devices, with the recent advances in power semiconductor technology, more power electronics devices, such as phase controllers, inverters, cycloconverters, are widely used for electromechanical loads in the industry.
2. The application of saturable core reactors for isolating dc components when the load to be controlled doesn't require dc.
3. The operation of arc furnaces and electric arc welders, etc., these non-continuous loads result in significant current distortion and the appearance of even harmonics in the transmission and distribution systems.
4. Shunt capacitor banks for power factor correction and voltage regulation in a distribution system.
5. Using series inductors to reduce the short circuit current of a transmission line. Inductors and capacitors do not produce harmonics; however, they may result the potential for resonant problems, which can magnify existing harmonic levels.

As the power electronics field is concerned, developments have lead to an ever-increasing use of static switching devices to control the torque and speed of ac motors. Invariably the output voltage and current waveforms of these devices contain numerous harmonics. The order and magnitude of these harmonics depend on the design as well as the nature of the load being supplied. Generally the voltage source inverters, current source inverters and pulse width modulated inverters are used to control the speed of the induction motor. The output voltage of these static controllers deviates substantially from the sinusoidal form and contains wide spectrum of time harmonics. Within them the lower order time harmonics in general, having frequency closer to the wanted output frequency and the sub harmonics in particular, are found potentially objectionable in practice and are at the same time found difficult to be filtered-off. These supply time harmonics give rise to an increase in the copper and iron losses, to pulsating shaft-torques and to a small reduction of the steady-state torque produced by the motor. The magnitude and the distribution of the additional losses and the related motor derating, in steady state, depend on the harmonic contents of the voltage and, in some way, on the motor design. However, in any case, the motor has to be derated for the harmonic effects due to the non-sinusoidal nature of the voltage supply. Besides directly reducing the rating of the machine, these time harmonics produce other undesirable effects on the performance. Since electric motors are designed on the basis of balanced three-phase sinusoidal input voltage. So non-sinusoidal wave-shapes have detrimental effect on induction motor performance.

Static frequency converters can be subdivided into two basic sub-groups: dc link converters and cycloconverters. In dc link converters, the power taken from the ac system is first of all converted into dc power by a system side converter and then back into ac power of the required voltage, frequency and phase number by a machine side.

In the case of cycloconverters, the network frequency is directly converted to a lower frequency without intermediate rectification, the output voltage being formed from the sections of the input voltage. The cycloconverters are not generally favoured for various speed induction motor drives due to certain drawbacks. These static frequency converters produce complex waveforms. The most satisfactory way of considering such complex waveforms is to split them into a sinusoidal fundamental wave of the desired frequency and then consider the effect of the remaining harmonics separately. The unwanted harmonics will normally be at frequencies higher than the fundamental and many of the drive systems are specially arranged to avoid frequencies below or near to the fundamental values. The frequencies of the harmonics may be directly related to the fundamental frequency or they may be unrelated to the fundamental as the case in cycloconverters and some of the voltage source PWM systems.

The type of drive system also decides whether the harmonics occur predominantly in the voltage or the current. Voltage source inverters usually produce complex voltage waveforms containing large amounts of harmonics and the resultant current waveforms tend to contain a lesser value of harmonics. Conversely, current source systems produce currents with high harmonics contents and in many cases the voltage waveforms tend to be more sinusoidal. It is not usual to find similarly large harmonic contents in both the current and voltage of any system.

In general, harmonics do not produce any useful torque in any of the systems and produce additional losses and harmonic pulsations in the output torque from the motor shaft. Considering the equivalent circuits of motors under harmonic conditions can make further understanding of these effects. In an induction motor the magnitude of the equivalent circuit components will alter when harmonic frequencies are considered. The impedance values of the circuit leakage inductances will increase. They are effectively iron-cored inductances; their values will increase roughly in proportion to frequency. The stator resistance value will remain approximately constant; frequencies up to 1000 Hz are not likely to cause significant change due to skin effects. The rotor resistance however, will change and it could rise to twice the standstill value due to the influence of the deep bar designs used to improve starting performance. The use of an equivalent resistance for the core loss is not entirely accurate for harmonic considerations but this item is not significant to the overall flow of harmonic currents. The effective value of slip increases considerably when the harmonics are considered.

The behaviour of an induction motor on the non-sinusoidal voltage waveforms is different from the behaviour on normal three phase sinusoidal supplies. A knowledge of the performance of the inverter fed induction motor on these non-sinusoidal voltage is necessary for the proper design of the motor as well as to know the derating factor of the motor. The exact steady-state behaviour of the motor fed by non-sinusoidal supply is provided in this work. The method of analysis is based on the splitting up of the non-sinusoidal voltage into fundamental and harmonics and solving the equivalent circuit for the performance. The accuracy of the result depends upon the number of harmonics considered. Larger the number of harmonics the more accurate the performance. The losses of the non-sinusoidal supply fed induction motor are calculated and compared with that of sinusoidal supply. The comparison of efficiencies when supplied from these sources is given.

1.2 Literature Survey

In power system, induction motors are the largest component of the load and they are widely used in industrial, commercial and residential applications. Some industrial applications require speed control of induction motors, this can be done through electronic converters. A converter-fed electrical machine operates with a high harmonic content of voltage and current. The motor behaviour, when applied a non-sinusoidal ac source, is quite different when applying a sinusoidal one. So the operation characteristics of motor will be affected. Therefore, to study the impacts of induction motors under harmonic voltage, it becomes essential to review the literature published in the area of design and analysis of induction motor under non-sinusoidal supply. The study of this issue has been found since the 1960's. As upto date literature is presented in this chapter.

G.C. Jain [14] has pointed towards the various voltage waveshapes' effect on the three-phase induction motor. He has discussed the contribution of harmonics to copper losses and torque. He has concluded that the torque of induction motor is effected by negligible amount because of voltage waveshape.

Klingshirn and Jordan [15] have proposed a three-phase induction motor performance under non-sinusoidal voltage source. They have given the method of calculation of harmonic currents by using equivalent circuit approach. They have also given a simple method to account for the effect of magnetic saturation. The losses are separated into various components and are concluded that the harmonic losses are nearly independent of motor load. They have also shown that the losses of induction motor with non-sinusoidal waveform impressed across its terminals can be markedly different from its sinusoidal losses depending upon the harmonic content of the impressed waveform.

Linders J.R. [16] has investigated the effects of poor quality power on AC motors and proposed the hidden costs and containment due to the electric wave distortion.

Chalmers and Sarkar [4] have presented the various components of additional losses produced in induction motors having non-sinusoidal supply. In particular, the importance of losses due to skew-leakage fluxes and end-leakage fluxes has been demonstrated. They have pointed out that the skew-leakage losses are increased when the frequencies of either the fundamental or the largest harmonics are increased. In such cases it may become desirable to use the unskewed construction.

Sen P.K. and Landa H.A. [23] have proposed derating operation to the material of an induction motor under waveform distortion. It has been discussed that the non-sinusoidal voltage has detrimental effect on the induction motor performance and derating of the machine is needed. They have suggested that drip-proof machines to be less affected by harmonic distortion than the totally enclosed machines and that less found efficient machines would require a higher derating.

Boglietti A., Ferraris P., Lazzari M. and Profumo F. [3] have presented a simple methodology in order to analyse the energetic behaviour of induction motors fed by PWM inverters, in comparison to classical sinusoidal supply. They have also pointed out the method in order to separate the loss items in six-step inverter fed motors and given the knowledge of the loss items used for energetic consideration in order to select a suitable power derating value.

Venkatesan K. and Lindsay J.F. [24] have calculated the losses in an induction motor fed from six-step voltage and current source inverters and also given a comparison of efficiencies when supplied by these sources. They have calculated the main and stray copper losses with the help of equivalent circuit that includes the effect of space harmonics and corrected for stray copper loss. Stray iron losses due to magnetomotive forces (MMF) and permeance harmonics; end leakage and skew leakage are also considered. With fundamental air gap flux held constant, a comparative study of the losses in the motor when operated from voltage and current source inverters has been made.

Debuck G.G.f., Gistelinck P. and Backer deD. [7] have developed a semi-empirical induction motor loss model which estimates time-harmonic dependent losses in a non-sinusoidal fed induction motor as a function of motor power rating. Their aim is the estimation of losses caused by harmonic-distorted current and voltage supplies. The model may consider harmonics between 100 and 20000 and is constructed for power ratings between 1 and 1000 kW. Losses in rotor, stator and iron (including most important stray losses) are separately estimated and suitably combined. The model is characterized by its simple and reliable nature. They have also given the relationship how rotor resistance depends on bar height, shape, and material, frequency and open or closed slot character.

Oguchi K. [20] has given a comparison between motor performance of multistepped voltage-fed and pulse-width modulation (PWM) voltage-fed designs. Motor performance, such as the torque ripple and the peak stator current ratio, is calculated for the inverter output voltage waveforms with six to forty-eight steps. They have analyzed the performance of a multiple, phase-shifted inverter-fed induction motor by the model approach.

Antonio G.D. and Roberto P. [2] have given a method to calculate the extra iron losses occurring in the cores of the inverter fed electromagnetic devices. The loss increase referred to the sinusoidal operation is evaluated, analyzing its dependence on the inverter and on the lamination characteristics. After a critical analyses of the models of the iron losses, the expressions of these losses have been developed, as a function of the PWM inverter voltage waveforms and of its corresponding amplitude spectrum, taking into accounts the voltage drops, the winding characteristics and the features of the used laminations.

Ching Yin and W.J. Lee [5] have suggested about a real load test to investigate the effects of each order of harmonics from 2 to 13 under various Voltage Distortion Factor on the performance of a three-phase induction motors. According to test results, they have given idea that the harmonics with an order below 5 have more impacts on a three-phase induction motor than the higher order harmonics. They have strongly suggested following points:

a) Provide a quantized comparison of the effects of each order harmonic from order 2 to 13 on the operation performance of three-phase induction motors, particularly including even order harmonics, zero-sequence harmonics and harmonics with an order below 5.
b) Investigate the adequacy and adaptability of relative regulations relative to the voltage harmonic limits.

c) Use the temperature rise curve obtained by a real full load test to define/revise the new derating factor of an induction motor.

Murphy J.M.D. [18] has estimated the average torque produced in a 2 HP motor by the 5th supply time-harmonic having a magnitude of 20% of that of the fundamental. In this calculation he assumes a threefold increase in rotor resistance due to skin effect at the harmonic frequency and the torque was found to be only 0.125% of the fundamental torque.

The study of literature shows that the operation of induction motor fed from a static frequency converter is modified considerably in comparison with operation from sinusoidal voltage source. Presence of time harmonics in the applied voltage waveform gives rise to additional losses and pulsating torques in the motor. The additional losses are about 20 percent of the fundamental losses for the motor and there is necessity to see the design and derating aspects. It is strongly suggested that even order harmonics and harmonics having an order below 5 should be considered in related regulations of harmonics control and limits.

1.3 OUTLINE OF THE PRESENT WORK

In the present work, the steady-state performance of an induction motor fed by non-sinusoidal supplies has been evaluated by Fourier analysis. Two cases are considered operation of induction motor supplied with sinusoidal waveform and operation with the supply of three-phase inverters, namely six-step, twelve-step voltage source inverters, sinusoidal pulse width inverter (SPWM), modified sinusoidal pulse width inverter (MSPWM) and harmonic injection sinusoidal pulse width inverter (HIPWM). The rotor resistance variation due to skin effect has been calculated in terms of the harmonic frequency and conduction height.

1.4 ORGANIZATION OF THE WORK

In this very first chapter a detailed introduction is given. The literature published in the concerned field is mentioned.

Chapter-2 gives the discussion on inverters, which are used for the control of the induction motor and also sources for non-sinusoidal supplies. The line to line voltage spectrum for different supplies is shown.

Chapter-3 deals with the performance of the induction motor for both sinusoidal supply and non-sinusoidal supplies. The flow charts of various step procedures and functions, which are used to calculate the performance of induction motor under sinusoidal and for non-sinusoidal supplies, are given. The harmonic equivalent circuit has been driven. The total rms current is calculated. The torque behaviour of the induction motor is described. The impacts of the non-sinusoidal supply have been described. The impacts of the non-sinusoidal supply on consumers and utility companies are discussed.

In the chapter-4, results are listed in form of performance characteristics of the motor. The results are discussed with different power supplies. A comparison between motor performance of sinusoidal and inverter fed voltage is made. It is shown that the supply time harmonics affect the average torque very less in amount. The harmonic content and their magnitude of various forms are tabulated.

Finally, conclusions have been drawn from the results obtained. The scope for further work is also indicated in this chapter.

NON-SINUSOIDAL SUPPLY SOURCES FOR INDUCTION MOTOR

Most variable frequency drive systems use switching circuits which do not inherently produce sine waves, they normally produce square or pulsating voltages or currents. The output of the static controllers contains wide spectrum of time harmonics. These static controllers can be classified as one of the two types of power sources– voltage or current. In voltage source, the magnitude of the voltage applied to motor is directly controlled and the current is a function of motor characteristics and the applied voltage. In current sources, the level of current flowing in motor is controlled and voltage is a function of the motor characteristics and the current. There are two types of link inverters namely, the voltage source inverter (VSI) and current source inverter (CSI). Voltage source inverter can be further classified into two types of sources square-wave inverters with variable link voltage, and pulse width modulated inverter (PWM). The output of a current source inverter is a square wave current which contains harmonics low orders, therefore current source inverter drives suffer specially at low load speed and speed ripple problems. The variable voltage inverter output is variable amplitude square wave voltage while the PWM output consists of pulses with constant amplitude. Basically there are three major sources for speed control of induction motor :

2.1 Voltage source inverter (VSI) control

2.2 Current source inverter (CSI) control

2.3 Pulse width modulated (PWM) control.

2.1 VOLTAGE SOURCE INVERTER (VSI) CONTROL

The voltage source inverter control is more common and this type of inverter creates a relatively well defined switched voltage waveform at the terminals of the motor. The voltage inverter consists of a phase–controlled bridge, an auxiliary dc commutating circuit and inverter bridge. Variable voltage is generated by controlling the firing angle of the SCR's in the phase controlled bridge. The dc power rectified by means of phase controlled rectifier from the ac at the normal frequency is inverted to desired frequency by a force commutated inverter. A stiff dc bus voltage is maintained by the use of a large capacitor in the dc link. The inverter thyristors can conduct either for 180^o or 120^o. Because of the advantages in the performance, normally 180^o conduction is applied. The voltage source inverter controller is shown in Fig. 2.1(a).

The harmonic voltages associated with the voltage source inverter output are determined by the low output impedance of the voltage source inverter. The harmonic currents are limited by the induction motor leakage reactance. Typical voltage and current waveforms are shown in Fig.2.2(a). The controller acts as a voltage source and is operated as an open-loop frequency control and closed loop voltage control.

2.1.1 Six-Step Voltage Source Inverter

The six–step voltage source inverter is shown in Fig.2.3. In six-step voltage source inverter, each semiconductor is turned on and off for 180^o intervals, and each output terminal is connected alternatively for a half-period to the positive and negative rails of the dc supply. A three–phase output is obtained by preserving a mutual phase displacement of 120^o between the switching sequences in the three legs of the inverter.

The line voltages are shown in Fig. (2.4). The line output voltages can be described by the Fourier series as follows

$$v_{AB} = \frac{2\sqrt{3}}{\pi} V_d \left[\sin \omega t - \frac{1}{5}\sin 5\omega t - \frac{1}{7}\sin 7\omega t + \frac{1}{11}\sin 11\omega t + \frac{1}{13}\sin 13\,\omega t + \ldots\ldots\ldots\ldots \right] \quad (2.1)$$

The rms value of the line voltage is $\sqrt{2/3}$ V_d or 0.186 V_d, and the fundamental component has an rms value of $\sqrt{6}$ V_d/π or 0.78 V_d. In the six-step output voltage all odd harmonics except multiple of three are present.

2.1.2 Twelve–Step Voltage Source Inverter

A twelve–step waveform is shown in Fig.2.5 and has been used in ac motor applications where the harmonic content of the six-step wave was deemed excessive. The fundamental amplitude V_1 is given by

$$V_1 = \frac{\pi V_m}{3}$$

where V_m is the maximum step amplitude. The Fourier series expansion of the twelve-step waveform is

$$V = \frac{\pi}{3} V_m \left[\sin\omega t + \frac{1}{11}\sin 11\omega t + \frac{1}{13}\sin 13\omega t + \text{-----------------} \right] \quad (2.2)$$

All triplen harmonics are eliminated and the total harmonic content is 15.22 percent of the fundamental, as compared with 31.08 percent for a six-step wave.

2.2 CURRENT SOURCE INVERTER (CSI) CONTROL

The current source inverter controller consists of a phase–controller bridge and an inverter bridge. Commutation is achieved by means of commutation capacitors and reactors. Variable voltage is generated by controlling the SCR firing angles as in the volt-

-age source inverter. A large dc link reactor then serves as a filter to provide constant current to the inverter section where variable frequency current is generated. A block diagram of this approach is shown in Fig. 2.1(b).

The harmonic currents are determined by the inverter output current and have similar content to the harmonics associated with the voltage in the voltage source inverter control. Typical voltage and current waveforms are shown in Fig.2.2(b). The harmonic currents are limited by the motor leakage reactance. That means the lower the value of the leakage reactance the lower the harmonic voltages.

Better thermal performance on a current source inverter can be accomplished by using a motor with lower leakage reactance. This leads to the conclusion that a current source inverter does not lend itself well to operate motor horsepower ratings less than the controller rating, but it is acceptable to operate standard horsepower ratings larger than the controller rating. However, the actual motor performance from the larger motors will be limited by the current capability of the controller.

The basic current source inverter circuit produces six step, or quasi-square wave currents and the low order harmonic components will cause torque pulsation and irregular shaft rotation at low speeds and may excite mechanical resonance in the motor load systems. The predominant pulsating torque in the sixth harmonic component due to the fifth and seventh harmonic currents.

2.3 PULSE WIDTH MODULATED (PWM) INVERTER CONTROL

In many industrial application it is often required to control the output voltage of inverters (1) to cope with the variations of dc input voltage, (2) for voltage regulation of inverters and (3) for the constant volts / frequency control requirement. There are various techniques to vary the inverter gain. The most efficient method of controlling the gain (and output voltage) is to incorporate pulse-width modulation (PWM) control within the inverters.

The pulse width modulated (PWM) inverter is the most popular adjustable frequency controller technology today for voltage inverter . It is characterized by pulsed output waveforms of varying width to form a simulated sinusoidal of controlled variable frequency and rms voltage. The drive consists to a fixed diode rectifier and an inverter section operated in a manner of produce variable voltage and frequency. The output has reactors to filter the output current to the motor to reduce the magnitude of the harmonic currents. A block diagram of this approach is shown in Fig.2.1(c) and typical voltage and current waveforms are shown in Fig 2.2(c).

Due to the many different control strategies that can be used, it is not practical to generalize about PWM output characteristics. The output voltage and frequency can be rapidly altered within the inverter circuit, and the PWM inverter drive has a transient response which is much superior to that of the six-step inverter. When a sophisticated PWM strategy is adopted , there are no low-order harmonics in the motor current, and low-speed torque pulsations and cogging effects are eliminated. Pulse width modulated inverter system, in general provide superior performance to the six step alternations:

(1) The range of speed control is much wider and operation at and around zero speed is quite satisfactory.

(2) Low frequency torque pulsations do not occur in the output and hence there is less chance of exciting mechanical load resonances.

(3) The current waveforms in the motor are always very near to sinusoidal leading to less motor derating.

(4) The diode input rectifier means that the input power factor is always high whatever the speed and load.

(5) In multidrive systems it is possible to connect a number of inverters to the same power dc link to allow transfer of regenerated power from same drives to help feed other motoring drives.

However these advantages are primarily balanced by the increased complexity and by the increased difficulty in protecting these systems. The drawbacks of the PWM inverter control are listed as:

(i) attenuation of the wanted fundamental component of the PWMed waveform.

(ii) Drastically increased switching frequencies stress on associated switching devices and therefore derating of the switching devices.

(iii) Generation of high-frequency harmonic components.

The commonly used techniques [17] are namely single-pulse-width modulation multi-pulse-width modulation, sinusoidal-pulse-width modulation, phase-displacement control, modified sinusoidal pulse width modulation and harmonic injection pulse width modulation technique. The motor performance has been carried out by applying certain techniques:

2.3.1 Sinusoidal Pulse Width Modulation (SPWM) Technique

Since the ac motors are designed to operate on a sine wave supply and the inverter output voltage should be a nearly sinusoidal as possible. It seems that the three-phase square wave-reference should be replaced by a three-phase sine wave reference, to give a PWM waveform in which the pulse width in sinusoidally modulated throughout the half-cycle. Instead of maintaining the width of all pulses the width of each pulse is varied in proportion to the amplitude of a sine wave evaluated at the center of the same pulse. The distortion factor and lower-order harmonics are reduced significantly.

Each inverter phase or half-bridge has a comparator which is fed with the reference voltage for that phase and a symmetrical triangular wave which is common to all three phases as shown in Fig.2.5(e) Again the ratio of carrier to reference frequency or carrier ratio, p, must be a multiple of three to ensure identical phase voltage waveforms in a three phase system. The triangular carrier has a fixed amplitude, and the output voltage control is achieved by variation of the sine wave amplitude. This variation alters the pulse widths in the output voltage waveform but preserves the sinusoidal modulation pattern.

The area of each pulse corresponds approximately to the area under the sine wave between the adjacent midpoints of off-periods on the gating signals. If δ_m is the width of m^{th} pulse. Then, the rms output voltage

$$V_o = V_s \left(\sum_{m=1}^{p} \frac{\delta_m}{p} \right)^{1/2} \tag{2.3}$$

The carrier ratio, p, determines the order of the predominant harmonics in the sinusoidally modulated pole voltage waveform. The harmonics occur as side-bands of the carrier frequency and its multiples and in general, the harmonics order is given by

$$K = np \pm m$$

where m, n are the side-band and the carrier harmonic respectively.

For even values of n, there is an odd side-band spectrum because the harmonics are nonexistent when n and m are both even. Thus, the odd values of n possess even side band spectrum, because the harmonics are nonexistent when n and m are both odd. Large harmonics are also present is the pole voltage waveform at odd multiple of p, but since p is a multiple of three, these carrier harmonics, and certain other harmonics such as those of order 2p $\pm$ 3, are triplen harmonics, are not applied to the three-phase load. This type of modulation eliminates all harmonics less than or equal to 2p-1. The frequency spectrum is shown in Fig.2.6.

2.3.2 Modified Sinusoidal Pulse Width Modulation (MSPWM) Technique

In this technique the widths of pulses that are nearer to the peak of the sine wave do not change significantly with the variations of modulation index. This is due to the characteristics of a sine wave, and the SPWM technique can be modified so that the carrier wave is applied during the first and last 60^0 intervals per half-cycle (e.g., 0 to 60^0 and 120^0 to 180^0). This technique spectra are shown in Fig 2.7 for VSI operation.

The particulars follow:

(a) this technique defines the ac term Fig.2.7(c) on a line- to- line basis for VSI.

(b) As shown in Figs. 2.7(a), and 2.7(b) only the first and last 60^0 intervals (per half-cycle) of the ac term waveform are directly defined through intersections of respective (sine reference) and triangular (carrier) waveforms. The 60^0 to 120^0 intervals are obtained by "folding" the first and last 60^0 intervals around the 60^0 and 120^0 points respectively.

(c) As shown in Fig. 2.7(d) this technique provides a substantially higher ac term gain as compared with the original sine PWM technique. However it follows that hardware implementation for this technique generates a substantial (21-percent) ac term third harmonic component on a line-to-neutral basis.

2.3.3 Harmonic Injection PWM Technique

This improved technique (Fig 2.8) has been driven from the original sinusoidal pulse-width modulated (PWM) technique through the addition of 17 percent third harmonic component to the original sine reference waveform. The further variation can be obtained by injecting additional harmonics in the respective reference waveform. The resulting flat-topped waveform Fig.2.8(a) allows overmodulation (with respect to the original sine PWM technique) while maintaining excellent ac term and dc term spectra. The particular follows:

(a) the analytical expression for the reference waveform is

$$V = Vm_1 \sin(\omega t) + Vm_3 \sin(3\omega t)$$

(b) the ac term gain Fig.2.8(d) is equal to the respective gain obtained with the modified sine PWM Fig. 2.7(d) and substantially higher than the one obtained with the original PWM technique Fig.2.6(d)

The hardware implementation of this technique in quite simple. However, this technique also generates a substantial ac term third-harmonic component (17 percent) on line-to- neutral basis.

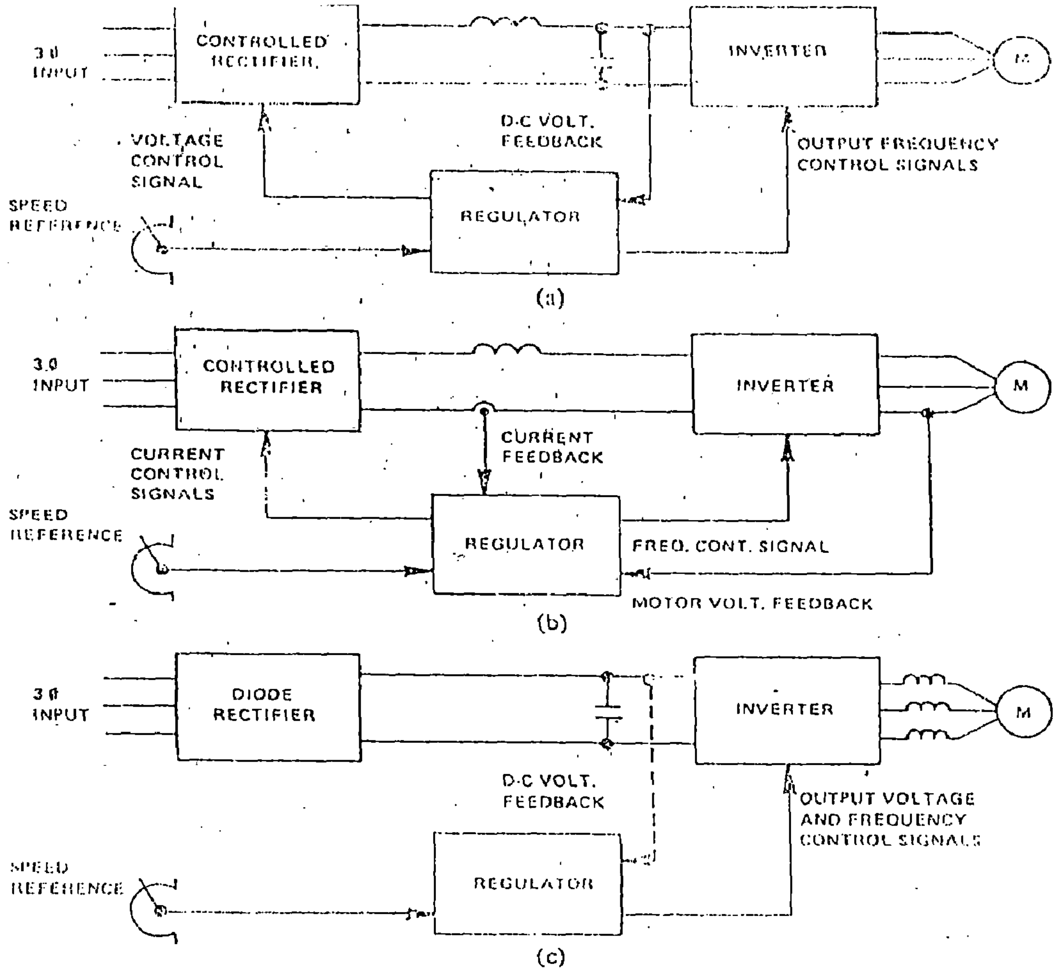


Fig. 2.1: **Block diagram of various controllers.(a) VSI. (b) CSI. (c) PWM.**

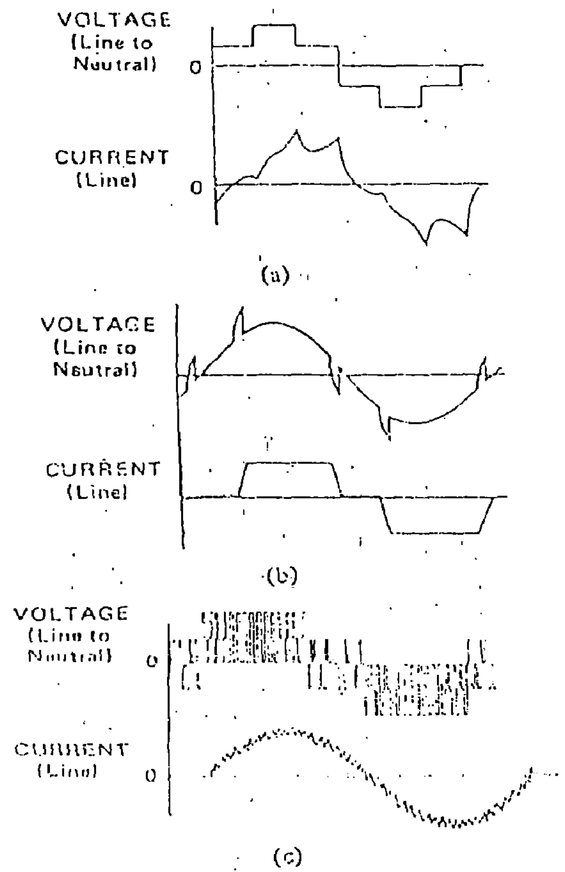


Fig. 2.2: **Voltage and current waveforms. (a) VSI. (b) CSI. (c) PWM.**

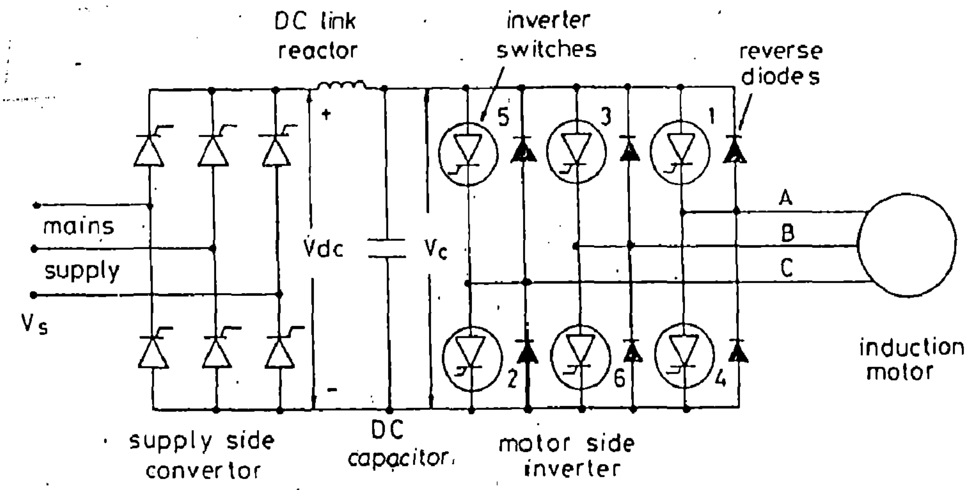


Fig. 2.3: Six-step voltage source inverter

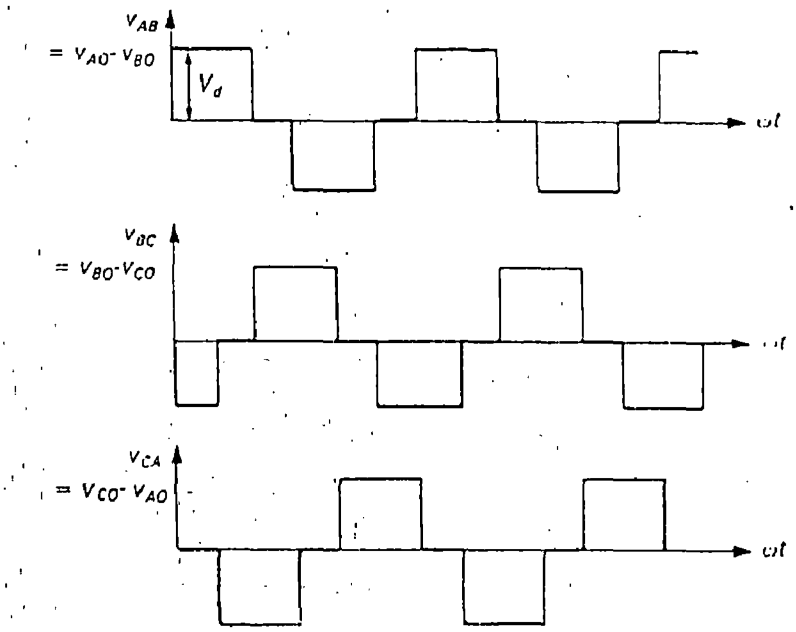


Fig. 2.4: Six-step line voltage

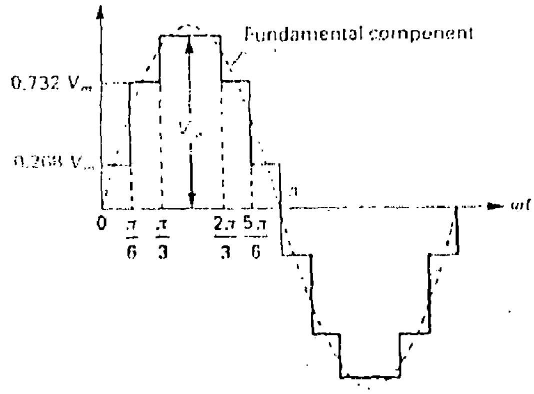


Fig. 2.5: Twelve-step voltage waveform

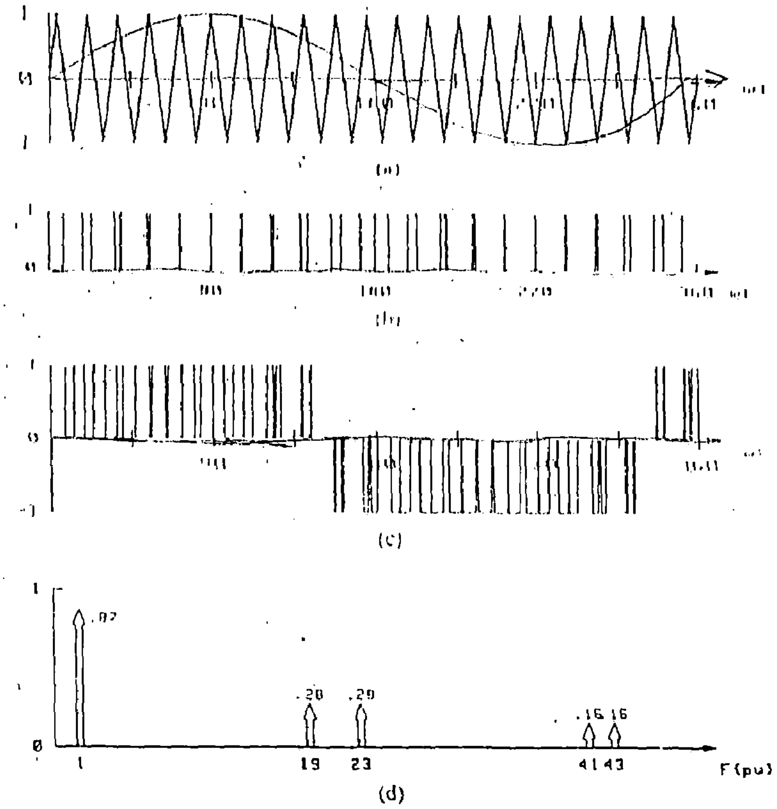


Fig. 2.6 Frequency spectrum of SPWM

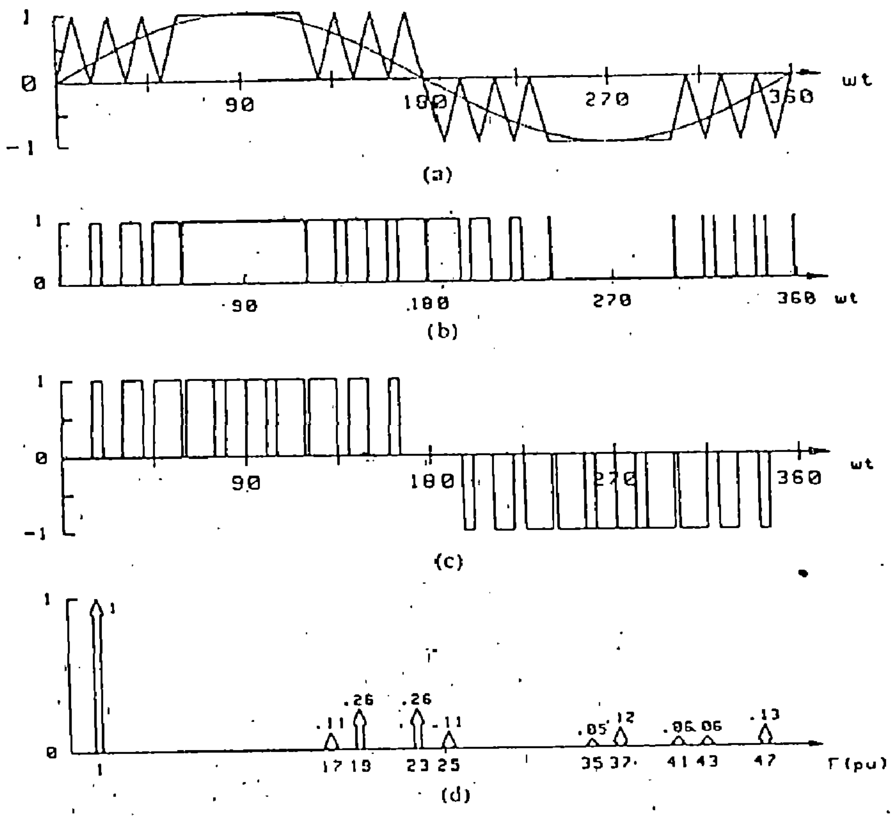


Fig. 2.7: Frequency spectrum of MSPWM

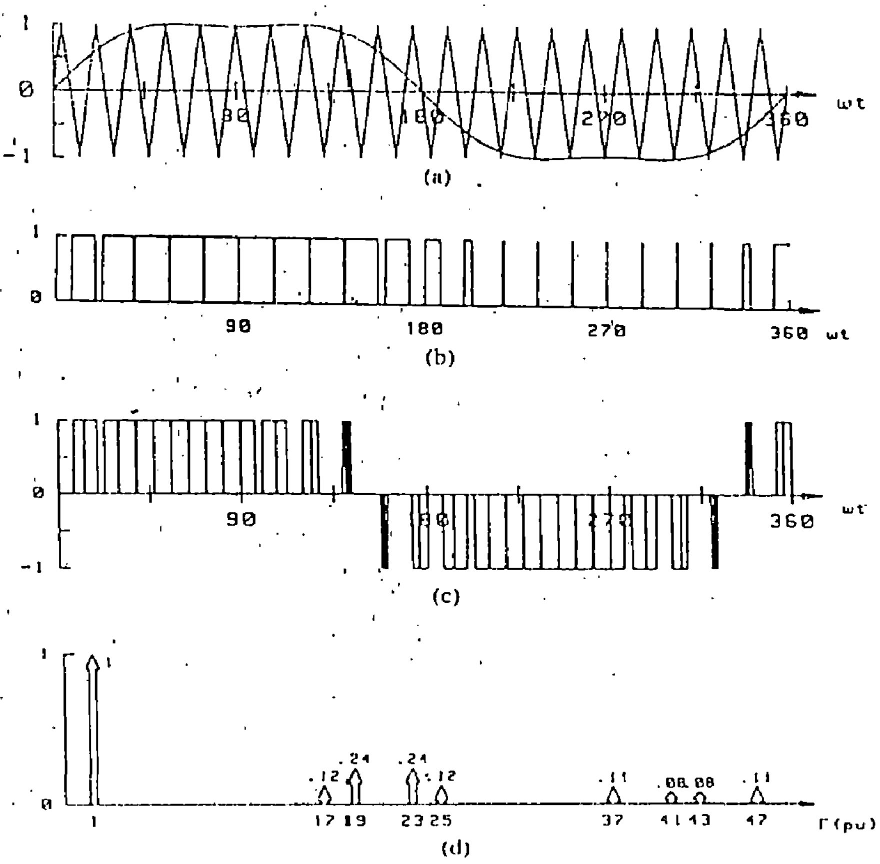


Fig. 2.8: Frequency spectrum of HIPWM

MOTOR PERFORMANCE ANALYSIS

Static converters which are being increasingly used to obtain flexible performance characteristics from robust but inherently constant speed cage induction motors, have the output voltage and current waveforms rich in harmonics. The effect on the motor performance due to the non-sinusoidal waveforms is to be concluded. So in this chapter, the performance analysis for sinusoidal and non-sinusoidal supplies is carried out.

3.1 ANALYSIS ON SINUSOIDAL SUPPLY

Conventional equivalent circuit is used for the steady state analysis of the induction motor. The equivalent circuit is shown in Fig. 3.1. Equivalent circuit can be solved by dividing into three parts in order to determine the performance of the induction motor.

Stator circuit impedance $Z_s = R_1 + jX_1$

Rotor circuit impedance $Z_r = R_2/s_1 + jX_2$

Magnetizing Reactance $= jX_m$

$$\text{hence } \overline{Z}_{m} = \frac{(R_2/s_1 + jX_2)(jX_m)}{[R_2/s_1 + j(X_2 + X_m)]} \tag{3.1}$$

$$m_3 = \frac{[R_2X_m(X_2 + X_m)]/s_1 - [X_2R_2X_m]/s_1}{(R_2/s_1)^2 + (X_2 + X_m)^2} \tag{3.2}$$

$$m_4 = \frac{[X_2 X_m (X_2 + X_m)] + [\frac{R_2}{s_1}.X_m.\frac{R_2}{s_1}]}{\left(\frac{R_2}{s_1}\right)^2 + (X_2 + X_m)^2} \quad (3.3)$$

So $\bar{Z}_{rm} = m_3 + jm_4$ (3.4)

Total per phase impedance of the motor

$$\bar{Z}_T = (R_1 + m_3) + j(X_1 + m_4) \quad (3.5)$$

The stator current per phase in calculated as;

$$\bar{I}_1 = \frac{\bar{V}}{(R_1 + m_3) + j(X_1 + m_4)} \quad (3.6)$$

Power factor,

$$PF = \frac{(R_1 + m_3)}{\left[(R_1 + m_3)^2 + (X_1 + m_4)^2\right]^{1/2}} \quad (3.7)$$

Voltage across magnetizing branch

$$\bar{E}_1 = \bar{V} - (R_1 + jX_1)(\bar{I}_1) \quad (3.8)$$

So current in magnetizing branch

$$\bar{I}_m = \frac{\bar{E}_1}{jX_m} \quad (3.9)$$

The rotor current

$$\bar{I}_2 = \bar{I}_1 - \bar{I}_m \quad (3.10)$$

$$\text{Air–gap power} = \frac{m_1 I_2^2 R_2}{s_1} \quad (3.11)$$

The electromagnetic torque is given by

$$T_e = \frac{P_G}{\omega'_s} \tag{3.12}$$

The input power to the motor is calculated as

$$P_{in} = m_1.V.I_1\ PF \tag{3.13}$$

Loss calculation

(i) Stator copper loss $P_1 = m_1\ I_1^2 R_1$ (3.14)

(ii) Rotor copper loss $P_2 = m_1\ I_2^2 R_2$ (3.15)

(iii) Iron loss

$$P_i = \frac{1}{2500}[21 + 28(B_M - 1.5)]f^2 \text{ W/kg} \tag{3.16}$$

where, B_m is the maximum flux density.

(iv) Friction and windage loss : It is calculated by the empirical formula given by Yermekova [26] as follows

$$P_{FW} = 0.016 \times HP \times 746 \times (1-s_1) \tag{3.17}$$

(v) Stray Load Loss: In the analysis, these are considered as the 2.07 percent of the output power.

Efficiency is calculated after adding all the losses

$$\eta = 1 - \frac{\sum \text{Losses}}{\text{Power Input}} \tag{3.18}$$

A flow chart is shown in Fig. 3.2 for the various performance calculations of the induction motor fed by the sinusoidal supply. A C-language program is made for the performance evaluation.

3.2 ANALYSIS ON NON-SINUSOIDAL SUPPLY

The input voltages to the induction motor operating on a voltage source inverter are periodic but non-sinusoidal. The waveform depends upon the type of converter and the period of conduction of thyristors. It is rectangular or stepped waveform for 180° conduction of thyristors. For a pulse width modulated inverter it is a pulse wave depending upon the method of modulation. For 120° conduction of thyristors the waveform depends upon the load also. If it is a motor load, the back emf of the motor affects this waveform. In this chapter a method of predicting currents and losses in the presence of time harmonics is presented. The following assumptions have been made to get simpler relations:

1. Saturation of iron parts of the machine is neglected.
2. The non-sinusoidal output voltage from three-phase bridge inverter is balance.

3.2.1 Method of Analysis

As the magnetic saturation is neglected, the motor may be regarded as linear device. The inverters produce output in the form of complex waveforms. The most satisfactory way of considering such complex waveforms is to split them into a sinusoidal fundamental wave of the desired frequency and then consider the effect of the remaining harmonics separately. The non-sinusoidal excitation is expressed as a Fourier series. Using Fourier series, a general expression can be written for the voltage applied to the motor as:

$$v(t) = \sqrt{2}\left[V_1 \sin\omega t + \sum_{k=2}^{n} V_k \sin(k\omega t + \theta_k)\right] \tag{3.19}$$

The voltage waveform expressed by equation (3.19) contains odd harmonics and does not contain even and triplen harmonics when the load is a three-phase induction motor. Principle of superposition can be applied to determine the overall performance of the induction motor for non-sinusoidal voltages; the motor behaviour for fundamental as well as for each harmonic is determined independently. These individual responses are added up to obtain overall performance.

The net current and torque of the machine are equal to the sum of the current and torque contributions of each harmonic of the voltage waveform. Fig.3.3 represents a series of independent generators, all connected in series, supplying the motor. Each generator would represent one of the voltage terms in equation (3.19).

3.2.2 Harmonic Equivalent Circuit

The conventional equivalent circuit is used in the steady-state analysis of the induction motor. The behaviour of the motor for a harmonic voltage is obtained by modifying the equivalent circuit for harmonic under consideration. Thus, a series of independent equivalent circuits, one for each harmonic is used to calculate the complete steady state behaviour of the motor. Harmonic equivalent circuit is shown in Fig.3.4.

In this case, the magnetizing resistance branch representing iron losses is neglected, since this branch contributes very little to the stator harmonic current. The effect of increase in iron losses is, however, considered separately as discussed in iron calculations, to be followed. The other differences between this circuit compared to the circuit at fundamental frequency are those need to take account of the harmonic frequencies. Thus for a time harmonic of order k^{th}, the modifications in the fundamental equivalent circuit are done in the following manner:

1. All reactances are evaluated at the harmonic frequency, which is k times the fundamental frequency.

2. Skin effect is taken into account in calculating the rotor resistance.

Hence, the rotor slip with respect to the fundamental rotating field, denoted by s_1, is given by

$$s_1 = \frac{n_1 - n}{n_1} \quad (3.20)$$

Where n_1 is the synchronous speed of the fundamental rotating field, and n is the actual rotor speed.

The harmonic voltages applied to the machine produce harmonic currents. These harmonic components in the phase currents produce time harmonic mmf. These mmf have a speed which is a multiple of the harmonic order and fundamental synchronous speed. Certain mmf's rotate in the forward direction and certain others rotate in the backward direction. Thus the speed of the k^{th} harmonic mmf is $kn1$. The rotor slip in a forward-rotating harmonic field is

$$s_k = \frac{kn_1 - n}{kn_1} \quad (3.21)$$

and for a backward-rotating field

$$s_k = \frac{kn_1 + n}{kn_1} \quad (3.22)$$

In general, therefore,

$$s_k = \frac{kn_1 \mp n}{kn_1} \quad (3.23)$$

Where the negative sign is valid for positive-sequence harmonics and the positive sign applies to negative-sequence harmonics. The harmonic slip, s_k, is expressed in terms of s_1 as follows,

$$s_k = \frac{(k \mp 1) \pm s_1}{k} \tag{3.24}$$

In equation (3.24), positive value refers to time harmonic fields rotating in the opposite direction to the fundamental field whereas negative value refers to time harmonic fields rotating in the same direction as the fundamental.

3.2.3 Performance Calculation

Fig. 3.4 shows the form of the equivalent-circuit of a 3-phase induction motor for k^{th} harmonic voltages where

$$R(k) = R_1 + \frac{kR_2}{(k \mp 1 \pm s_1)} \times \frac{(k \mp 1 \pm s_1)^2 X_m^2}{R_2^2 + (k \mp 1 \pm s_1)^2 (X_2 + X_m)^2} \tag{3.25}$$

and
$$X(k) = kX_1' + k \times \frac{X_m \left[R_2^2 + (k \mp 1 \pm s_1)^2 X_2^2\right] + (k \mp 1 \pm s_1)^2 X_m^2 X_2}{R_2^2 + (k \mp 1 \pm s_1)^2 (X_2 + X_m)^2} \tag{3.26}$$

for non triplen values of k. The top signs (of $\pm$ and $\mp$) refer to the forward rotating fields; i.e; for k=1,7,13….., whereas the bottom signs refer to the backward rotating fields,i.e; for k=5,11,17,

3.2.3(a) Harmonic Currents

The calculation of harmonic currents from harmonic equivalent circuits is done exactly in the same way as for fundamental currents.

If V_k is the maximum amplitude of k^{th} harmonic component of applied voltage per phase, then with proper choice of origin of time;

$$v(t) = \sum V_k \sin k\omega t \tag{3.27}$$

And then the rms value of k^{th} harmonic component of the stator phase current is given by

$$I_k = \frac{(V_k/\sqrt{2})}{\sqrt{R(k)^2 + X(k)^2}} \tag{3.28}$$

Usually, there are no zero-sequence harmonics and no even-numbered harmonics; hence the total rms harmonic current is given by

$$I_H = \left[I_5^2 + I_7^2 + I_{11}^2 + I_{13}^2 + --- + I_k^2\right]^{1/2} \tag{3.29}$$

$$= \sqrt{\sum_{k=2}^{n} I_k^2} \tag{3.30}$$

If I_1 is the fundamental rms current of the motor, the total rms stator current, including the fundamental, is

$$I_{rms} = \left[I_1^2 + I_5^2 + I_7^2 + I_{11}^2 + --- + I_k^2\right]^{1/2} \tag{3.31}$$

$$= \left[I_1^2 + I_H^2\right]^{1/2} \tag{3.32}$$

3.2.3(b) Motor Losses

Time harmonics of the voltage waveform applied to the motor result in a harmonic content in the stator and rotor currents, harmonic fluxes in the air gap and cause additional losses. The additional losses owing to time harmonic currents increase the heating of a given machine, and may lead to a reduction of its available continuous output. If these losses are high, locations and relationship to the motor and converter design features are to be investigated. The losses for the fundamental and each time harmonic are as follows:

(i) Stator Copper Loss:

The harmonic currents contribute to the total rms input current. Skin effect in the primary conductors may be neglected in small-wire wound machines, but it should be taken into account when the primary conductor depth is appreciable. Assuming the skin effect in the stator winding to be negligible, the stator copper loss on a non-sinusoidal supply is proportional to the square of the total rms current.

For the k^{th} time harmonic, the stator copper loss in a m_1-phase cage motor can be calculated as follows:

$$P_{cu1} = m_1 I_{rms}^2 R_1 \tag{3.33}$$

$$= m_1 \left[I_1^2 + I_H^2\right] R_1 \tag{3.34}$$

(ii) Rotor Copper Loss:

When considering the additional secondary copper losses, skin effect must be taken into account for all sizes of the motor since the secondary frequencies concerned are high (slip is equal to 1 for the harmonic with order > 1). Loss due to each harmonic must be considered separately and then added. In the case of a deep bar machine rotor, the total rotor copper loss per phase is given by

$$P_{cu2} = \sum_{k=2}^{n} I_{2k}^2 R_{2k} + I_2^2 R_2 \tag{3.35}$$

where, the first term represents the loss due to harmonics, and the second term will give the rotor copper loss due to the fundamental. At harmonic frequencies, the rotor resistance is much greater than the dc value. The actual increase depends on the geometrical shape of the conductor cross-section and of the rotor slot in which it is placed. Considering the geometrical shape, the rotor resistance depends on bar height, shape, and material, frequency, and open or closed slot character. As a very first approximation, slots are assumed to be open and rectangular. The rotor resistance for one singular rectangular bar as a function of frequency f_k and conductor height H has been calculated in following steps;

Bar resistance, as compared to its dc resistance,

$$\frac{R_{R,bar}(f_k)}{R_{R,bar,dc}} = 1 + C_1^{'} H_{(cm)} f_k^{0.5} \qquad (3.36)$$

With $C_1^{'}$ =0.1 .

With end-ring and bar dc resistance being approximately equal, total rotor cage resistance becomes

$$\frac{R_R(f_k)}{R_{R,dc}} = \frac{1 + (1 + C_1^{'} H_{(cm)} f_k^{0.5})}{2} \qquad (3.37)$$

or

$$\frac{R_R(f_k)}{R_{R,dc}} = 1 + C_1^{'} H_{(cm)} f_k^{0.5}, \qquad k > 1 \qquad (3.38)$$

with C_1=0.05.

The value of the constant C_1 depends upon the power rating of the motor which are listed below:

C_1=0.025----0.05, for P<=10 kW (when nominally saturated)

C_1=0.15, for P>30 kW

The additional rotor copper losses form a large portion of additional losses in the induction motors operating on non-sinusoidal supplies and are the main cause for reduced efficiency. The variation of the rotor resistance with the harmonic frequency is shown in the Fig. 3.5.

(iii) Iron Loss:

The core losses in the machine are also increased by the presence of harmonics in the supply voltage and current. These occur due to harmonic main fluxes. These core losses occur at high frequencies, but the fluxes are highly damped by induced secondary currents. A time harmonic mmf wave is established in the air-gap by each stator current harmonic. These time harmonic waves of mmf have the same number of poles as the fundamental field, but rotate forward or backward at a multiple of the fundamental speed. However, the resultant time harmonic air gap fluxes are small. These small harmonic main fluxes cause negligible increase in the core loss of the motor. The core loss due to space harmonic air-gap fluxes is also negligible, but the end-leakage and skew-leakage fluxes, which normally contribute to the stray load loss, may produce an appreciable core loss at harmonic frequencies. An exact determination of these losses is rather difficult. An approximate analysis shows that these losses are very small in comparison with the other additional losses. The iron losses are calculated by knowing the flux density, frequency of pulsation, mass of stator and rotor teeth and using the specific iron loss curves pertaining to the laminations at different frequencies.

Klingshirn et. al., Chalmers et. al. [15] & has stated that increase in iron loss due to non-sinusoidal excitation is a small fraction. Since the voltage of k^{th} harmonic is $1/k$ times the fundamental, the k^{th} harmonic flux density then is $1/k^2$ times the fundamental flux density.

$$B_m(k) = \frac{1}{k^2} B_m(1) \tag{3.39}$$

Where $B_m(k)$ is the k^{th} harmonic flux density and $B_m(1)$ is the maximum flux density of fundamental wave.

Specific iron loss (W/kg) for different flux densities and at frequencies higher than 50 Hz is calculated using the following expression:

$$P_i = \frac{1}{2.4025}\left[(0.1351f + 0.000136f^2)B_m^2\right] \text{ W/Kg} \tag{3.40}$$

For frequencies equal to or lesser than 50 Hz and flux densities greater than 0.8 Wb/m^2, for 0.5 mm following expression obtained by approximating the iron loss curve [22] given in Fig 3.6 at 50Hz for lamination thickness is used

$$P_s = \frac{1}{2500}\left[21 + 28(B - 1.5)\right]f^2 \text{ W/kg} \tag{3.41}$$

For flux densities less than 0.8 Wb/m^2, equation (3.40) is used.

(iv) Friction and windage loss:

Friction losses in bearings will vary in direct proportion to the speed of the motor. The windage losses are caused by the fans; which may be mounted on the rotor for cooling purposes and by the rotation itself. The power losses caused by windage will be proportional to the third power of the rotor speed and hence they will drop to very low levels under low speed operating conditions.

These losses are not influenced by the harmonics in voltage waveform. When the slip or frequency changes, however, then these losses are changed. These losses are empirically determined. An approximate expression, developed on the lines of Yermekova [26] is given below:

$$W_{FD} = 0.016 \times HP \times 746 \times (1 - s_1) \text{ W} \tag{3.42}$$

(v) Stray Losses:

Stray losses are due to space harmonics; crosscurrents in iron. The normal stray load losses are contributed by end leakage and skew leakage fluxes. At harmonic frequencies these leakage fluxes are very much affected and consequently there is an increase in these losses. The end leakage flux and the associated stray load losses are present in both the stator and rotor. The skew leakage flux is present only in squirrel cage motors. These stray load losses are normally 0.5 percent of rated output with sinusoidal excitation. In small machines, this may even be 1.5 to 3 percent. The increase in these losses can be taken to be proportional to the square of the current. These losses can be empirically determined using the current harmonics at a particular frequency. As a result of complexity the harmonic stray losses are taken only about 8 percent of the fundamental stray losses [24].

Thus the sum of total losses under non-sinusoidal supply is calculated as

$$P_{lossh} = P_{cu_1} + P_{cu_2} + TIL + PFW + PS \tag{3.43}$$

3.2.4 Torque Behaviour of Induction Motor

The presence of time harmonic mmf wave in the air gap results in additional harmonic torque on the rotor. These torques are of two types:

(a) Steady Harmonic Torques:

Constant or steady torques are developed by the reaction of harmonic air gap fluxes with harmonic rotor mmfs, or currents, of the same order. However, these steady harmonic torques, which are very small fraction of rated torque, have negligible effect on motor operation. The fundamental torque is given by

$$T_1 = \frac{pm_1}{2\pi f_1}(I_2)^2 \frac{R_2}{s_1} \tag{3.44}$$

The k^{th} time harmonic currents produce mmf rotating forwards or backwards at a speed kNs. The harmonic slip s_k is expressed in terms of the fundamental slip s_1 as

$$s_k = [(k \mp 1) \pm s_1]/k \tag{3.45}$$

So, the k^{th} harmonic torque contribution, T_k is given by;

$$T_k = \pm \frac{pm_1}{2\pi n_k}(I_{2k})^2 \frac{R_{2k}}{s_k} \quad (3.46)$$

The k^{th} harmonic currents of order (6m + 1), where m is an integer, has the same phase sequence as the fundamental current with positive phase rotation, and the k^{th} harmonic currents of order (6m-1) have a phase rotation opposite to that of the fundamental currents.

The steady state harmonic torques are useful or braking torques in an induction motor. The average torque of the induction motor is the algebraic sum of these harmonic torques and the fundamental torque.

(b) Pulsating Harmonic Torques

Pulsating torque components are produced by the reaction of harmonic rotor mmfs with the harmonic rotating fluxes of a different order. The harmonic air gap fluxes are small, and the dominant pulsating torques arise from the interaction between harmonic rotor currents, or mmfs, and fundamental rotating flux. By the excitation of the induction motor, stator magnetic field is developed which rotates at a speed of kN_s relative to stator. Each of these fields induces the rotor harmonic currents. These produce rotor mmfs which rotate at a speed kN_s with respect to stator. The torques developed by the reaction of stator and rotor fields having the same speed with respect to stator are constant. The torques developed by the fields of different speed pulsate with a frequency corresponding to the difference of the speeds of rotor and stator mmfs. The mmfs have the same number of poles. So each stator mmf reacts with other rotor mmf to produce a torque.

Both motoring and braking torques are possible. The fifth harmonic produces a braking torque whereas seventh harmonic, motoring torque.. These are very small in the inverter operation of the motor and can be neglected. Sixth harmonic pulsating torque occurs when fundamental of stator mmf reacts with either fifth or seventh harmonic rotor mmfs. Fifth and seventh stator harmonics reacting with fundamental of rotor mmf also produce sixth harmonic pulsating torque. Reaction of fundamental with eleventh and thirteenth harmonics results in a pulsating torque having 12 times the fundamental frequency.

The pulsating torques have zero average value, but their presence causes the angular velocity of the rotor to vary during a revolution. At very low irregular cogging motion sets a lower limit to the useful speed range of the motor. The point at which the speed pulsation becomes objectionable depends on the inertia of the rotating system. In certain applications, such as machine tool drives, the speed fluctuation is intolerable. Abnormal wearing of gear teeth can also occur, particularly if the torque pulsation coincides with a shaft mechanical resonance.

This resonant frequency is usually below 100 Hz, however, and a sixth harmonic torque pulsation is outside the range of shaft resonance over most of the speed range.

The pulsating torque amplitude is given by

$$T_{6k} = \phi_1 \left[I_{6K-1} - I_{6K+1} \right] \tag{3.47}$$

$$T_{6k} = \frac{\phi_1^2}{X_{pu}} \left[\frac{1}{(6K-1)^2} - \frac{1}{(6K-1)^2} \right] \tag{3.48}$$

where X_{pu} is the per unit leakage reactance of the induction motor, ϕ_1 is the fundamental flux and K is an integer.

The negative sign in the equation indicates the individual pulsating torque components due to I_{6k-1} and I_{6k+1} are in direct phase opposition, thereby reducing the resultant torque amplitude.

Equation (3.48) indicates that a significant reduction in pulsating torque is possible when fundamental air gap flux is reduced. This also diminishes fundamental torque, of course, but if a particular application does not require high torque at low speeds, then field weakening can be employed to reduce pulsating torque amplitude and so extend the useful low-speed range of operation.

Low order torque pulsations can be avoided by supplying the motor with an improved voltage or current waveform, such as a PWM waveform employing a sinusoidal modulation strategy. However, it is a characteristic of PWM techniques that large amplitude torque pulsations are produced at high switching frequencies.

3.2.5 Motor Efficiency

The harmonic content of the voltage wave has a direct bearing on the efficiency of the motor. The magnitude of the harmonic losses depends on the harmonic content of the motor voltage and current. Large harmonic voltages at low harmonic frequencies cause significantly increased machine losses and reduced efficiency. However, most inverters do not generate harmonics of lower order than the fifth, and high order harmonic currents usually have small amplitudes.

The main cause to reduce the efficiency is the increase in motor resistance due to skin effect. So the increase in copper losses is higher than the other losses. The increase in core loss is less predictable because it is influenced by the machine construction and magnetic materials used. If the high order harmonic content of the stator voltage waveform is relatively low, the total harmonic core loss due to skew–leakage and end–leakage effects should not exceed 25 percent of the fundamental core loss.

In induction motor, the harmonic currents and losses are practically independent of load, and the no load losses of an inverter fed motor may be appreciably greater than those which are excited by sinusoidal supply.

3.2.6 Drive Stability

Instability in the induction motor drive may occurs for certain critical frequency ranges and loading conditions when supplied by non-sinusoidal supply. Machine that are perfectly stable on an ac utility network may become unstable with an inverter supply, and machines that are stable when operated individually may become unstable when several of the motors are operated simultaneously as a group drive. There are two types instability.

(a) Inherent low–frequency instability in the motor

(b) Instability due to interaction between the motor and inverter.

The transient response of the induction motor becomes more oscillatory as the supply frequency is reduced, but the normal machine does not usually become unstable on an infinite system. However, small motors with a low inertia constant may be unstable Reducing the magnetizing reactance and increasing the stator and rotor resistances may improve induction motor stability. Instability between motor and converter arises when the converter which supplies the adjustable–speed motor has a finite source impedance. The impedance may be introduced by a transformer or by the filter, which smooths the dc link supply in a static inverter drive. System instability usually occurs at frequencies below 25 Hz when an inter change of energy taken place between motor inertia and filter capacitor. In an open loop adjustable–speed drive, the unstable region of operation is normally confirmed to a certain torque and frequency drive. A torque frequency diagram can be prepared, as in Fig. 3.7 in which the unstable zone is enclosed by the contour. The size of the instable zone is affected by the system inertia, the load damping and the electrical parameters of the motor and supply source, the harmonic content of the stator voltage waveform may slightly affect system stability, particularly if the inertia is small. Stability of the inverter–fed induction motor drive is generally enhanced..

(a) by increasing load torque and inertia

(b) by reducing stator voltage

(c) by increasing filter capacitance, reducing filter inductance and resistance.

3.2.7 Vibration and Overheating

The potential damage caused by harmonics voltage to a three-phase induction motor mainly come from rotor vibration and overheating. Rotor vibration originates from pulsation torque caused by positive and negative–sequence harmonics. The vibration of the rotor can increase friction losses of the bearings and reduce the life span of the bearings, thus largely increase the probability of mechanical failure. If the rotor axis does not have enough strength, the vibration is likely to twist the rotor axis and the rotor will rub the stator causing overheating until the wedges are damaged.

Overheating originates from harmonic currents. The temperature rise of the induction motor in any harmonic order case is greater than in the normal (fundamental wave) case. It means that an induction motor overheating may happen in any harmonic condition. This will lead to the damage of stator windings, wedges and bearings, shortening of the life span of the induction motor.

3.2.8 Flow Chart for Determination of the Performance of the Induction Motor

A simple flow chart to achieve the motor performance fed by non-sinusoidal supplies is shown in Fig. 3.8(a) & Fig. 3.8(b). Fig.3.8(a) & Fig. 3.8(b) are concerned with stepped inverter supply and PWM inverter supply respectively. The input data for a computer program based on this flow chart includes motor parameters (rotor parameters referred to stator), maximum core flux density, maximum stator teeth density, stator core weight, stator teeth weight, power rating of the motor and the ratio voltage. The detailed input has been tabulated in Appendix. The output data for this program consists of the values of various harmonic order, stator currents, rotor currents, power output, efficiency, torque and power factor all for individual harmonics as well as for the whole harmonics at no load to full load speed. A C-language program based on the flow charts is developed to investigate the motor performance. In the performance evaluation harmonics upto 31st order are considered. For each harmonic, motor is run from on load to full load. After adding all the power inputs, losses, torques of each harmonic at a particular speed, the efficiency and torque are calculated.

Thus due to the non-sinusoidal supply, the motor performance is deteriorated. The efficiency reduces as the motor resistance increases due to the skin effect. So the supply, which is given for the induction motor control should be more close to the sinusoidal to get better performance.

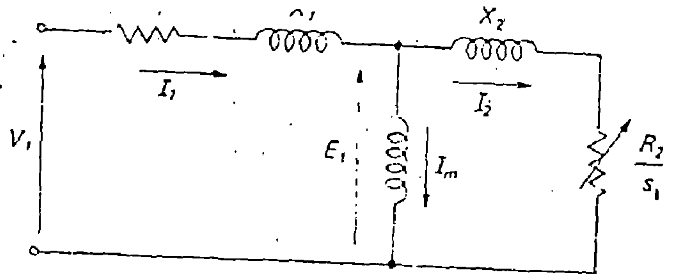


Fig. 3.1: Conventional equivalent circuit of the induction motor

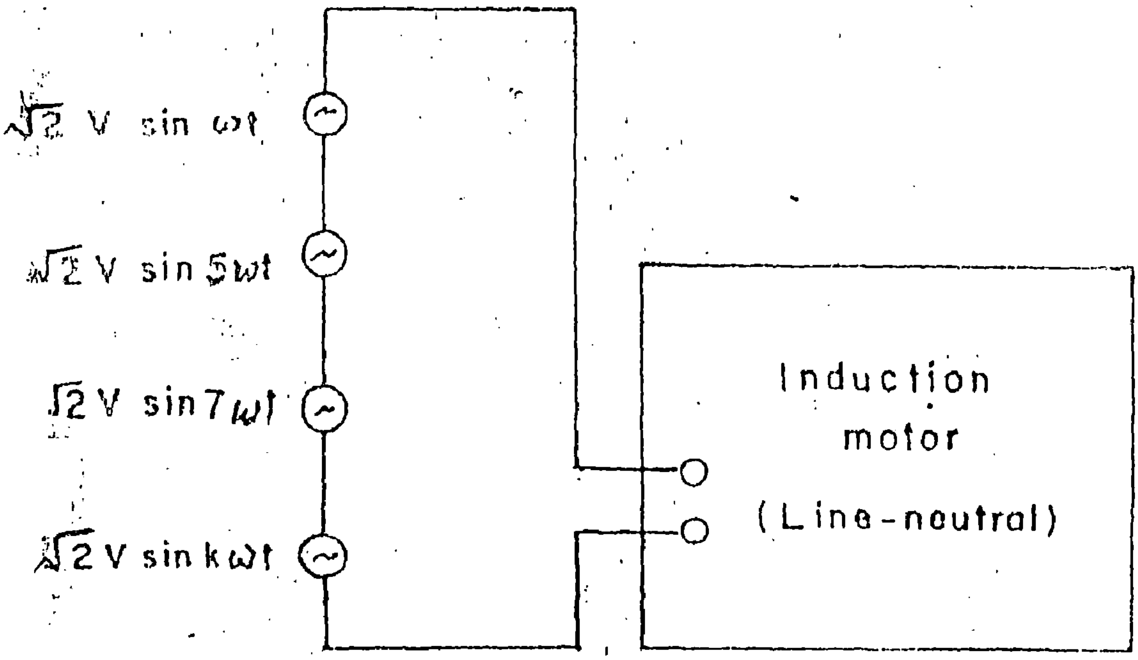


Fig. 3.3: Non-sinusoidal excitation of induction motor per phase

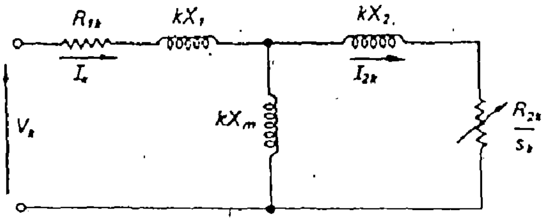


Fig. 3.4: Harmonic equivalent circuit of the induction motor

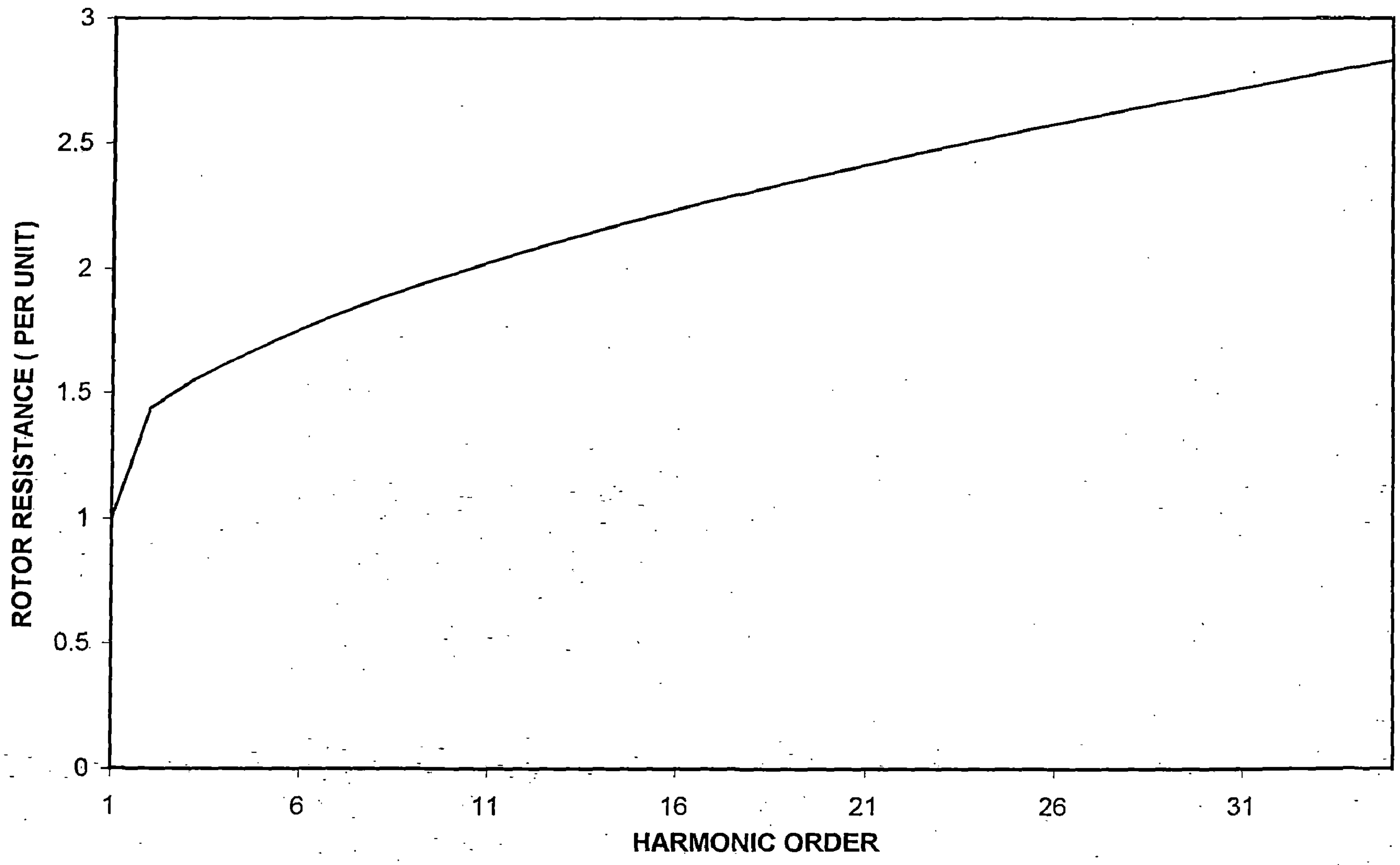


Fig.3.5 ROTOR RESISTANCE VARIATION WITH HARMONICS

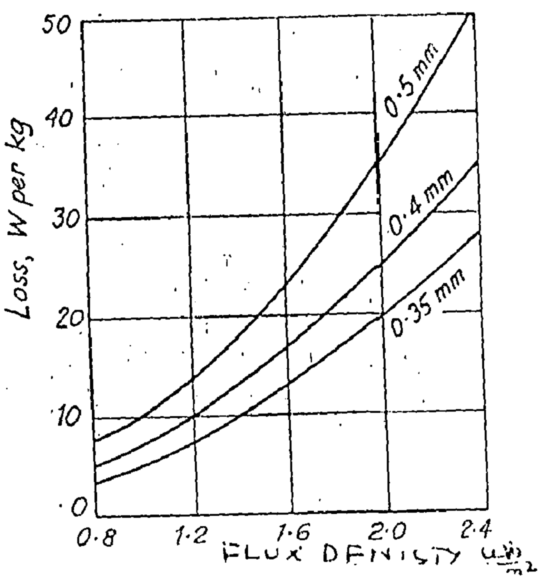


Fig. 3.6: Iron loss curve

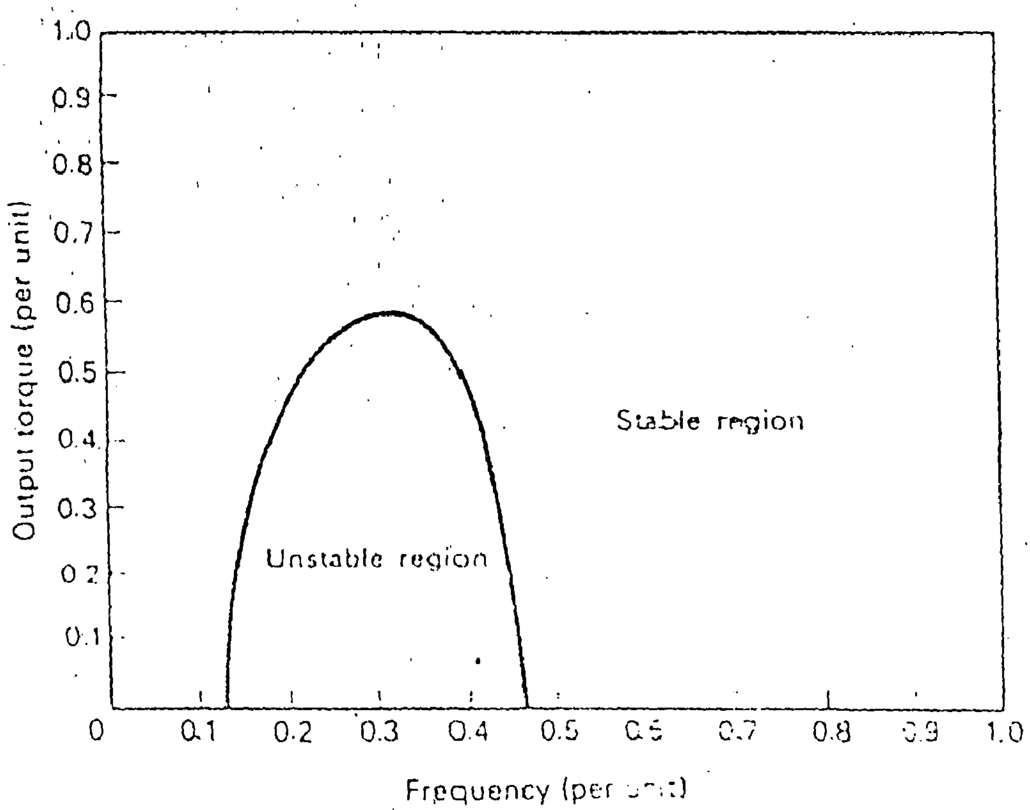
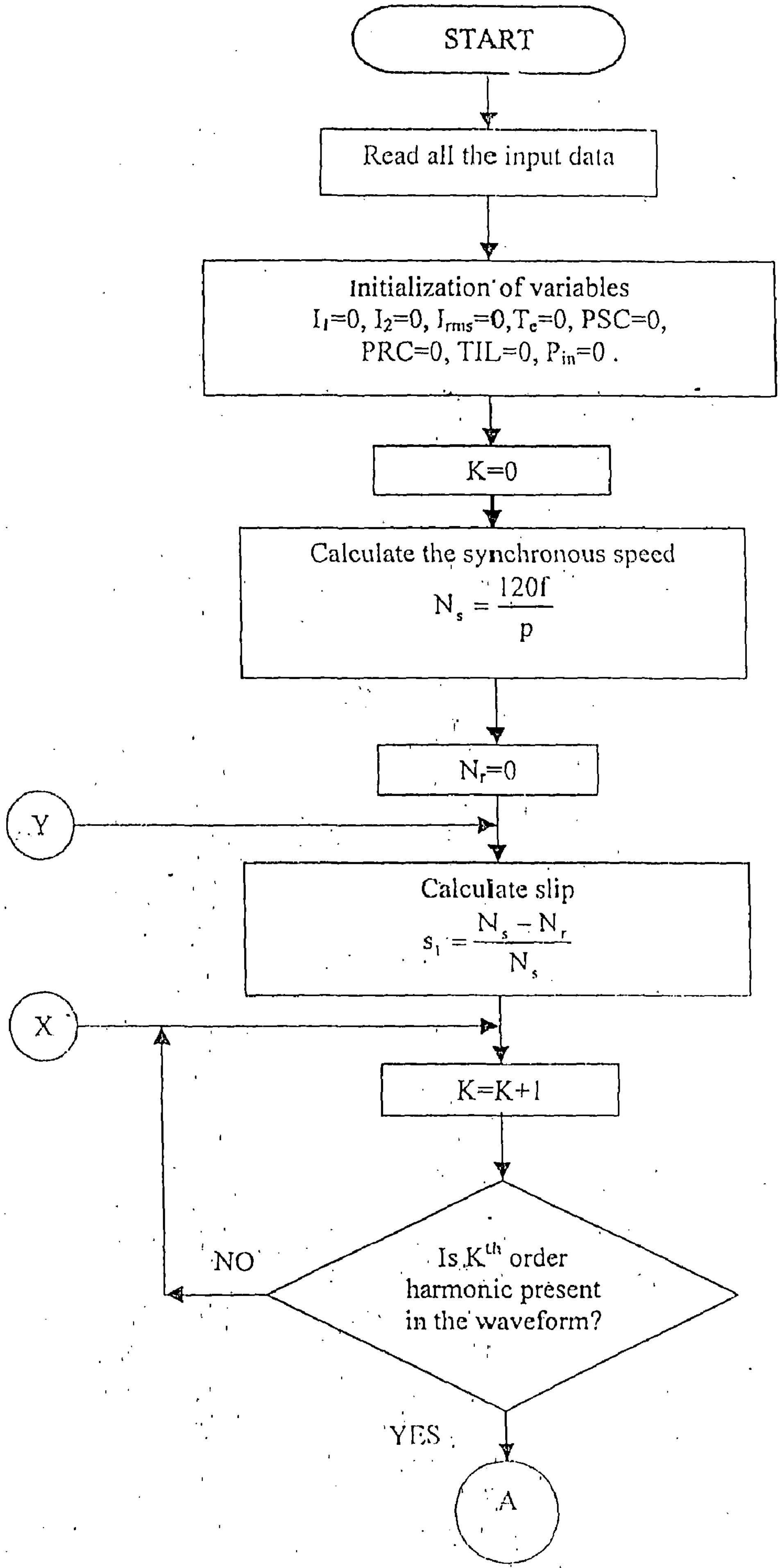
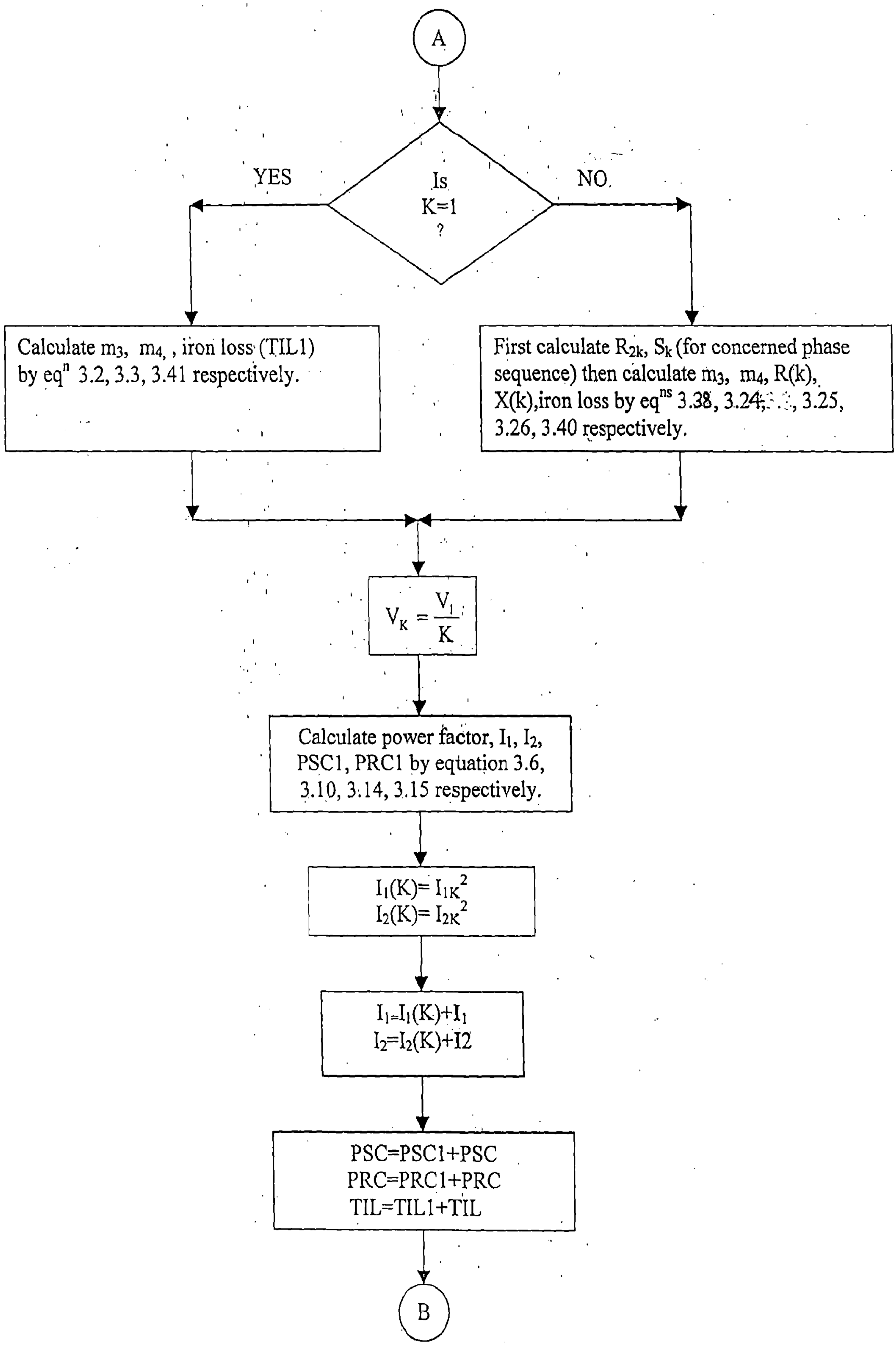
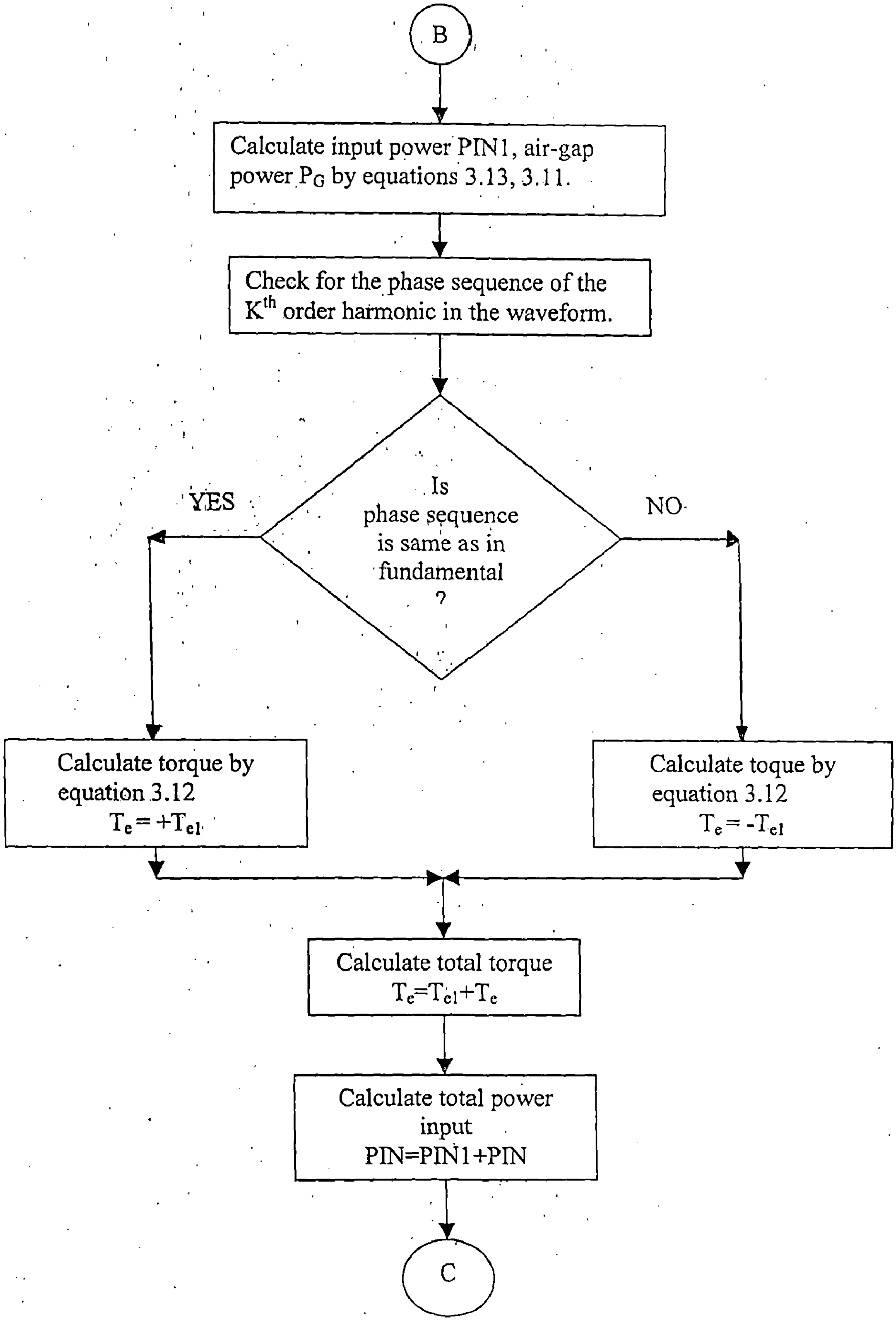


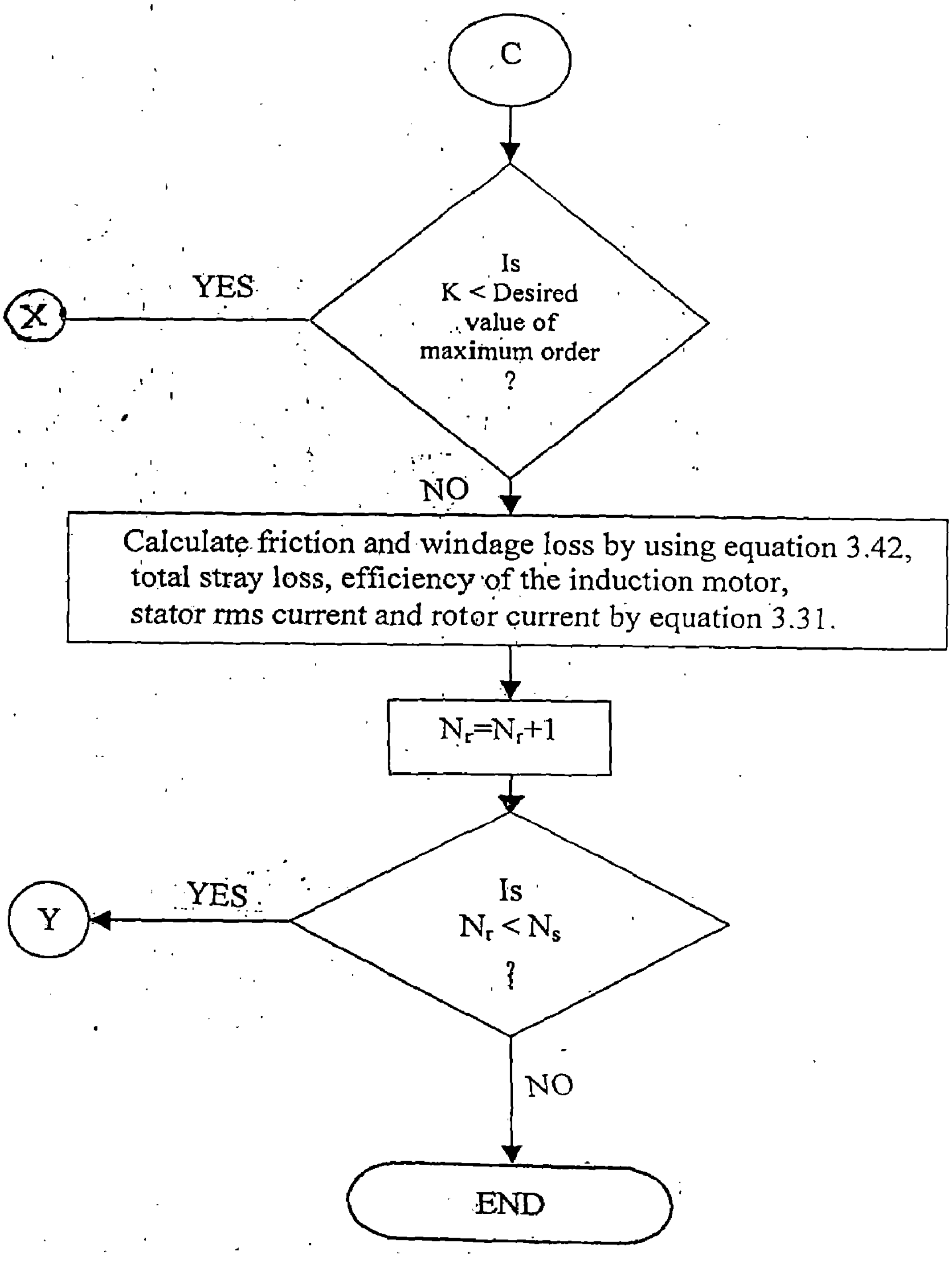
Fig. 3.7: Stability boundary for a converter-fed ac motor drive

Fig. 3.8(a): Flow chart for stepped VSI supplies




A
Is
K=1
?
YES
NO
Calculate m3, m4, , iron loss (TIL1) by eqn 3.2, 3.3, 3.41 respectively.
First calculate R2k, Sk (for concerned phase sequence) then calculate m3, m4, R(k), X(k),iron loss by eqns 3.38, 3.24, 3.25, 3.26, 3.40 respectively.
VK = V1/K
Calculate power factor, I1, I2, PSC1, PRC1 by equation 3.6, 3.10, 3.14, 3.15 respectively.
I1(K)= I1K2
I2(K)= I2K2
I1=I1(K)+I1
I2=I2(K)+I2
PSC=PSC1+PSC
PRC=PRC1+PRC
TIL=TIL1+TIL
B

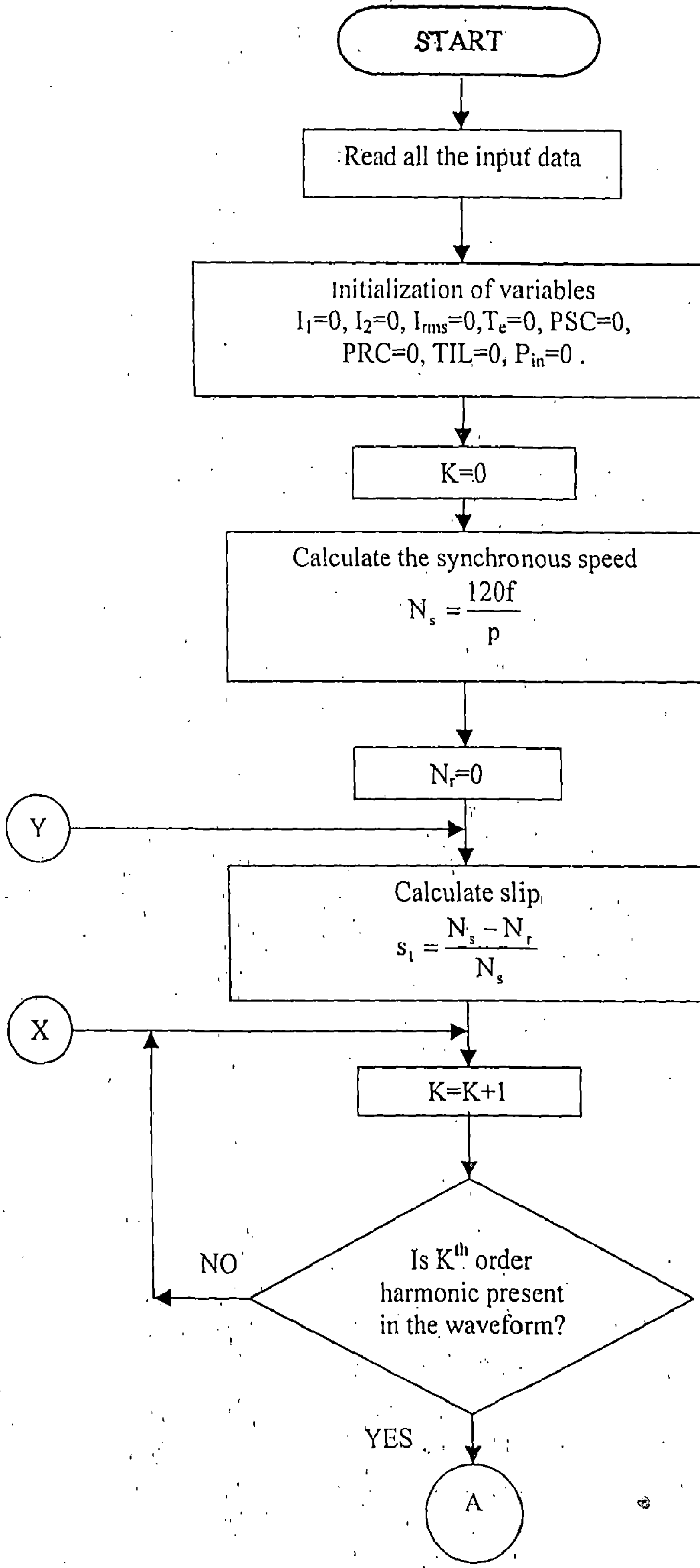

B
Calculate input power PIN1, air-gap power P_G by equations 3.13, 3.11.
Check for the phase sequence of the K^{th} order harmonic in the waveform.
Is phase sequence is same as in fundamental ?
YES
NO
Calculate torque by equation 3.12 $T_e = +T_{el}$
Calculate toque by equation 3.12 $T_e = -T_{el}$
Calculate total torque $T_e=T_{el}+T_e$
Calculate total power input PIN=PIN1+PIN
C

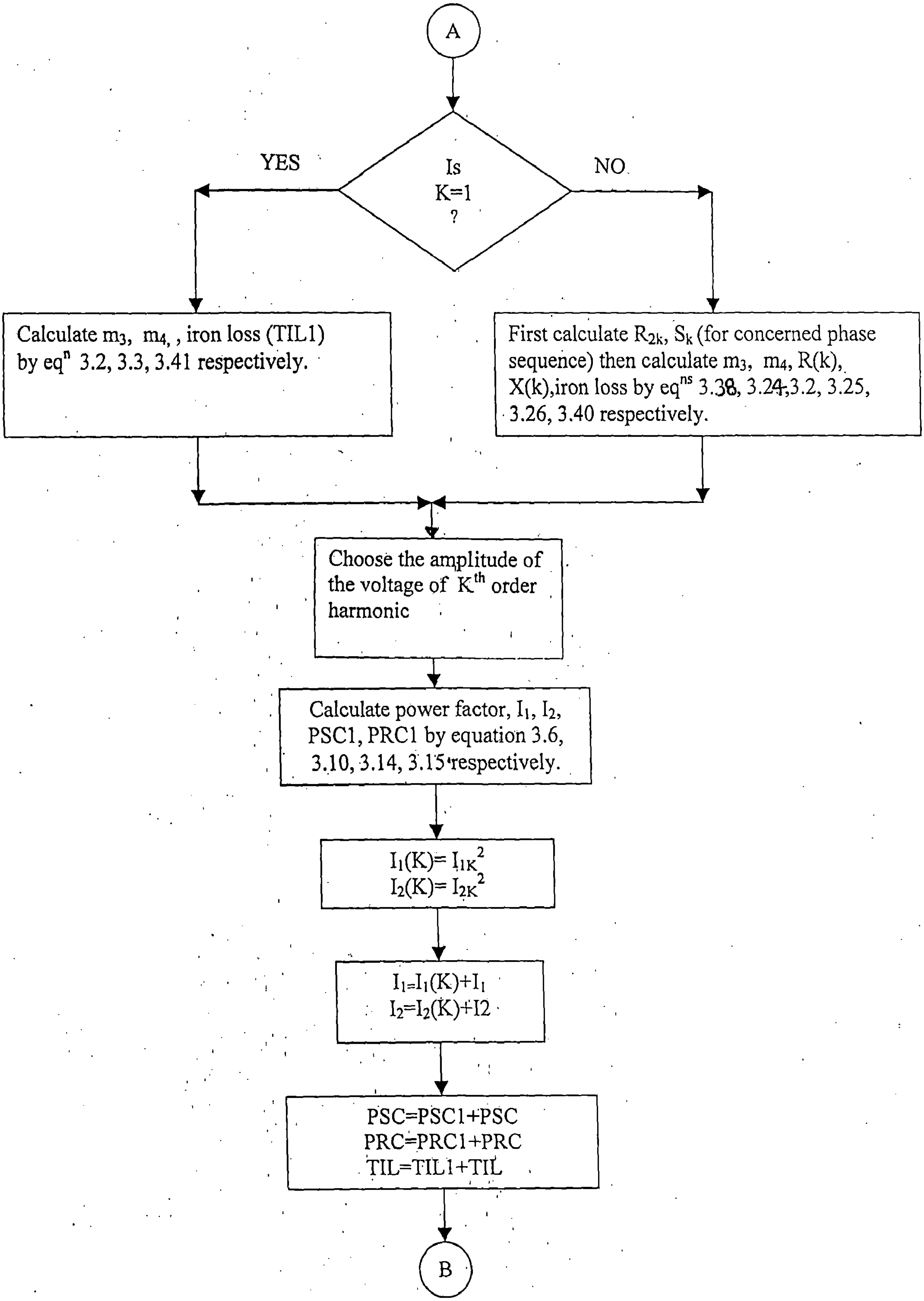


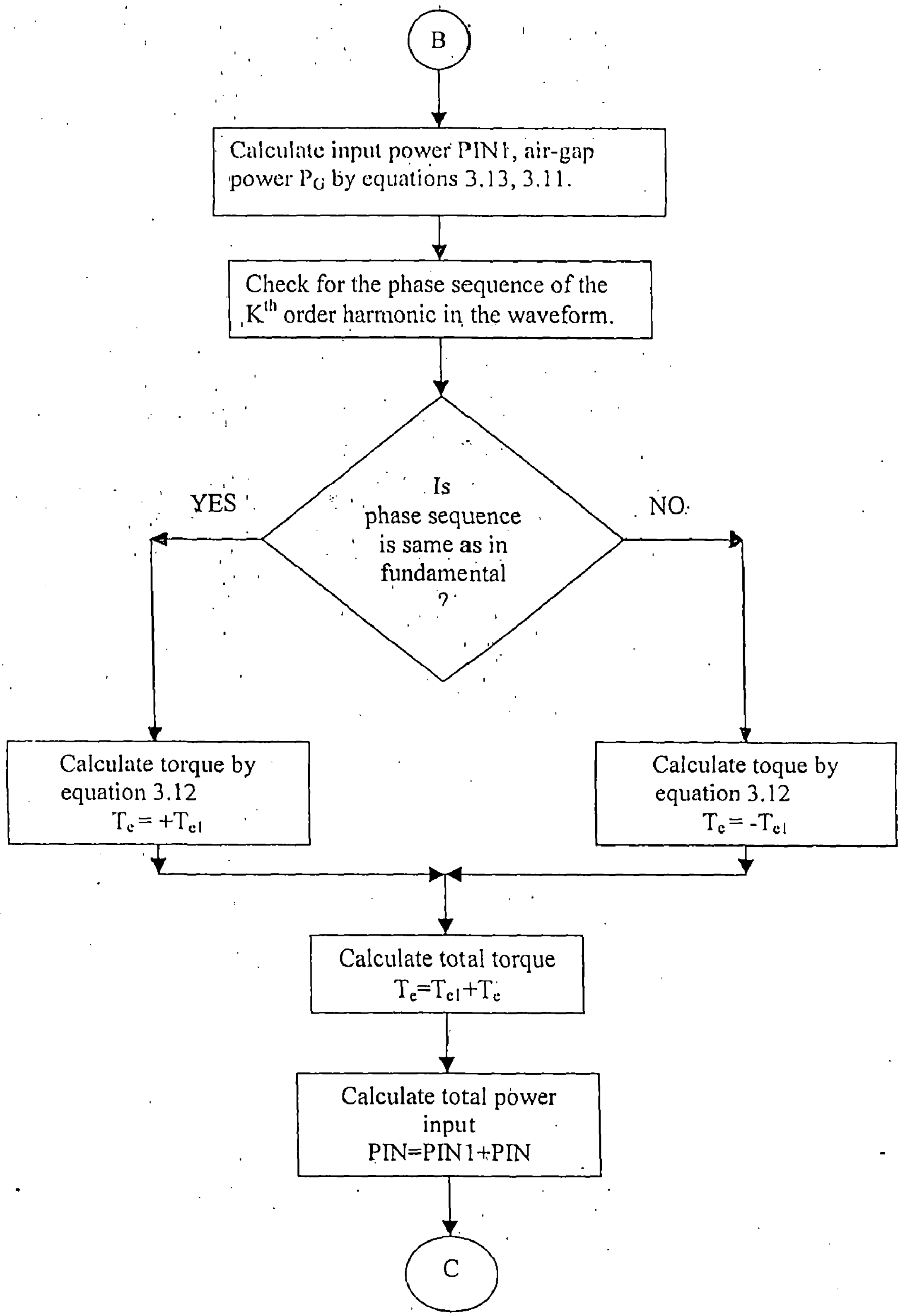
C
Is
K < Desired
value of
maximum order
?
YES
X
NO
Calculate friction and windage loss by using equation 3.42, total stray loss, efficiency of the induction motor, stator rms current and rotor current by equation 3.31.
Nr=Nr+1
Is
Nr < Ns
?
YES
Y
NO
END

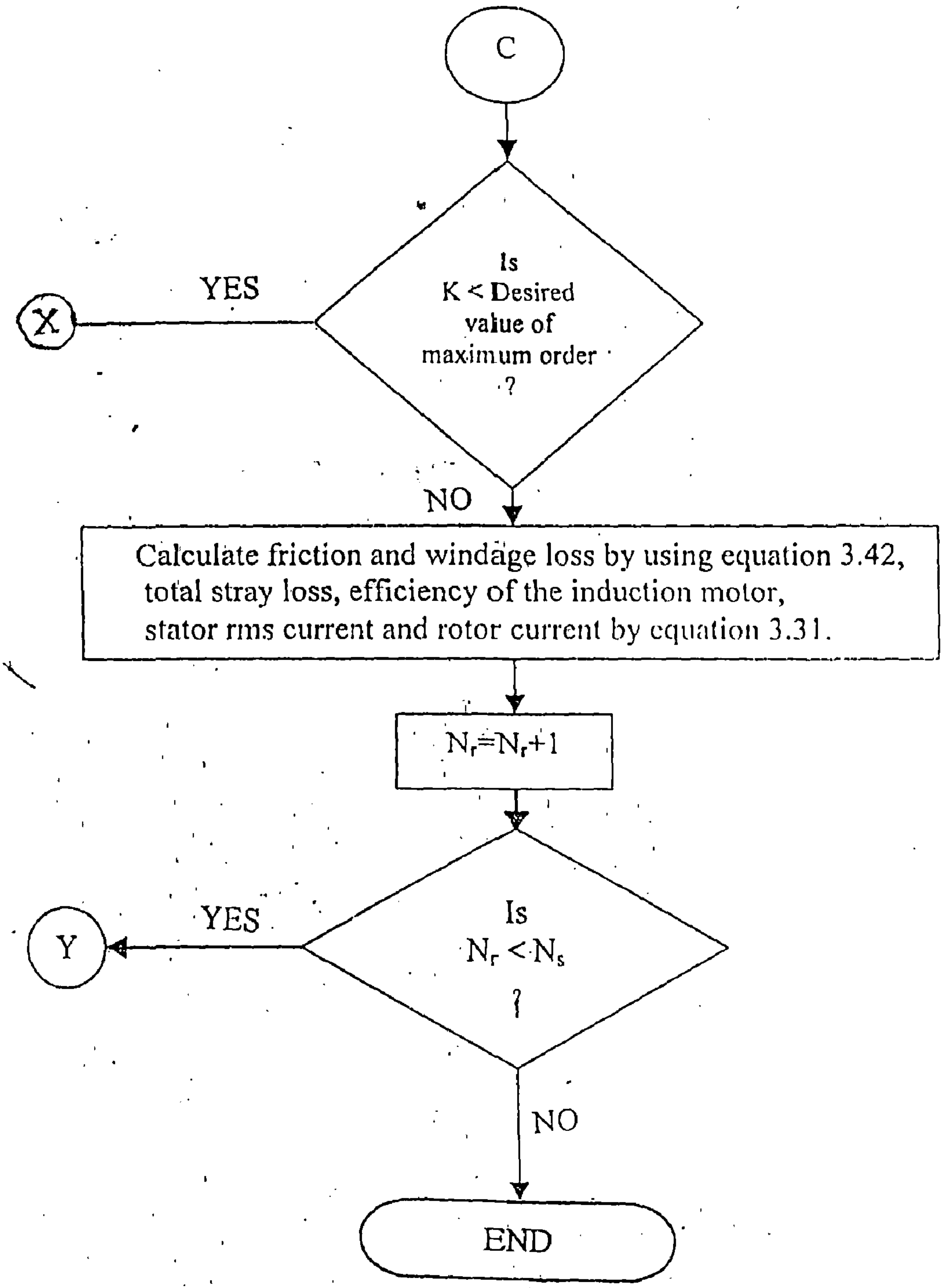
G10,121

Fig. 3.8 FLOW CHART FOR THE PERFORMANCE EVALUATION OF THE INDUCTION MOTOR FED WITH NON-SINUSOIDAL SUPPLY




A
YES
Is K=1 ?
NO
Calculate m3, m4, , iron loss (TIL1) by eqn 3.2, 3.3, 3.41 respectively.
First calculate R2k, Sk (for concerned phase sequence) then calculate m3, m4, R(k), X(k),iron loss by eqns 3.38, 3.24,3.2, 3.25, 3.26, 3.40 respectively.
Choose the amplitude of the voltage of Kth order harmonic
Calculate power factor, I1, I2, PSC1, PRC1 by equation 3.6, 3.10, 3.14, 3.15 respectively.
I1(K)= I1K2
I2(K)= I2K2
I1=I1(K)+I1
I2=I2(K)+I2
PSC=PSC1+PSC
PRC=PRC1+PRC
TIL=TIL1+TIL
B


B
Calculate input power PIN1, air-gap power P_G by equations 3.13, 3.11.
Check for the phase sequence of the Kth order harmonic in the waveform.
Is phase sequence is same as in fundamental ?
YES
NO
Calculate torque by equation 3.12 Te= +Te1
Calculate toque by equation 3.12 Te= -Te1
Calculate total torque Te=Te1+Te
Calculate total power input PIN=PIN1+PIN
C


C
YES
X
Is
K < Desired
value of
maximum order
?
NO
Calculate friction and windage loss by using equation 3.42,
total stray loss, efficiency of the induction motor,
stator rms current and rotor current by equation 3.31.
N_r=N_r+1
YES
Y
Is
N_r < N_s
?
NO
END

RESULTS AND DISCUSSION

In order to investigate the performance of an induction motor under sinusoidal and non-sinusoidal supplies, a 5HP, 3-phase, delta connected induction motor is taken. The motor parameters are given in the Appendix. In this chapter a discussion on the performance of the induction motor fed by sinusoidal and non-sinusoidal supplies is given with the help of various characteristics. These characteristics are drawn with the results obtained by running the C language programs, which are based on the flow charts shown in the chapter-3. The results are also tabulated for different non-sinusoidal sources.

In the induction motor control, the static switching devices are used. Due to the presence of the harmonics within the output of these switching devices, the motor performance is deteriorated. These harmonic voltages produce the harmonic currents. So the total rms current of the motor is increased. The variation of the stator current with the speed is shown in Fig. 4.1 for both sinusoidal and non-sinusoidal supplies. To judge the clear difference, an another curve is shown in Fig. 4.2 at the full load condition. It is observed in Fig. 4.2 that an increase in the number of steps significantly decreases in the stator current. So by increasing the number of steps, the non-sinusoidal supply for the motor reaches near to sinusoidal supply. It can be seen that the increase in the number of the steps considerably improves the stator current waveform.

The increase in the stator current causes the additional losses. The increase in the stator and rotor copper losses are shown in Fig. 4.3 & in Fig. 4.4 respectively. From these figures it is concluded that the additional losses can be decreased by reducing the harmonic components of the voltage waveform by using a multi-stepped or PWM inverters. The difference between the rotor copper losses due to the six-step supply and twelve-step supply is more than the stator copper losses in both cases due to increase in the rotor resistance. The rotor resistance increases due to the skin effect. The variation of the rotor resistance is shown in Fig. 3.5 in the chapter-3. The friction and windage losses are 57.09W at the base speed and they have been assumed to vary with the speed.

The time harmonics have detrimental effects on the average steady-state torque of the motor. A combined torque-speed characteristic for both types of supplies is shown in Fig. 4.5. Fig. 4.6 shows the variation of the torque with the speed at full load for both sinusoidal and non-sinusoidal fed induction motor. It is cleared from Fig. 4.6 that the time harmonics have very small variation in the torque when supplied by non-sinusoidal supplies. The torque conditions at different supplies are listed in the Table-3. The bold digits represent the torque at full load.

The harmonic contents of the voltage waveform have a direct bearing on the motor efficiency. Efficiency curves are shown in Fig. 4.7 & in Fig. 4.9. The motor efficiency, which is 82.049% in the sine wave, is improved from 81.093% in the six-step case, to 81.850%, in the twelve-step case. In the efficiency curve shown in the Fig.4.9 harmonic injection PWM technique provides good efficiency, which is near to the efficiency obtained by sinusoidal fed induction motor, than the modified sinusoidal PWM and sinusoidal PWM techniques. It is concluded that with the use of multi-stepped and modified PWM voltage source inverters, the motor efficiency can be achieved as good as for the sinusoidal fed induction motor. Further increase in the number of steps has little effect on the motor efficiency. Comparison between performance of the induction motor fed by stepped and PWM waveforms, shows that the torque, stator current waveform are little improved by PWM than by stepped waveform. This is because:

(a) for SPWM, the waveform has no 5^{th},7^{th},11^{th},13^{th},17^{th},25^{th},29^{th},31^{st} harmonics. On the other hand, the 19^{th}, 23^{rd} harmonic increases to 32% from 5.2% and from 4.3% compared with six-step and twelve-step waveforms respectively.

(b) for MSPWM, the waveform has no 5^{th},7^{th},11^{th},13^{th},29^{th},31^{st} harmonics. On the other hand, 17^{th} & 25^{th} harmonic increases to 11% from 5.8% & from 4.0% respectively and 19^{th}, 23^{rd} harmonic increases to 26% from 5.2% and from 4.3% compared with six-step and twelve-step waveforms respectively.

(c) for HIPWM, the waveform has no 5^{th},7^{th},11^{th};13^{th},29^{th},31^{st} harmonics. On the other hand, 17^{th}, 25^{th} harmonic increases to 12% from 5.8% & from 4.3% respectively and 19^{th}, 23^{rd} harmonic increases to 24% from 5.2% and from 4.3% compared with six-step and twelve-step waveforms respectively.

The efficiencies at the sinusoidal, six-step and twelve-step supplies are given in the Table-2. The efficiencies at full load are written in italics. It is cleared that the difference between the efficiencies of the motor for sinusoidal and non-sinusoidal supplies is about 2%.

The behaviour of the induction motor at a particular harmonic as well as at the fundamental is given in the Table-4. In the Table-1 (a) & 1(b) various harmonics, which are present in the non-sinusoidal supplies, are presented with their amplitudes.

Due to the non-sinusoidal supply, the induction motor is excited by a series of the voltage sources. Since each harmonic voltage source produces its individual effects on the motor. So these effects can be reduced by eliminating the lower order harmonics. Lower order harmonics elimination can be achieved by the multi-stepped or modified PWM inverters. Then the motor performance can be upgraded which might be near to that for sinusoidal supply.

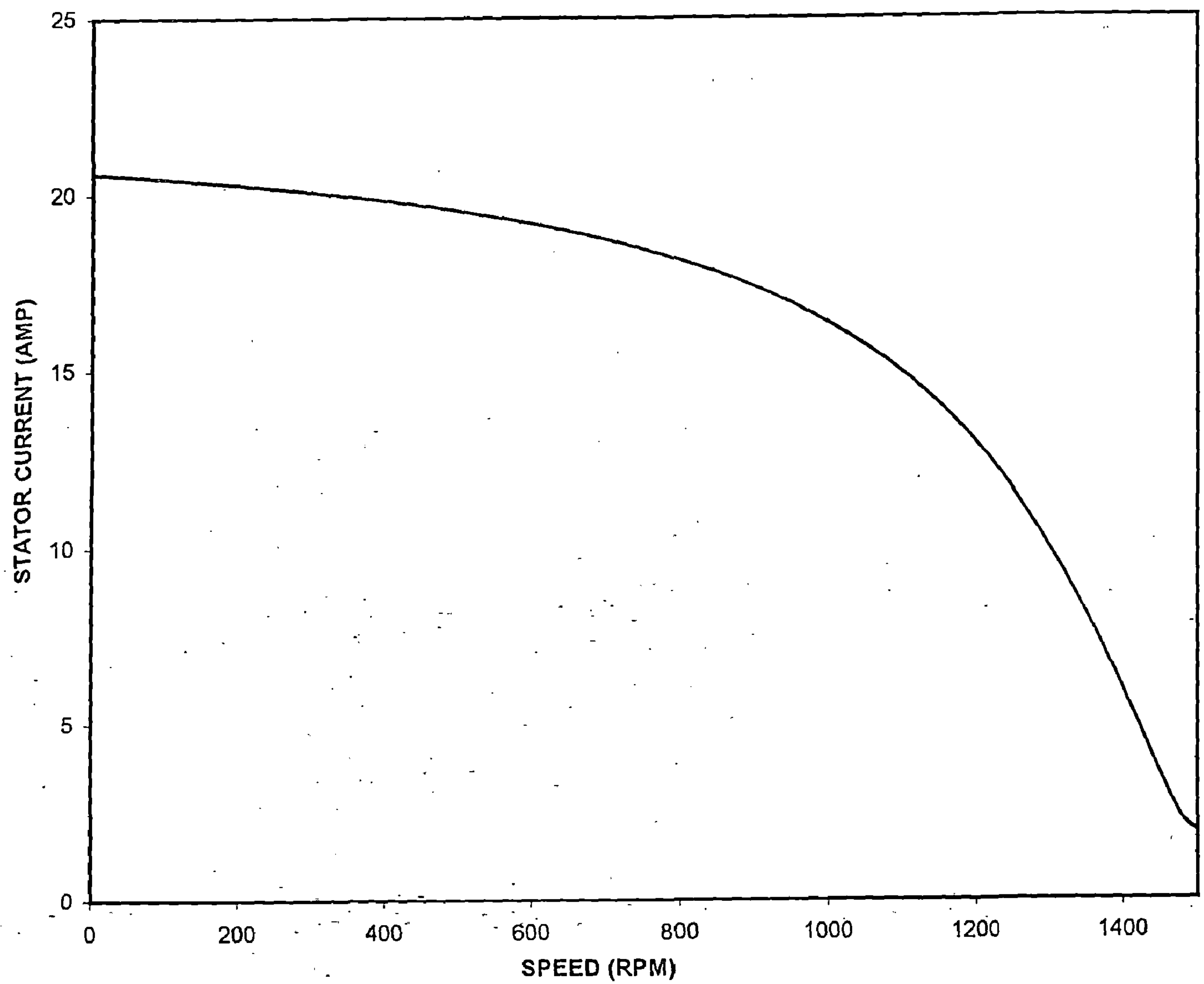


Fig. 4.1 : STATOR CURRENT VARIATION WITH THE MOTOR SPEED

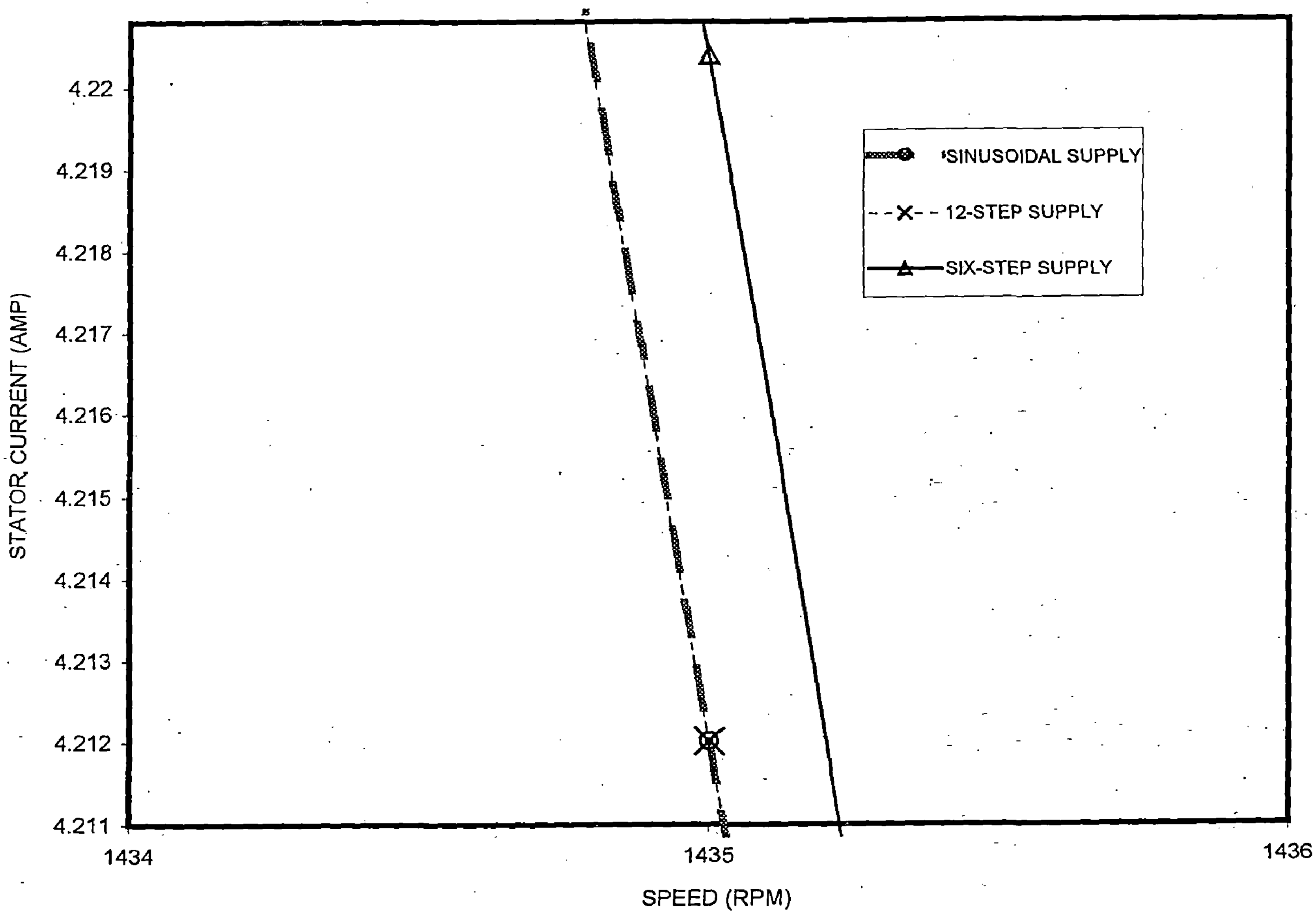


Fig. 4.2: STATOR CURRENT VARIATION WITH THE MOTOR SPEED AT FULL LOAD

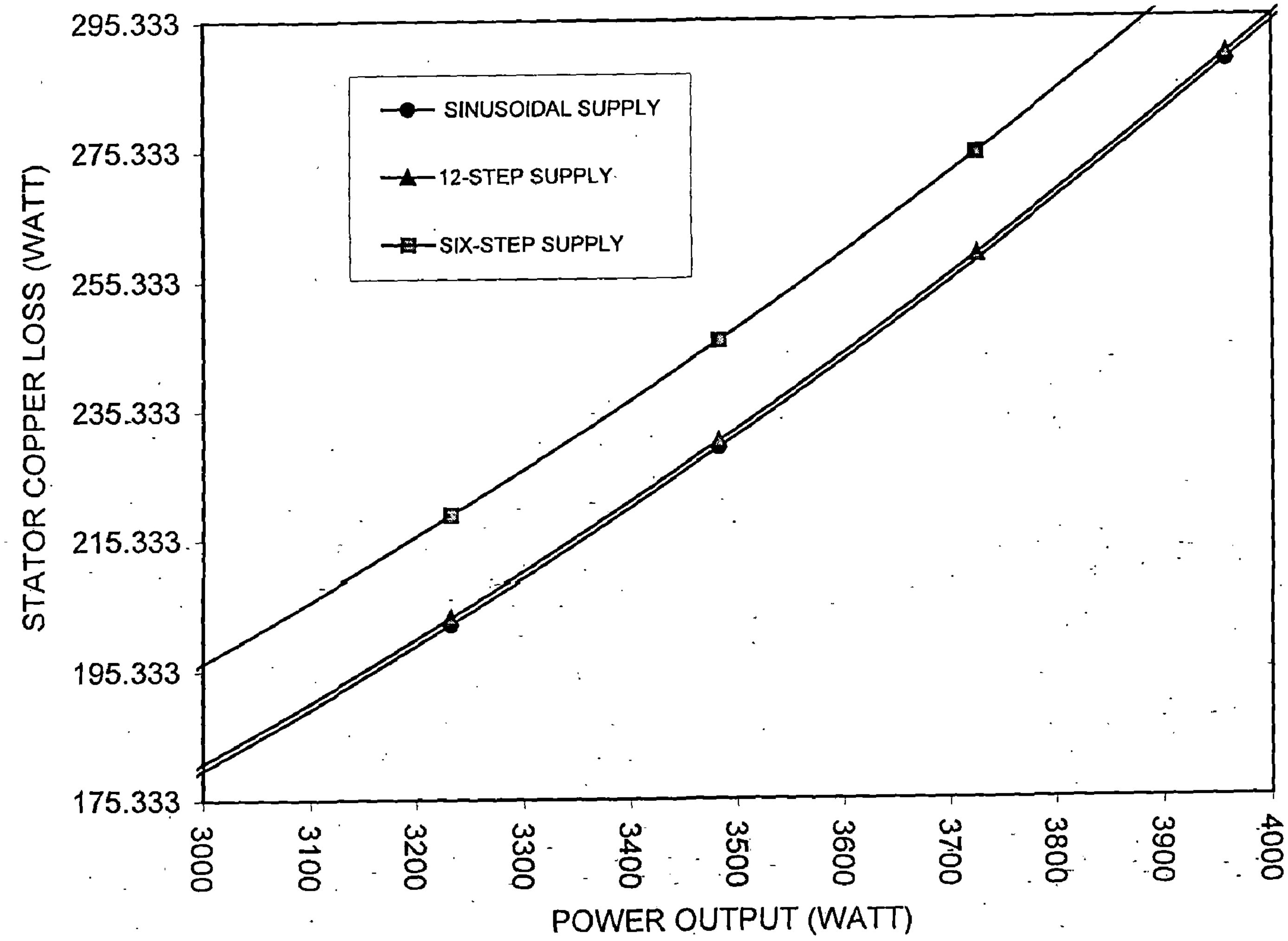


Fig. 4.3: VARIATION OF STATOR COPPER LOSS WITH POWER OUTPUT

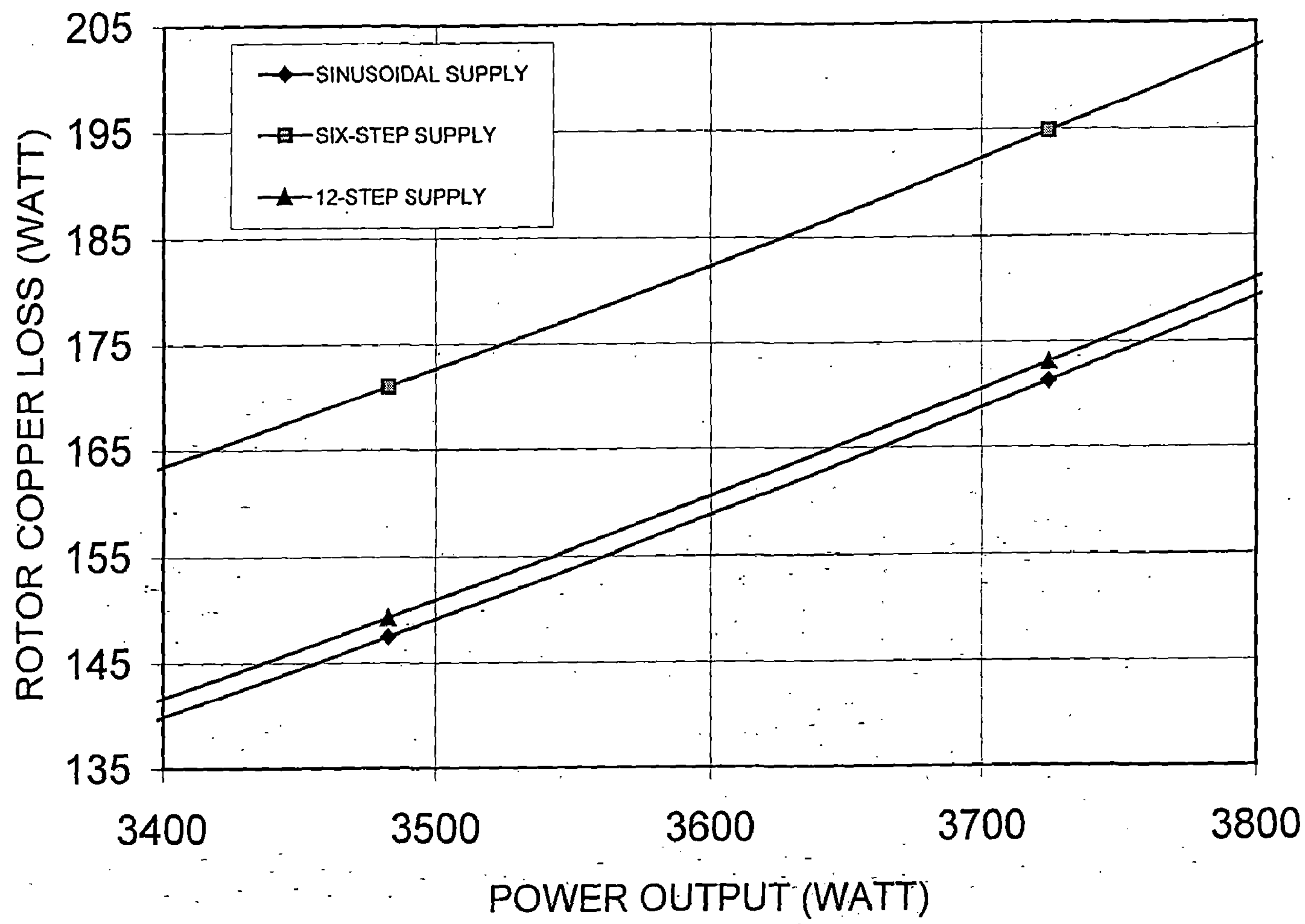


Fig.4.4: VARIATION OF ROTOR COPPER LOSS WITH POWER OUTPUT

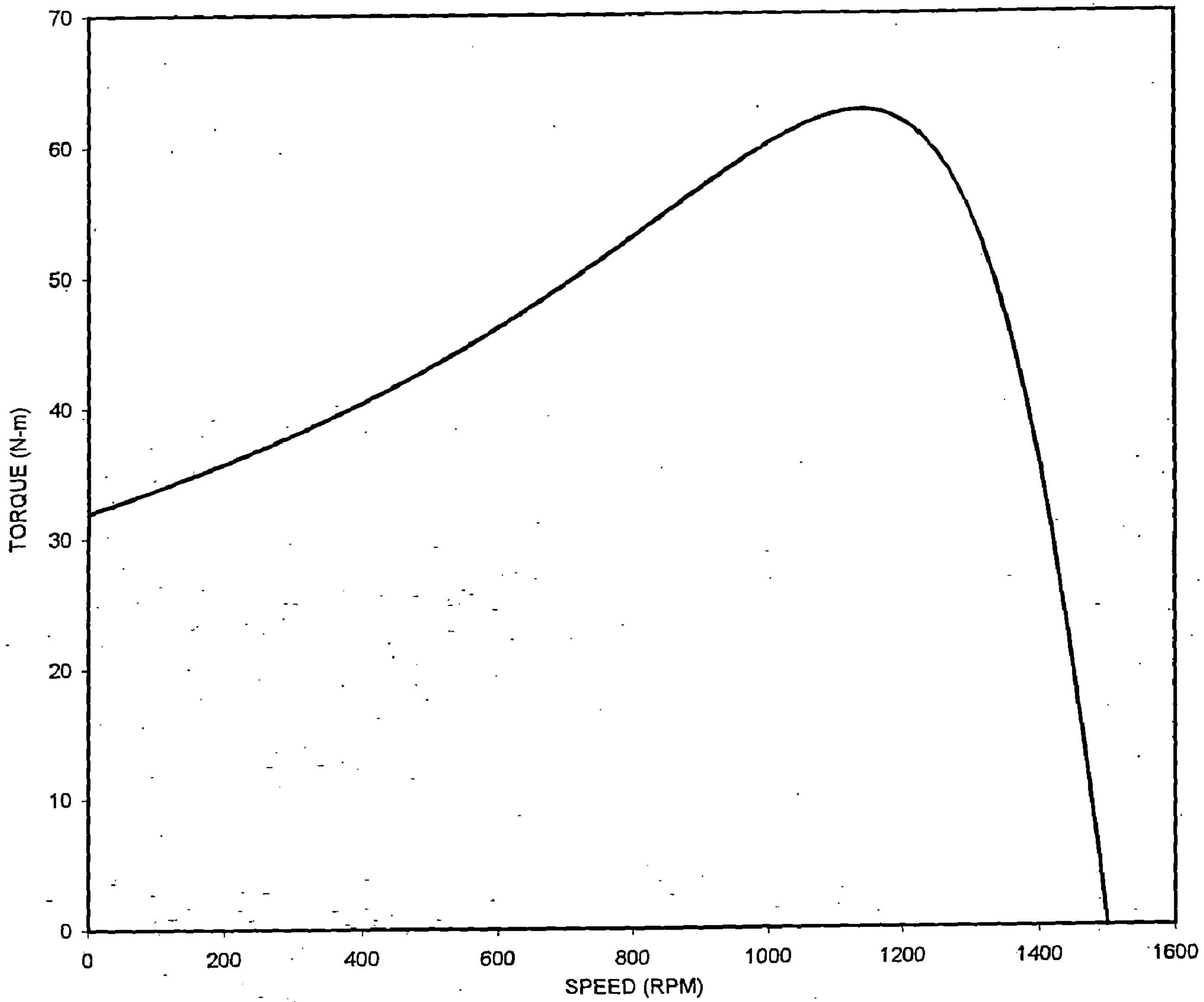


FIG45 : TORQUE-SPEED CHARACTERISTIC OF THE INDUCTION MOTOR

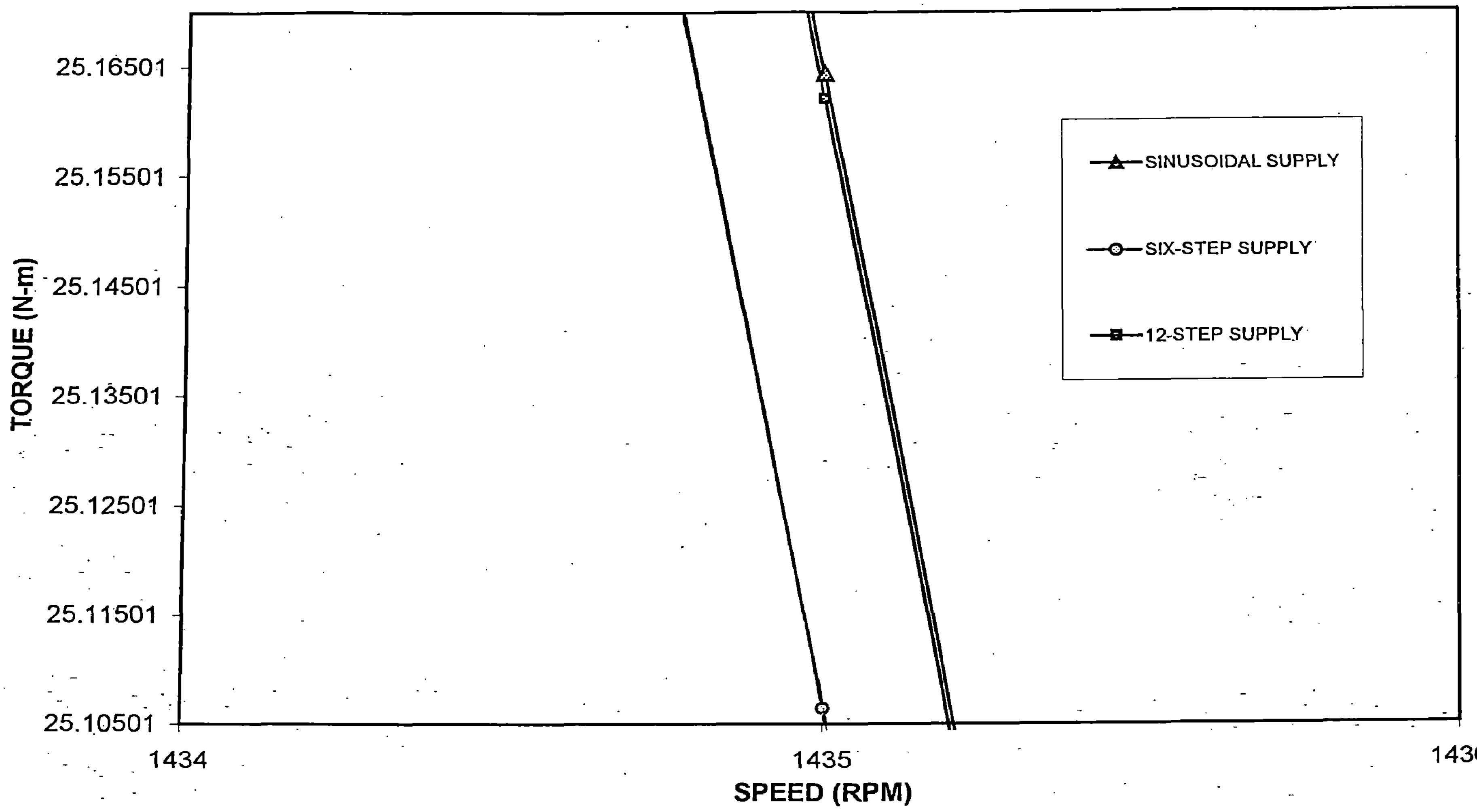


Fig. 4.6 : TORQUE-SPEED CHARACTERISTIC OF THE INDUCTION MOTOR AT FULL LOAD

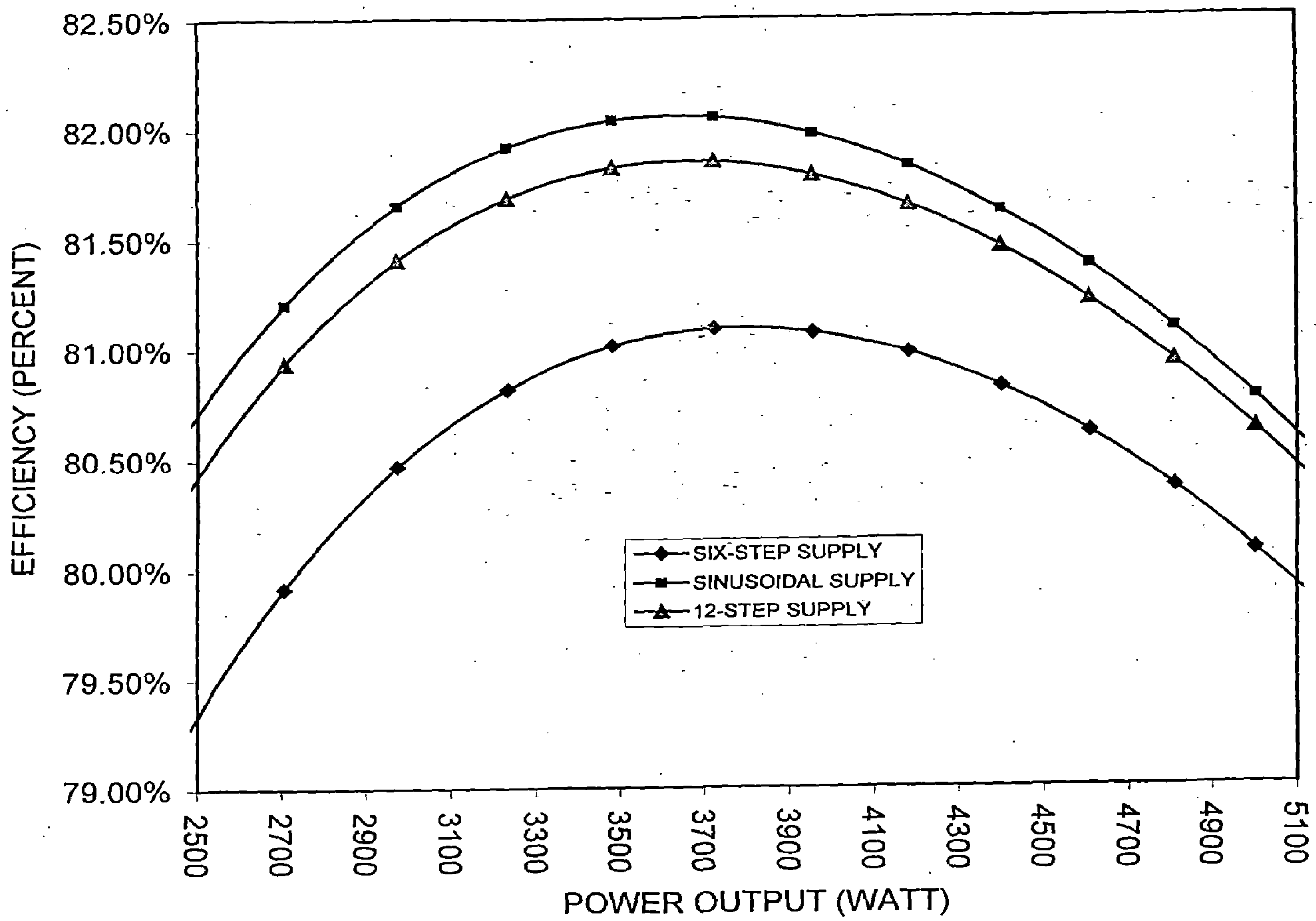


Fig 4.7 EFFICIENCY CURVES AT DIFFERENT SUPPLIES

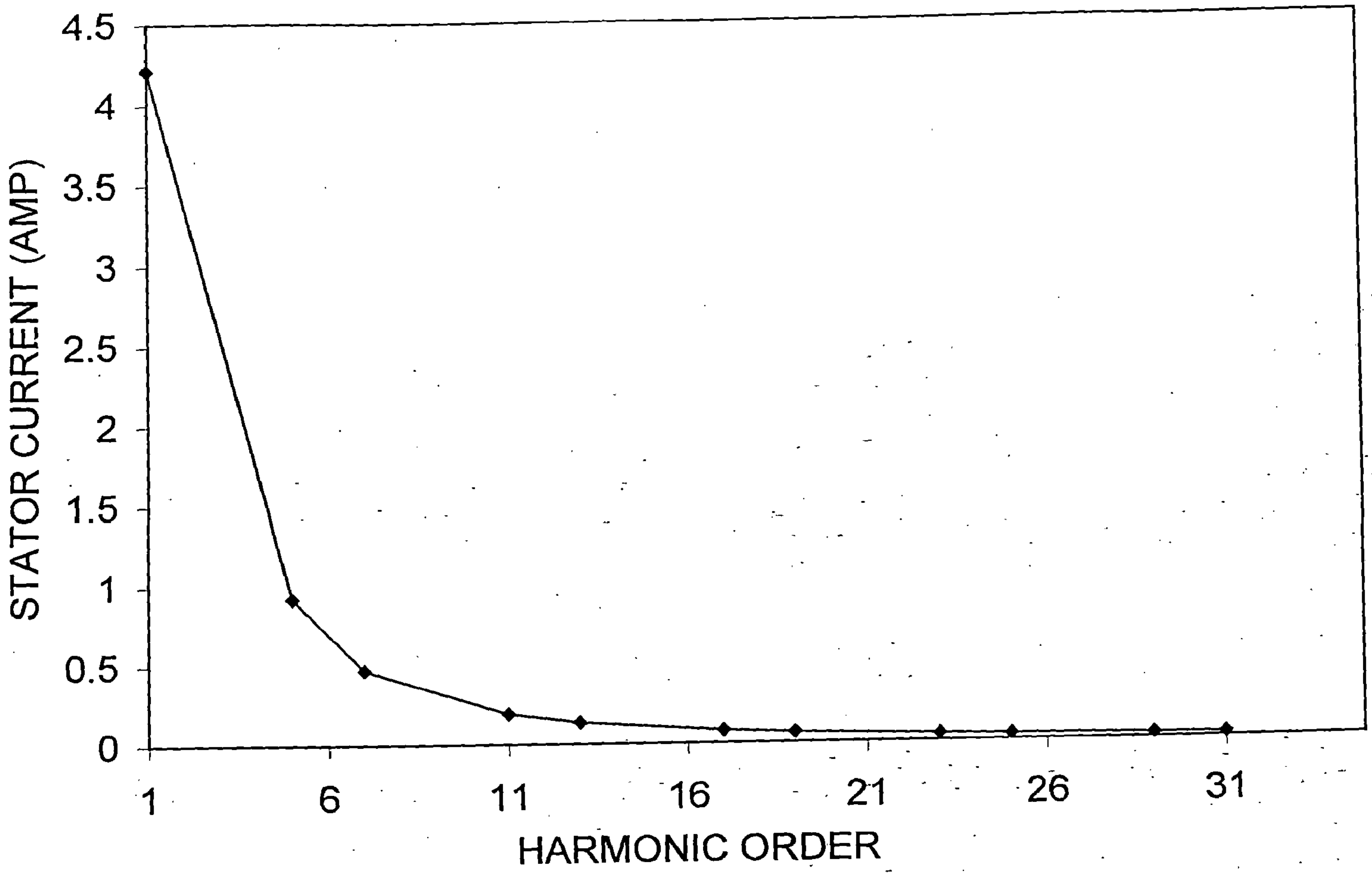


Fig. 4.8: STATOR CURRENT UNDER VARIOUS HARMONIC ORDERS

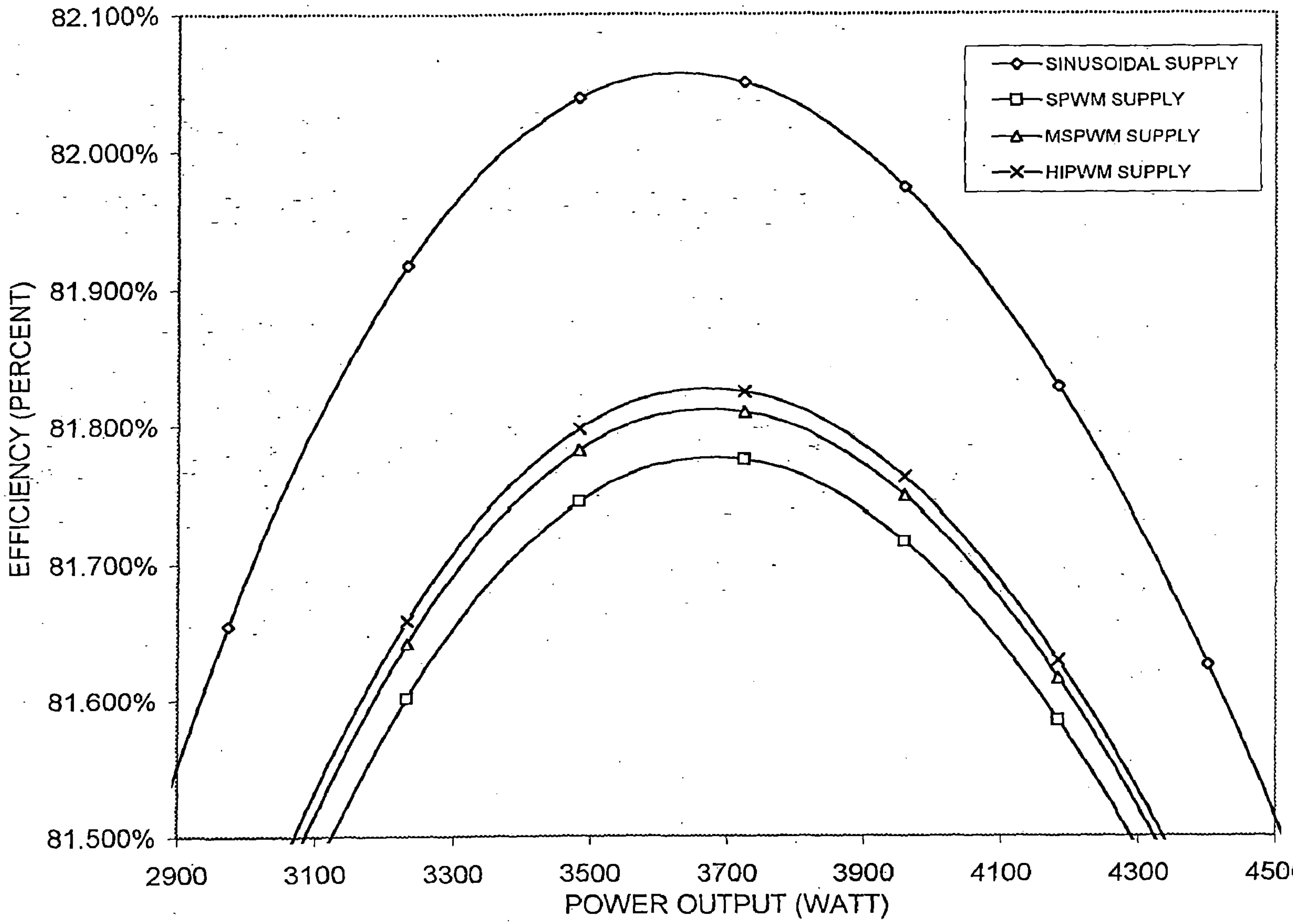


Fig. 4.9: EFFICIENCY CURVES AT DIFFERENT SUPPLIES

HARMONIC CONTENT OF VOLTAGE WAVEFORMS

Table: 1(a)

Harmonic Order	Amplitude (per unit)	
	Six-step Waveform	Twelve-step Waveform
1	1	1
-5	0.200	-
7	0.142	-
-11	0.090	0.090
13	0.076	0.076
-17	0.058	0.058
19	0.052	0.052
-23	0.043	0.043
25	0.040	0.040
-29	0.034	0.034
31	0.032	0.032

Table: 1(b)

Harmonic Order	Amplitude (per unit)		
	SPWM	MSPWM	HIPWM
1	1	1	1
-5	-	-	-
7	-	-	-
-11	-	-	-
13	-	-	-
-17	-	0.11	0.12
19	0.32	0.26	0.24
-23	0.32	0.26	0.24
25	-	0.11	0.12
-29	-	-	-
31	-	-	-

Table: 2

VARIATION OF THE STATOR COPPER LOSS, ROTOR COPPER LOSS AND EFFICIENCY OF THE INDUCTION MOTOR WITH LOAD FOR DIFFERENT SUPPLIES

Stator Copper Loss (Watt)			Rotor Copper Loss (Watt)			Efficiency		
Sinusoidal Supply	Six-step VSI Supply	12-step VSI Supply	Sinusoidal Supply	Six-step VSI Supply	12-step VSI Supply	Sinusoidal Supply	Six-step VSI Supply	12-step VSI Supply
288.664	305.274	289.692	196.536	220.089	198.315	81.973	81.071	81.786
282.420	299.030	283.448	191.374	214.927	193.153	81.995	81.082	81.805
276.244	292.854	277.272	186.268	209.821	188.046	82.013	81.089	81.821
270.137	286.747	271.165	181.218	204.771	182.996	82.028	81.090	81.834
264.099	280.709	265.127	176.214	199.777	178.003	82.041	81.091	81.843
258.132	*274.742*	*259.160*	*171.288*	*194.841*	*173.067*	*82.049*	*81.093*	*81.850*
252.236	268.846	253.264	166.410	189.963	168.189	82.055	81.094	81.852
246.411	263.021	247.439	161.590	185.144	163.369	82.055	81.091	81.851
240.659	257.269	241.687	156.830	180.383	158.608	82.054	81.089	81.847
234.980	251.590	236.008	152.129	175.682	153.907	82.048	81.081	81.838
229.375	245.985	230.403	147.488	171.041	149.267	82.038	81.077	81.825
223.844	240.454	224.872	142.908	166.461	144.687	82.023	80.984	81.807
218.388	234.998	219.036	138.390	161.943	140.168	82.004	80.950	81.785
213.008	229.619	214.036	133.933	157.486	135.711	81.983	80.911	81.755
207.705	224.315	208.733	129.539	153.092	131.317	81.954	80.866	81.727
202.479	219.089	203.507	125.207	148.761	126.986	81.912	80.816	81.687
197.331	213.941	198.359	120.940	144.493	122.718	81.873	80.760	81.644
192.261	208.872	193.289	116.736	140.290	118.515	81.839	80.697	81.594
187.271	203.881	188.299	112.598	136.151	114.376	81.773	80.622	81.538
182.361	198.971	183.389	108.524	132.078	110.303	81.725	80.557	81.476
177.531	194.142	178.559	104.517	128.070	106.265	81.659	80.461	81.406

Table: 3

VARIATION OF THE STATOR CURRENT AND TORQUE WITH LOAD FOR DIFFERENT SUPPLIES

Speed N_r (RPM)	Stator current (Amp)			Torque (N-m)		
	Sinusoidal Supply	Six-step VSI Supply	12-step VSI Supply	Sinusoidal Supply	Six-step VSI Supply	12-step VSI Supply
1430	4.454	4.580	4.462	26.811	26.753	26.808
1431	4.405	4.533	4.413	26.485	26.427	26.483
1432	4.357	4.486	4.365	26.157	29.099	26.155
1433	4.308	4.439	4.317	25.828	25.770	25.826
1434	4.260	4.392	4.268	25.497	25.439	25.495
1435	**4.212**	**4.345**	**4.220**	**25.164**	**25.106**	**25.162**
1436	4.163	4.298	4.172	25.829	24.771	24.827
1437	4.115	4.251	4.123	24.493	24.435	24.491
1438	4.066	4.204	4.075	24.155	24.097	24.152
1439	4.018	4.158	4.027	23.815	23.757	23.812
1440	3.970	4.111	3.979	23.473	23.415	23.471
1441	3.922	4.065	3.931	23.130	23.072	23.127
1442	3.874	4.018	3.883	22.784	22.727	22.782
1443	3.826	3.973	3.835	22.438	22.380	22.435
1444	3.778	3.926	3.787	22.089	22.031	22.087
1445	3.730	3.880	3.739	21.739	21.681	21.736
1446	3.682	3.834	3.692	21.386	21.329	21.384
1447	3.635	3.788	3.644	21.033	20.975	21.030
1448	3.587	3.743	3.597	20.677	20.619	20.675
1449	3.540	3.697	3.550	20.320	20.262	20.318
1450	3.493	3.652	3.503	19.961	19.903	19.959

TABLE:4

OPERATION FROM VOLTAGE SOURCE INVERTERS

Harmonic Order	Stator Phase Current (Amp.)	Stator Copper Loss (Watt)	Rotor Copper Loss (Watt)	Core Loss (Watt)
1	4.212	258.132	171.288	251.076
5	0.921	12.356	17.004	3.193
7	0.470	3.225	4.770	1.629
11	0.191	0.053	0.878	0.659
13	0.137	0.273	0.470	0.472
17	0.080	0.093	0.173	0.276
19	0.064	0.060	0.114	0.221
23	0.043	0.027	0.056	0.150
25	0.037	0.020	0.041	0.127
29	0.027	0.011	0.024	0.094
31	0.024	0.008	0.018	0.083

CONCLUSION

The steady-state performance of the induction motor is analyzed by Fourier analysis. The motor performance is carried out for two commonly used adjustable frequency controllers, voltage source inverter and for pulse width modulated inverter. For voltage source inverter, two cases namely six-step, twelve-step voltage source inverters and for pulse width modulated inverter, sinusoidal pulse width modulated (SPWM), modified sinusoidal pulse width modulated (MSPWM), and harmonic injection pulse width modulated (HIPWM) inverters are considered.

A comparison between the motor performance under sinusoidal and non-sinusoidal supplies (six-step, twelve-step & PWM voltage source inverters) is made. The following results are obtained:

1. The detrimental effect of harmonics on the motor performance is substantially decreased by increasing the number of steps of the inverter output voltage from six to twelve.
2. Modified sinusoidal PWM and harmonic injection PWM technique provide better performance of the motor than the sinusoidal PWM technique.
3. The overall average torque of the motor due to harmonics is calculated. It is observed that there is a slight reduction (less than 1.0%) in steady state torque because the average torques due to the positive-sequence harmonic currents tend to cancel the average torques due to the negative-sequence harmonic currents.

4. For a given fundamental supply frequency, harmonic currents and motor losses are higher in the reduced torque output region when motor is fed by non-sinusoidal supply.
5. The efficiency of the non-sinusoidal fed induction motor can be improved by using multi-stepped and PWM waveforms. Further increase in the number of steps has little effect on the motor efficiency.

Scope for future work

Further work of this problem can be taken up on the following lines:

1. Instead of the Fourier analysis in frequency domain, the time domain analysis whereby all quantities are obtained explicitly and not as the sum of infinite series can be applied for the performance investigation of the induction motor when fed by non-sinusoidal supply.
2. In the present work, iron and stray losses are considered constant, so by applying design parameters, performance of the motor can be evaluated for about the exact value of the iron and stray losses. Optimization of motor losses can be carried out.
3. One of the major problems in inverter-fed motors is the high level of audible noise produced by harmonic current and voltage components. So analysis can be done to predict the spectrum components produced by the motor and to relate it to the air-gap flux density distribution time harmonics caused by the non-sinusoidal supply.
4. The constant-speed performance of the induction motor with applied stator phase voltages of any periodic form can also be evaluated by the method of multiple reference frames and also a digital model can be developed, which is particularly adapted for studying its dynamic performance when fed from an inverter.

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APPENDIX

APPENDIX

The data given below pertains to a 5 H.P, 400 volts, 3-phase, 50 Hz, 4-pole, delta connected squirrel-cage induction motor taken for the performance investigation [20],

Sl. No.	Input Quantity	Symbol	Value
1	Stator per phase resistance	R_1	4.85 ohm
2	Stator per phase leakage reactance at normal frequency	X_1	8.80 ohm
3	Rotor per phase resistance referred to stator	R_2	4.30 ohm
4	Rotor per phase leakage reactance at the normal frequency referred to stator	X_2	8.80 ohm
5	Magnetizing reactance at normal frequency	X_m	200 ohm
6	Maximum stator core density	B_{cm}	1.44 Wb/m^2
7	Maximum stator teeth density	B_{tm}	1.68 Wb/m^2
8	Stator core weight	SCW	8.75 Kg
9	Stator teeth weight	STW	3.15 Kg