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Calorimetric measurement of induction motor harmonic losses

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Calorimetric Measurement of Induction Motor Harmonic Losses

A thesis submitted in fulfilment of the requirements for the award of the degree

PhD

from

University of Wollongong

by

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School of Electrical Computer and Telecommunication Engineering

June 1997
In the name of God, the merciful and compassionate
Declaration

This is to certify that the work presented in this thesis is entirely my own and has not been submitted for any other degree.

Alireza Jalilian

June 1997
To my wife and to my dearest daughters, Elham and Farhaneh...
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Wollongong, June 1997

Alireza Jalilian
# Table of Contents

Declaration ........................................................................................................ ii

Acknowledgments ............................................................................................. iv

Table of Contents ............................................................................................... v

List of Figures .................................................................................................... viii

List of Tables ...................................................................................................... x

Abstract ............................................................................................................... xi

1. Introduction
   1.1 Introduction .............................................................................................. 1
   1.2 Objectives and Goals .............................................................................. 2
   1.3 Outline of the Thesis .............................................................................. 4

2. Background: Induction Motor Losses
   2.1 Introduction .............................................................................................. 7
   2.2 Space and Time Harmonics .................................................................. 8
   2.3 Rotating MMFs due to the Time Harmonics ........................................... 9
   2.4 Fundamental Equivalent Circuit ........................................................... 11
   2.5 Harmonic Behaviour of Induction Motors .............................................. 13
      2.5.1 Equivalent Circuit ........................................................................... 13
      2.5.2 Rotor deep bar effect ....................................................................... 15
      2.5.3 Stator winding resistance variation (R1n) ....................................... 16
      2.5.4 Rotor resistance variation (R2n) ..................................................... 17
      2.5.5 Stator and rotor leakage reactance ............................................... 18
   2.6 Losses in Polyphase Induction Motors ..................................................... 21
      2.6.1 Stator winding losses ...................................................................... 24
      2.6.2 Rotor winding losses ...................................................................... 25
      2.6.3 Core losses ..................................................................................... 26
      2.6.4 Stray load losses ............................................................................ 30
      2.6.5 Friction and windage losses ........................................................... 33
   2.7 Standard Methods for Determining Losses ............................................ 33
   2.8 Calorimetric Method .............................................................................. 35
   2.9 Conclusions ............................................................................................. 37
3. Calorimeter Design to Measure Induction Motor Losses

3.1. Introduction ................................................................. 38
3.2. Principle of the Calorimetric Method ................................ 39
3.3 Open and Closed Type Calorimeters .................................. 40
3.4 Balance Calorimetric Method ........................................... 42
3.5 Double Chamber Calorimeter (DCC) .................................... 44
  3.5.1 Heat transfer mechanism within the DCC ......................... 45
  3.5.2 Design and construction of the DCC ............................... 49
  3.5.3 Reference heater ........................................................ 50
  3.5.4 Loaded machine mechanism ......................................... 51
  3.5.5 Instrumentation and measurement system ....................... 53
    3.5.5.1 Temperature measurement system ........................... 53
      A. Absolute temperature measurement using RTDs ............... 53
      B. Relative temperature measurement using thermopiles ........ 54
      C. Calibration of thermopiles ....................................... 55
  3.5.5.2 Voltage and current measurement circuits ................... 59
3.6. Conclusions ............................................................... 60

4. Double Chamber Calorimeter: Test and Calibration

4.1 Introduction ......................................................................... 62
4.2 Conducted Heat Leakage ..................................................... 63
  4.2.1 Heat Conduction through a Plane Slab ............................. 63
  4.2.2 Heat Conduction through the extended shaft ................... 64
4.3 Measurement of Calorimeter Heat Leakage ............................. 65
  4.3.1 Test Procedure and test results .................................... 66
  4.3.2 Heat leakage through the calorimeter edges and corners .... 67
4.4 Calorimeter Calibration using Two Identical Heaters ............... 69
  4.4.1 Test Procedure ............................................................ 70
  4.4.2 Temperature distribution and heat leakage calculation ...... 71
  4.4.3 Air temperature rise across each chamber ..................... 73
  4.4.4 Estimation of dissipated heat by the test heater ................ 75
  4.4.5 Deriving limits for the DCC ........................................... 77
4.5 Error Analysis and Accuracy of the Loss Measurement ............. 81
4.6 Conclusions ...................................................................... 82

5. Induction Motor Harmonic Tests

5.1 Introduction .......................................................................... 84
5.2 Test Induction Motor .......................................................... 85
  5.2.1 Initial tests ...................................................................... 85
List of Figures

Figure 1.1: Illustration of Power Quality Testing Facility 3
Figure 1.2: Double chamber calorimeter developed for accurate measurement of induction motor losses 5
Figure 2.1: Conventional per phase equivalent circuit of the induction motor operating under fundamental frequency 12
Figure 2.2: Approximate single phase equivalent circuit of the induction motor 12
Figure 2.3: Single phase equivalent circuit corresponding to the nth harmonic order 14
Figure 2.4: Harmonic equivalent circuit including stray load loss resistor 14
Figure 2.5: Different rotor bar shapes of an induction motor [Buc84] 16
Figure 2.6: Variation of nominal leakage inductance with harmonic order 20
Figure 2.7: Effective total machine resistance and leakage reactance as a function of harmonic order [Wil82] [Cum86] 21
Figure 2.8: Variation of induction motor losses versus load under ideal supply conditions [Cum81] 22
Figure 3.1: Open and closed type calorimeters 41
Figure 3.2: Conventional heat transfer mechanism within the Double Chamber Calorimeter (DCC) 46
Figure 3.3: Schematic diagram of the constructed double chamber calorimeter (DCC) housing the test motor and the reference heater 50
Figure 3.4: Schematic diagram of the calorimeter for loaded machine tests 52
Figure 3.5: A simple arrangement of thermocouples to form a thermopile 55
Figure 3.6: Experimental data and fitted line for thermopile slope (mV/°C) as a function of average working temperature, $T_{avg}$ 58
Figure 3.7: Induction motor voltage and current measurement circuits 59
Figure 4.1: Experimental setup for measurement of calorimeter conducted heat leakage through the walls 65
Figure 4.2: Measured ($P_{in}$) and calculated ($q_{total}$) conducted heat leakage through the calorimeter vs temperature difference ($\Delta T$) 69
Figure 4.3: Double chamber calorimeter (DCC) housing the test and reference heaters for calibration 70
Figure 4.4: Estimated heat leakage through the calorimeter chambers at different heater input power levels and various air flow rates 72
Figure 4.5: Air temperature rise across chamber 1 vs test heater input power at different air flow rates 74
Figure 4.6: Absolute (top) and percentage (bottom) error between the estimated and measured losses in the test heater with $P = 200-500$ W and air flow rate = 55 L/s 76
Figure 5.1: Un-scaled supply line-to-line voltage waveform
Figure 5.2: Approximate single phase equivalent circuit for the test induction motor
Figure 5.3: LC power filter connected at the output terminals of the HG
Figure 5.4: Un-scaled fundamental voltage waveform produced by the HG and measured at the output of the filter
Figure 5.5: Un-scaled voltage waveform containing 10% of 11th harmonic produced by the HG and measured at the output of the filter
Figure 5.6: Un-scaled motor voltage and current waveforms containing the 5th harmonic
Figure 6.1: Calculated values of total machine resistance for different harmonic tests and under different loading conditions
Figure 6.2: Experimental data for total machine resistance $R_n$ and the best fitted curves using Equation (6.2)
Figure 6.3: Experimental values for $R_n$ at half load and the fitted curves according to the error bars
Figure 6.4: Experimental data and fitted curves for $R_n$ vs harmonic order under different loading conditions,
Figure 6.5: Variation of $Z_n$ with harmonic order at different tests
Figure 6.6: Experimental and estimated values for $X_n$ at different tests
Figure 6.7: Experimental values and fitted curve for total leakage inductance vs harmonic order at different tests
List of Tables

Table 3.1: Calibration of thermopiles using mercury-in-glass thermometers 57
Table 4.1: Measured and calculated values for the calorimeter conducted heat leakage in different tests 68
Table 4.2: Limits for different parameters derived for the DCC 80
Table 5.1: Specifications of the test induction motor 85
Table 5.2: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{11} = 370 \) W 94
Table 5.3: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{11} = 385 \) W 96
Table 5.4: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{11} = 500 \) W at half load conditions 97
Table 5.5: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{11} = 1015 \) W at full load conditions 98
Table 5.6: Separation of fundamental losses in the test motor under different loading conditions 99
Table 6.1: Calculated values for \( R_n \) corresponding to different harmonic tests 102
Table 6.2: Calculated values of \( K_R \) and \( a \) to give the best estimate for \( R_n \) in different tests using Equation (6.2) 105
Table 6.3: Calculated values for \( Z_n(=V_n/I_n) \) corresponding to different tests 108
Table 6.4: Calculated values of \( K_X \) and exponent \( c \) to give the best estimate for \( X_n \) in different tests using Equation (6.5) 110
Table 6.5: Typical parameters for three induction motors with different power ratings 117
Table 6.6: Different distorted waveforms having the same THD but different WTHD 118
Table 6.7: Derating factor due to distorted waveforms corresponding to different machines 118
Table 6.8: Maximum allowable value for WTHD to give a DF based on 1.15 service factor 119
Abstract

An original contribution of this project is the design and construction of a new type calorimeter, a double chamber calorimeter (DCC), to directly and accurately estimate total losses of a 7.5 kW induction motor. The DCC is utilised to investigate the additional losses due to the presence of time harmonics in the supply of mains-connected induction motors. The main advantage of using the DCC is that it enables estimation of machine losses independent of the level of the supply distortion and motor loading conditions.

The DCC is made of class VH expanded polystyrene insulation material, large enough to house the test motor and a reference heater for balance type of loss measurement. A variable speed fan is used to maintain the air with sufficient flow rate through the calorimeter to remove the generated heat within the calorimeter chambers. Motor losses are estimated as a function of reference heater input power and the air temperature rise across the calorimeter chambers after thermal equilibrium has been achieved.

One-dimensional conducted heat leakage through the calorimeter walls is estimated by developing a simple loss model for the calorimeter and validated using experimental tests. The model utilises conduction shape factors to evaluate the heat leakage through the calorimeter walls, edges and corners. Dynamic operation of the DCC is examined by performing substantial experimental tests using two identical heaters. Accordingly, limits for the air flow rate through the calorimeter, air temperature rise inside and across each chamber and heater input power are derived. Experimental results confirmed that motor losses up to 1 kW can be estimated using the DCC with a resolution of 10 W and an accuracy of 4%.

Motor line-line voltages and line currents are measured by developing voltage and current measurement circuits. For data collection a PC-based data acquisition (DA) system in conjunction with a computer software package is employed in this project.
The DA system is also used for absolute temperature measurement using RTDs and relative temperature measurement using thermopiles.

A 10 kVA inverter capable of producing harmonically distorted waveforms (up to 1 kHz) is employed to conduct harmonic tests on the test induction motor. Experimental tests are performed under non-distorted (nominal) and various distorted supply conditions and with the motor operating under no load, half load and full load. Experimental results confirmed that a distorted voltage containing low order harmonic causes more losses in a motor when compared with a distorted voltage having a higher order harmonic. A weighted THD (WTHD) is defined to specify the limits for additional losses in a motor supplied by distorted voltages. In terms of loading effect, the additional losses significantly increase with load mainly due to the increased resistance with temperature. Therefore, one can conclude that the harmonic losses are load independent and are constant for a known voltage distortion level except for the temperature effect.

The variation of test motor parameters with harmonic order as well as the variation of additional losses with WTHD has led to establishment of derating factor (DF) for induction motors. Depending upon the supply WTHD, a DF can be determined which suggests the fraction of machine loading under which the additional losses due to the distorted supply can be safely tolerated by the machine. This figure has been calculated using the data for several machines with various power ratings from 3.7 kW to 1.6 MW. The results confirmed that a higher WTHD can be applied to the larger machines as compared with smaller machines. In other words, larger machines are more capable of handling additional losses due to the supply distortion. It has also been demonstrated that most induction motors can afford a WTHD up to 8% if a service factor of 1.15 is applied. The 8% figure corresponds to an average THD of about 15% which is much larger than the commonly used 5% limit for THD in utility power networks as specified by standards.
Chapter 1

Introduction

1.1 Introduction

Three phase cage induction motors are recognised as major power system loads due to their wide use in industry and utilities as well as in home appliances. They are considered to be very reliable and rugged motors having a relatively simple design, low cost and virtually no maintenance. Their application in Variable Speed Drive (VSD) systems make them even more popular particularly as a suitable alternative for DC motors. The development of reliable frequency converters has made it possible to increasingly utilise induction motors in speed control applications such as pumps, fans, compressors, mixers and conveyors. Electric vehicles and traction drive systems are other important applications of inverter-fed induction motors.

Increasing application of power semiconductor switching devices has resulted in a significant presence of time harmonics in the power grid. This has caused mains-connected induction motors to experience distorted waveforms in their supply voltage. Inverter-fed induction motors are subject to even a greater voltage distortion levels since the inverter output voltage contains significant time harmonics with different magnitudes and order.

The presence of time harmonics in the supply of induction motors causes unwanted effects such as additional losses and hot spot temperatures. Additional losses increase heating within different parts of the machine leading to reduced motor efficiency. Hot spot temperatures, however, lead to thermal stress and hence insulation degradation and loss of motor lifetime. In addition to extra losses and overheating, oscillating torques, magnetic noise, mechanical oscillations and rotor bearing currents may occur in induction motors when fed from distorted voltage waveforms [Bur67] [Kli68] [Bon80] [Psh83] [Ema91] [IEC92] [Wag93] [Yac95].
Unlike the fundamental losses, calculation and estimation of harmonic losses in induction motors is not a simple task. This issue has been a subject of research for many years [Jai64] [Kli68] [Mcl69] [Buc79] [Cum86] [Dew90] [Hub93] [Nee93] [Rap93] [Cza94] [Eld95] [Man96] where different approaches have been recommended for the evaluation of motor harmonic losses. Development of harmonic equivalent circuits is one of the commonly used methods to investigate the behaviour of machine under the influence of time harmonics [Cha68] [Cum81].

Assuming linearity, harmonic equivalent circuit parameters can be evaluated at different harmonic frequencies separately and their effects can be added based on the superposition principle. Both rotor resistance and leakage inductance are complex functions of the rotor current frequency [Cum86]. This is basically due to the well-known phenomenon of deep bar effect [Alg51] which results in an increased effective rotor resistance and a reduced leakage inductance as the rotor current frequency increases.

Measurement of additional losses in induction motors using standard methods is subject to difficulties and inaccuracies especially when the machine is loaded and supplied by distorted voltages. Standard laboratory instruments have limited frequency response and are inaccurate under harmonically distorted conditions. Even the motor fundamental losses cannot be measured accurately due to the difficulties involved in the measurement of motor output power.

1.2 Objectives and Goals

One aim of this thesis is to develop a technique to measure losses of a 7.5 kW induction motor precisely, conveniently and regardless of the voltage distortion level supplied to the motor. For this purpose, design and construction of a new type of calorimeter, a double chamber calorimeter (DCC), is suggested [Jal95] [Gos95] [Jal96] [JalI97] [JalII97]. The DCC is part of a Power Quality Testing Facility with an overall arrangement shown in Figure 1.1 which has already been developed for harmonic assessment of common power system loads. The facility consists of a
harmonic generator (HG) [Gos93], a PC-based data acquisition (DA) system and some auxiliary equipment.

![Illustration of Power Quality Testing Facility](image)

**Figure 1.1: Illustration of Power Quality Testing Facility**

The HG is a 3-phase 10 kVA controllable inverter which is employed as the harmonic source to carry out the experimental harmonic tests. This thesis does not give detailed information on the design and implementation of the HG. However, some basic information on its operation is provided in Appendix A.

Another aim is to have a better understanding of induction motor behaviour under harmonically distorted supply conditions. This thesis deals with the effect of time harmonics on additional losses in mains-connected induction motors under different loading conditions. It is intended to experimentally conduct harmonic tests on a 3-phase 7.5 kW high efficiency cage induction motor supplied by distorted waveforms. The emphasis will be placed on the variation of motor additional losses as a function of voltage distortion and harmonic order as well as the motor loading level. The variation of motor parameters with harmonic order will be examined and compared with harmonic loss models which have been presented in the past.

The experimental results will be utilised to derive harmonic limits for induction motors with different power ratings. Also a derating factor will be defined as a function of motor parameters and harmonic voltages, to alleviate the machine from overheating due to the additional harmonic losses. The suitability of power system
harmonic limits recommended by standards and/or enforced by electricity suppliers will also be evaluated in relation to induction motors.

1.3 Outline of the Thesis

A review of induction motor losses both at fundamental and harmonic conditions is presented in Chapter 2. The corresponding equivalent circuits are discussed and different methods for evaluation of motor parameters are described. Different approaches including the standard methods for calculation of fundamental and harmonic losses in induction motors are presented. Common harmonic loss models available in the literature are introduced and those relevant to the scope of this thesis are presented in detail. The variation of motor parameters with harmonic frequency are also described in accordance with skin effect in the stator and rotor conductors.

Description of the calorimetric method suitable for direct measurement of electric machine losses is described in Chapter 3. Open and closed type calorimeters are introduced and their basic operation is highlighted. Details of the design and construction of a new type calorimeter, a double chamber calorimeter (DCC), suitable for accurate measurement of harmonic losses in an induction motor is also described. A picture of the constructed DCC housing the test motor and the reference heater is shown in Figure 1.2.

Heat transfer process within the calorimeter and the dynamic operation of the DCC is investigated in this chapter. The arrangement for loaded machine tests along with the design of a stuffing box is also presented. The computer data acquisition (DA) system including voltage, current and temperature measurement circuits are described in Chapter 3 as part of the instrumentation and measurement system. Calibration of the measurement system in conjunction with the DA system is also presented in this chapter.
Figure 1.1: Double chamber calorimeter developed for accurate measurement of induction motor losses

In Chapter 4, results of some basic tests performed to measure the conducted heat leakage through the calorimeter walls, edges and corners are given. A simple thermal model is developed to estimate the calorimeter conducted heat leakage through the insulation material. Calibration of the DCC is performed using two identical heaters in separate chambers of the DCC. Test results are employed to derive limits for different aspects of the DCC, including the air flow rate, temperature rise and heat loss measurement. Finally, a theoretical analysis of the accuracy of the loss measurement using the DCC is presented and compared with
the experimental results. It is demonstrated that the DCC is a reliable setup for accurate measurement of the total motor losses with a resolution of about 10 W.

Chapter 5 deals with the specifications of the test motor and initial experiments for calculation of machine parameters. The accuracy of the motor loss measurement using the DCC is verified using a standard high accuracy AC power meter. The suitability of the HG to supply the motor with the predetermined voltage distortion is highlighted. Experimental setup and methodology for conducting harmonic tests using the HG and the DCC are described in this chapter. Calculation of total machine losses under different supply and load conditions is described with the emphasis on the separation of harmonic losses from the total machine losses. Finally, experimental results corresponding to the motor harmonic tests conducted under no-load, half load and full load conditions are presented.

Analysis of the experimental results is given in Chapter 6. The variation of motor additional losses as a function of voltage distortion (both magnitude and harmonic order) are investigated under different loading conditions. Experimental results are also utilised to examine the frequency variation of the machine parameters. Wherever applicable, comparisons are made between the theoretical models and the calculated experimental data. Accordingly, harmonic limits, being a function of voltage distortion, harmonic order and motor parameters, are specified for induction motors. The maximum overheating that the machine can tolerate is determined by defining a derating factor which can be applied to a wide range of induction motors.

Finally, Chapter 7 presents a summary of the conclusions made from this research work. Also some suggestions and comments are given for further work in this area.
Chapter 2

Background: Induction Motor Losses

2.1 Introduction

The issue of losses in induction motors have been of concern for many years. Reduction of motor losses is considered to be one of the major issues in the design of induction motors. This is not only for the purpose of energy saving but also to keep the motor heating under specified limits to gain the maximum possible lifetime. Improved motor design can result in reduced losses and hence higher efficiency which yields significant energy savings especially in large machines.

The presence of time harmonics in the supply of induction motors causes additional losses as well as hot spot temperatures in the machine. Additional losses increase the motor heating and results in reduction of motor efficiency. Hot spot temperatures cause insulation degradation and loss of motor lifetime.

In this chapter a brief description of space and time harmonics along with the mmf produced due to the time harmonics is given. Fundamental and harmonic losses in induction motors are reviewed and the corresponding equivalent circuits are presented. The variation of motor parameters such as stator and rotor resistance and leakage reactance with harmonic frequency and the significance of deep bar effect in determining rotor resistance and leakage reactance are investigated.

Different methods used for the calculation of fundamental and harmonic losses in induction motors are discussed. Some of the loss models available for estimation of motor harmonic losses are also presented. A brief review of the standard methods in determining induction motor losses is given with an emphasis being placed on the calorimetric measurement of electric machine losses.
2.2 Space and Time Harmonics

There are two types of harmonics associated with induction motors, namely, space and time harmonics. Space harmonics are produced by the discrete nature of the winding regardless of the input voltage waveform. Although, the presence of space harmonics are inevitable, unwanted effects such as the associated extra losses can be reduced by improvement in the motor design [Raw51] [Cha63] [Cha66] [Cha67] [Cha68] [Cha69] [Cha70] [Bin75]. For instance, an appropriate pitched windings has been used to reduce the effect of the low order fundamental space harmonics (5th and 7th) on motor performance [Cha63]. The relation between the design variables and losses due to the space harmonics have been investigated in motors having skewed or unskewed rotor slots [Cha70].

Time harmonics are those which exist in the supply voltage of both mains-connected and inverter-fed induction motors. Mains-connected motors experience harmonics caused by nearby distorting loads. Rectifiers, DC motor drives, adjustable frequency AC drives, solid state static voltage controllers, uninterruptable power supplies (UPS), arc furnaces, static var compensators, cycloconverters, HVDC systems, static motor starters and even household appliances such as microwave ovens, TVs and VCRs are the most common sources of harmonics in a power system [Psh83] [Ort85] [Han89] [Lu93] [Cza94].

A study by Fuchs et al [Fuc87] showed that the power system voltages in a particular distribution centre contains a dominant 5th harmonic voltage (about 2% of the fundamental voltage on average) during 24 hours. It has also been reported that the harmonic content is significantly influenced by different transformer connections used at the distribution systems. Another example demonstrated a maximum Total Harmonic Distortion (THD) of 2.7% in the supply voltage of an apartment building [Ema93].

The presence of time harmonics in the supply of power system loads, including induction motors, can cause many unwanted effects. Insulation stress due to the
voltage distortion, thermal stress due to the flow of distorted currents and disruption (abnormal operation or failure caused by harmonic voltages or currents) are some examples [Ort85].

Recently, with the development of power semiconductor switching devices, application of variable-voltage variable-frequency motor drive systems has significantly increased. Inverter-fed induction motors can be found in many applications such as chemical, steel, wood and paper industries as well as in power plants, traction systems and electric vehicles [Pea85] [Hyu90] [Bog94]. However, depending on their design, the output voltage of the static converters contains different order harmonics with different magnitudes. This has led to situations where induction motors experience a greater harmonic distortion and are forced to operate under conditions different from their original design.

Time harmonics present in the supply of induction motors produce rotating mmfs fields in the machine's air gap as described in the next section.

2.3 Rotating MMFs due to the Time Harmonics

When the stator windings of a polyphase induction motor are excited by balanced alternating currents, a magnetic field will be produced in the air gap which rotates with the synchronous speed, \( N_s \), given by:

\[
N_s = \frac{120f}{p}
\]  \hspace{1cm} (2.1)

where \( f \) is the supply frequency in Hz and \( p \) stands for the number of poles in the machine. Magnetic fields produced by each of the three phase balanced supply will have a phase shift of 120 degrees with respect to each other.

Similarly, \( n \)th order time harmonic components of the supply voltage produce magnetic fields which have the same number of poles as the machine but rotate at \( n \) times the synchronous speed. In a three phase power system, the supply generally does not contain any triplen order harmonics. In most practical cases the supply
waveforms are symmetrical and hence even order harmonics do not exist. Therefore, the only significant harmonic components which exist in the mains and/or inverter output voltages are the non-triplen odd harmonics:

\[ n = 6k \pm 1 \quad k = 1, 2, 3, \ldots \quad (2.2) \]

where \( n \) is the harmonic order and \( k \) is any integer giving \( n = 5, 7, 11, 13, \text{etc.} \). The fundamental component of the magnetic field produced by the \( n \)th time harmonic current has a rotational speed:

\[ N_{sn} = \pm \frac{120nf}{p} = \pm nN_s \quad (2.3) \]

where + and - signs indicate that some harmonics produce rotating mmfs in the same direction as the motion of the rotor (positive sequence components) while others produce rotating mmfs in a opposite direction of the rotor motion (negative sequence components) [Jai64] [Kli68] [Rap77] [Fuc87].

The rotor is always travelling backward with respect to the fundamental stator magnetic field having a slip \( s \) which is given as:

\[ s = \frac{N_s - N}{N_s} \quad (2.4) \]

where \( N \) is the rotor speed. The slip corresponding to the \( n \)th harmonic magnetic field, \( s_n \), is:

\[ s_n = \frac{\pm nN_s - N}{\pm nN_s} = \frac{\pm n - 1 + s}{\pm n} \quad (2.5) \]

Under normal load conditions, the rotor slip due to the fundamental frequency is near zero (\( 0 < s < 0.04 \) for low slip machines), and hence the slip for any harmonic frequency can be given as:
\[ s_n \approx \frac{n \pm 1}{n} \]  

(2.6)

which is approximately equal to unity for any reasonable value of n giving the greatest error of 20\% in the presence of 5th harmonic. Also this assumption implies that any time harmonic present in the supply of an induction motor induces a rotor current which has a frequency almost the same as that of the supply. In other words, the rotor is seen as being at standstill relative to the motion of the rotating harmonic magnetic field.

### 2.4 Fundamental Equivalent Circuit

In general, the induction motor equivalent circuit is similar to the usual transformer circuit since the induction motor is essentially a transformer with a rotating secondary. It is a single phase AC circuit which can be used to study the performance of the induction motor at fundamental frequency and under steady state conditions [Cum81] [Del84] [Fuc84] [Fuc86] [Cum86] [San93]. Figure 2.1 illustrates a per phase conventional (exact) equivalent circuit for a three phase induction motor where

- \( V_1 \) = input phase voltage
- \( R_1 \) = stator phase winding resistance
- \( X_1 \) = stator phase leakage reactance
- \( R_c \) = core loss resistance
- \( X_m \) = magnetising reactance
- \( R_2 \) = rotor phase resistance referred to the stator
- \( X_2 \) = rotor phase leakage reactance referred to the stator
- \( s \) = fundamental slip

The use of equivalent circuit is assumed to be the best approach for determining losses and efficiency of induction motors as recommended by most standards [IEEE91] [IEC72].
An approximate equivalent circuit can be achieved by moving the magnetising branch to the machine's terminal as shown in Figure 2.2. From an analytical point of view, this type of circuit is much simpler than the exact equivalent circuit without introducing a great inaccuracy.

The use of equivalent circuits of whatever form to determine the performance of induction motors is simple and accurate in many situations, but there are always many approximations and it is difficult to predict losses accurately in machines with different design parameters. According to the literature, equivalent circuit parameters can be calculated from the no-load and locked rotor impedance test data [IEEE91] [IEC72] [Cum81]. These parameters are non-linear and vary with current, frequency and temperature. Therefore, adequate knowledge of the motor design parameters are required to obtain accurate values for equivalent circuit parameters.

Although most standards use only one form of equivalent circuit, many modifications have been suggested. One such a circuit has been presented in [Del84] where two rotor loops suitable for both single and double cage induction motors have been incorporated. A detailed equivalent circuit appropriate for considering the core losses due to the main and leakage fluxes has been given in
where modifications are required to account for the effect of harmonics and skin effect.

2.5 Harmonic Behaviour of Induction Motors

2.5.1 Equivalent Circuit

The effect of harmonics on induction motor performance can be evaluated by developing a series of independent equivalent circuits supplied by each individual harmonic voltage source [Cha68] [Kli68] [Cum86] [Maa90]. Assuming linearity, superposition can be applied to add effects of individual harmonics and hence determine the machine performance under a harmonically distorted situation. Using this approach, total harmonic losses in the machine can be calculated by summation of separate losses corresponding to each harmonic frequency.

Although, in most cases it is assumed that individual harmonics affect the machine performance independently, there are still some arguments in which the interaction between different order harmonics becomes appreciable. One such an example shows that the pair of harmonics such as 5th and 7th might produce negative active power in a particular inverter-fed induction motor having iron bridges in its rotor [Nee93].

A harmonic equivalent circuit can be approximated by the circuit shown in Figure 2.3 where the magnetising branch is neglected. This assumption is valid since the harmonic slip, $s_n$, is close to unity and values of $R_c$ and $X_m$ are much larger than $R_1$, $R_2$ and $X$ [Cum86] [Fuc86]. Typical values of $R_c = 10$ to 40 pu and $X_m = 1$ to 3 pu can be compared with $R_2 = 0.03$ to 0.1 pu and $X_2 = 0.05$ to 0.15 pu [IEEE87]. The given harmonic equivalent circuit is then similar to the locked rotor equivalent circuit [Kli68] [Cum86]. However, the magnetising reactance (representing the air gap flux) can be assumed to be saturated at harmonic frequencies and hence $X_{mn} < n X_m$ [Wil82].
The harmonic equivalent circuit shown in Figure 2.3 does not include any parameters to represent the extra harmonic losses in the iron and/or stray load losses. This was based on the assumption that the additional iron and stray losses can be neglected [Kli68]. This argument was valid in the past where the inverter-fed induction motors were not commonly used and thus the iron losses due to the high frequencies (e.g., switching frequencies as well as the voltage time harmonics) were not so significant. In extreme cases, and in inverter-fed induction motors, these losses could be as high as the fundamental core losses [Ric85] where the motor experiences distorted waveforms containing high switching frequencies.

A more accurate harmonic equivalent circuit is presented in [Cum86] where a stray load loss resistor $R_{\text{ln}}$ is added as shown in Figure 2.4. This resistor represents extra harmonic losses in the iron and time harmonic space fundamental stray losses.

Although no equivalent circuit has been suggested in [Buc84], the given loss model consists of parameters which results in a harmonic equivalent circuit compatible with that shown in Figure 2.4. The variation of harmonic equivalent circuit parameters is given in the following sub-sections.
In some cases the equivalent circuit has been modified to represent the forward and backward field components as well as the harmonic behaviour of the motor [FucI84] [FucII84] [Fuc86]. The copper losses due to time harmonics are calculated using an equivalent circuit which considers the skin effect in rotor bars [Ven82]. A relatively complicated harmonic equivalent circuit has been suggested by Honsinger [Hon80]. The additional losses associated with stator and rotor leakage fluxes including the stray load losses due to presence of time harmonics have been accounted for by introducing additional components. The values of these components have been derived for different order time harmonics and used for calculation of total harmonic losses.

### 2.5.2 Rotor deep bar effect

When AC currents pass through rotor bars, the corresponding leakage flux will be distributed across the cross section of the bar in a non-uniform manner. The bottom sections of the bar are linked by more leakage flux when compared to the top sections close to the air gap. This, in turn, increases the effective resistance of the bar and reduces its effective leakage inductance. Therefore, the current distribution will be non-uniform and accumulated towards the top of the bar. This phenomenon is called deep bar effect [Alg51] which describes the current displacement in the rotor bars.

In addition to the frequency, the rotor resistance and leakage inductance depends on the bar height, shape, material and open or closed slot character [Buc84] [Cum86]. Different rotor bar shapes including the double cage rotor arrangements have been utilised by the designers to take advantage of deep bar effect. As a result various torque-speed characteristics can be obtained to meet the general requirements for different types of induction motors [Alg51] [Fit90]. Some of the typical rotor bar shapes are shown in Figure 2.5.
2.5.3 Stator winding resistance variation ($R_{1n}$)

In small induction motors, the stator winding resistance at fundamental frequency, $R_1$, is often considered to be equal to its DC value but subject to variations only due to the temperature. However, in large machines having multilayer conductors lying in deep stator slots [Cha68] or when the primary conductor depth (diameter) is appreciable [Buc84], $R_1$ is subject to variation due to skin effect. A greater increase in stator resistance can be expected when harmonics are present in the input voltage supplied to the motor [Buc84] [ Ort85]. According to the work presented by DeBuck et al [Buc84], total stator winding resistance corresponding to the nth harmonic can be expressed as:

$$R_{1n} = R_{1dc}(1 + C_1 h^4 n^2)$$ (2.7)

where $R_{1dc}$ is the stator DC winding resistance, the constant $C_1 = 1.58 \times 10^{-5}$, $n$ is the harmonic order of the stator current frequency and $h$ is the stator slot depth in cm. According to [Buc84] the stator winding resistance can be as 2.5 times larger than its DC value when experiencing harmonic frequencies. Also some influencing parameters such as ratio of coil end resistance to slot resistance, number of conductor layers, conductor cross sectional shape, material and temperature have been indicated which affect the stator resistance.

Stator and rotor slot depths can be different by a maximum factor of 25% but for simplicity these two are considered to be equal [Buc84] and can be approximated using an empirical equation as:
where \( P \) is the motor power rating in kW.

According to Cummings [Cum86], with the presence of the stray load loss resistor \( R_{lln} \), \( R_{ln} \) can be considered identical with its DC value, \( R_{1dc} \), which only varies with temperature.

### 2.5.4 Rotor resistance variation (\( R_{2n} \))

For a given machine where the rotor bar height, shape, material and open or closed slot cannot be changed the only dominant influencing factor on rotor parameters (resistance and leakage reactance) is the frequency of the rotor current. The variation of rotor bar resistance as a function harmonic frequency has been investigated for many years [Wil82] [Buc84] [Cum86] [Lan89] [Muk89] [Lev90] [Lip92] [Maa90] [Whi94] [Zha94]. There have been some investigations on the influence of deep bar effect on rotor end ring resistance [Wil86] [Wil87].

Frequency variation of rotor resistance with assumption that the end ring and bar DC resistances are equal is given as:

\[
R_{2n} = R_{2dc}(1 + C_2 h n^{0.5})
\]

where \( R_{2dc} \) is the total rotor DC resistance, \( h \) is the useful conductor (bar) height already defined by Equation (2.8) and \( n \) is the harmonic order of the rotor current [Buc84]. The constant \( C_2 \) is a function of motor power rating and is different for various types of bar shapes depending on the cage material and temperature. For motors with \( P < 10 \text{ kW} \) a value in the range 0.18 to 0.35 has been suggested for \( C_2 \) and for motors with \( P > 30 \text{ kW} \) \( C_2 = 1.06 \).

The variation of rotor resistance with the square root of the harmonic order has been reported by other researchers [Maa90] [Ort85]. However, an examination on a graph given in [Muk89] showed that the rotor bar resistance changes with square of the harmonic order due to skin effect. For instance, it could be as 1.07 and 25 times
larger than its nominal value at a harmonic order \( n = 5 \) and \( n = 100 \) respectively. As compared with Equation (2.9), this figure underestimates the variation of rotor resistance at low order harmonics and significantly overestimates at higher order harmonics.

Three different graphs corresponding to different rotor bars have been given to relate the variation of \( R_{2n} \) with harmonic order, \( n \) [Cum86]. It has been shown that the rotor resistance could be about five times larger than its DC value when experiencing 5th harmonic. Although, no expression has been stated for \( R_{2n} \) as a function of harmonic order, total machine resistance, \( R_n \), has been expressed as:

\[
R_n = R \ n^{0.6} \tag{2.10}
\]

where \( R \) is the total pu machine resistance (neglecting stray loss resistance, \( R_{ll} \)) at fundamental frequency. The exponent 0.6 is an empirical constant which mostly represents the frequency variation of the rotor resistance.

### 2.5.5 Stator and rotor leakage reactance

The stator leakage inductance is usually considered to be unchanged with harmonic frequency [Cum86] and hence the stator leakage reactance, \( X_{1n} \), is proportional to the harmonic order as:

\[
X_{1n} = n \ X_1 \tag{2.11}
\]

where \( X_1 \) represents the stator leakage reactance at fundamental frequency.

The rotor leakage inductance, however, decreases as the rotor current frequency increases due to deep bar effect. Therefore, rotor leakage reactance, \( X_{2n} \), increases with harmonic order but not linearly:

\[
X_{2n} < n \ X_2 \tag{2.12}
\]

where \( X_2 \) is the rotor leakage reactance at fundamental frequency (without saturation).
The total effective leakage reactance can be calculated as sum of the individual components as \( X_n = nX_1 + X_{2n} \). Although, it is common to give the stator and rotor leakage reactances individually, in practice they are often presented in a combined form. Such a variation is given in \([\text{Cum86}]\) as:

\[
X_n = Xn^{0.8}
\]  

(2.13)

where \( X \) stands for the nominal total leakage reactance and can be calculated as sum of stator and rotor leakage reactances at fundamental frequency (ie \( X = X_1 + X_2 \)).

Equation (2.13) mostly represents the frequency variation of the rotor leakage inductance under deep bar effect. An empirical equation is given in \([\text{Buc84}]\) which describes the total leakage inductance as a function of harmonic order, \( L_n(\text{pu}) \), as compared with the nominal leakage inductance \( L(\text{pu}) \):

\[
\frac{L_n(\text{pu})}{L(\text{pu})} \approx 1.07n^{-0.16} \quad n \geq 2
\]  

(2.14)

These variations are also illustrated in Figure 2.6 where a reduction of 35% in nominal leakage inductance can be calculated at \( n = 20 \). Equation (2.14) has been derived by performing experimental tests on motors with different power ratings (2.2, 3, 10 and 160 kW) while the rotor was locked \([\text{Buc84}]\). The tests have been performed by supplying DC power to the two of the phases in order to saturate the motor. The third phase has been supplied by a high frequency power source (150 V, 0-30 A) producing square waveforms (rather than sinusoidal) with a frequency range of 30 Hz to 12 kHz. However, using this procedure, it has been reported that the motor has experienced time harmonics as they were superimposed in the fundamental field.

It was also claimed in \([\text{Buc84}]\) that square waveforms did not substantially influence the measurements and calculations and so no corrections were made. However, no comments were given to clarify how the effects of different order harmonics of the square waveform were segregated. Based upon the measurements, four different
curves were obtained to represent the leakage inductance at different frequencies corresponding to test motors. Equation (2.14) represents an average estimation of leakage inductances measured for different motors and is reported to be independent of motor power rating. An expression similar to Equation (2.14) is given in [Mal92] where the exponent changes from -0.1 to -0.27 with increase of harmonic frequency.

![Figure 2.6: Variation of nominal leakage inductance with harmonic order](image)

Using Equation (2.14), the total leakage reactance, $X_n$, can be expresses as:

$$X_n \approx 1.07 X_0 n^{0.84}$$  \hspace{1cm} n \geq 2 \hspace{1cm} (2.15)$$

where $X$ is the total leakage reactance at fundamental frequency. As stated, both Equations (2.14) and (2.15) are valid for $n \geq 2$ and not for fundamental frequency (ie $n = 1$).

A typical variation of $R_n = R_1 + R_{2n}$ (the effective resistance neglecting $R_{2n}$) and $X_n = nX_1 + X_{2n}$ (the effective leakage reactance) is given in [Wil82] and [Cum86] as shown in Figure 2.7. It can be seen that the effective resistance $R_n$ is much less than the effective reactance $X_n$ and hence $Z_n = R_n + jX_n \approx jX_n$. Therefore, the nth harmonic current $I_n$ can be approximated as:

$$I_n = \frac{V_n}{Z_n} \approx \frac{V_n}{X_n}$$ \hspace{1cm} (2.16)
where \( V_n \) is the pu harmonic component of the motor input voltage.

With typical values of \( R = 0.05 \) pu and \( X = 0.2 \) pu, this approximation is subject to a maximum error of 2% in the presence of the 5th harmonic. With higher order harmonics, the error will be significantly reduced and can be neglected since \( X_n \) is the dominant part of the total machine impedance. However, it should be noted that \( R_n \) has to be taken into account when harmonic losses are to be calculated. In other words, \( X_n \) determines the harmonic currents while \( R_n \) specifies the actual harmonic losses in the machine.

![Diagram](image)

Figure 2.7: Effective total machine resistance and leakage reactance as a function of harmonic order [Wil82] [Cum86]

### 2.6 Losses in Polyphase Induction Motors

The issue of losses in polyphase induction motors has been a major subject of investigation for many years [Cha68] [Kli68] [Alg70] [Cum81] [Ven82] [Ric85]. This has attracted even more attention with the increasing application of inverter-fed induction motors in industries and electric utilities. The motor efficiency become very significant especially in large machines where the losses could be as high as 5% of the motor power rating. Improvements of 0.1% in efficiency of an induction motor could contribute to an average energy saving of more than $1/hp output power [Gra96].

In general, machine losses are defined as the difference between the input and the output power and are classified as:
- Stator winding losses ($W_1$)
- Rotor winding losses ($W_2$)
- Core losses ($W_c$)
- Stray load losses ($W_{ii}$)
- Friction and windage losses ($W_{fw}$).

A typical characteristic of different losses versus load for induction motors operating under sinusoidal supply conditions is shown in Figure 2.8. These losses are often referred to as fundamental losses. It can be seen that the core losses and windage and friction losses are almost independent of the motor loading. Therefore, their no-load values are often considered in determining motor losses and/or motor efficiency which takes place at full load. Stator, rotor and stray losses, however, increase with machine load.

![Figure 2.8: Variation of induction motor losses versus load under ideal supply conditions [Cum81]](image-url)
A typical example of different component of losses associated with a cage induction motor operating under rated torque and speed and supplied by sinusoidal voltage is given as the percentage of total losses as [IEC92]:

\[
W_1 = 41\%
\]
\[
W_2 = 26\%
\]
\[
W_c = 25\%
\]
\[
W_{ll} = 2\%
\]
\[
W_{fw} = 6\%.
\]

There have been many studies regarding the additional losses associated with the presence of time harmonics in the supply of induction motors [Jai64] [Cha68] [Kli68] [Mcl69] [Rap77] [Buc79] [Ven82] [Buc83] [Buc84] [Ric85] [Cec86] [Cum86] [Mal92] [Nee93] [Cza94] [Lam94] [Eld95]. Most of these references investigate the significance of additional harmonic losses in inverter-fed induction motors in order to control the electric drives efficiently.

As an example, additional losses in a 10 hp motor fed by a six step inverter could be as high as 80% of the fundamental losses on full load [Hon80]. Another example demonstrated that harmonic losses in a 7.5 kW induction motor supplied by PWM inverter could be about 60% of the rated motor losses [Bog93]. Additional core losses of 54% and 98% have been reported for a standard and energy efficient motor due to distorted waveforms (e.g. a VSI) [Ric85].

A theoretical study shows that the harmonic losses in a typical 50 hp induction motor supplied by an inverter can be between 5% to 11% of the rated motor power [Lip92]. Additional losses in a 15 kW induction motor supplied by quasi-square voltage with 30% THD has been reported to be about 23% of the motor losses under rated conditions [Kli68]. There has been no comment on the level of distortion due to the individual harmonics.
Harmonic losses also can be classified as in the case of fundamental losses and can be separated into various components as rotor and stator copper losses as well as additional core and stray losses. Harmonic losses are reported to be independent of motor load [Kli68] [Cha68] [Wil82] [Nee93], however, no experimental justification have been reported.

### 2.6.1 Stator winding losses

Stator winding losses, $W_1$, are the copper losses in the stator winding which are assumed to be proportional to the square of the stator rms phase current, $I_1$:

$$W_1 = 3R_1I_1^2$$  \hspace{1cm} (2.17)

where $R_1$ is the stator winding resistance per phase. Under nominal conditions $R_1$ is assumed to be equal to its DC value which only varies with temperature [Cum81]. The stator winding resistance can be simply calculated from the number of turns and their average length, the wire size and material. It can also be measured using a DC voltage source and/or a multimeter. Experimental results on the 7.5 kW high efficiency test motor confirmed that the stator resistance could increase by up to 40% due to the motor temperature increase as from no-load to full load.

The stator $I^2R$ losses considerably increase due to the flow of harmonic currents in the stator windings [Kli68]. As an example, an extra stator copper loss of 9.6% and 2.5% was measured respectively for a standard and an energy efficient motor supplied by distorted waveforms (e.g. a VSI or CSI) as reported in [Ric85]. It has been stated that the latter has less harmonic losses because of its higher leakage reactance limiting harmonic currents. Another example indicates that with 10% of 5th harmonic voltage, the stator copper losses can be increased by 4% when compared with the rated stator winding losses [Fuc87].

Neglecting the skin effect, the stator winding harmonic losses can be estimated as:

$$W_{i_n} = 3R_{i_n}I_n^2$$  \hspace{1cm} (2.18)
where $I_n$ is the rms value of the nth harmonic current which can be considered as the same in the stator and rotor circuits as shown in equivalent circuit of Figure 2.4 (ie $I_{1n} = I_{2n} = I_n$). In most cases $R_{1n}$ is replaced by the stator DC winding resistance which only varies with temperature. However, as presented in [Buc84], $R_{1n}$ can be calculated using Equation (2.7) where the skin effect in the stator conductors are taken into account.

### 2.6.2 Rotor winding losses

Rotor $I^2R$ or slip losses, $W_2$, are losses in the rotor bars and end rings due to the flow of current induced by the fundamental flux. In general, the rotor $I^2R$ losses under fundamental frequency are proportional to the square of the rotor current and can be expressed as:

$$W_2 = 3R_2I_2^2$$  \hspace{1cm} (2.19)

where $R_2$ is the total rotor cage resistance (including bars and end rings) per phase and $I_2$ is the rotor phase current. Under nominal conditions, $R_2$ is assumed to be equal to its DC (low slip) value which only varies with temperature [Cum81]. In cage induction motors the resistance of the rotor cannot be measured directly and is often calculated using well defined standard tests. Alternatively, when all the equivalent circuit parameters are not available, the rotor losses can be approximated as:

$$W_2 = sW_{ag}$$  \hspace{1cm} (2.20)

where $s$ is the fundamental slip and $W_{ag}$ is the air-gap power [Cum81] [San93]. The IEC and Australian Standards allow an approximation for estimation of $W_{ag}$ as:

$$W_{ag} = W_m - W_1 - W_{nl}$$  \hspace{1cm} (2.21)
where $W_{in}$ and $W_1$ are the input power and stator losses under any load conditions respectively and $W_{nl}$ is the no load input power [San93]. This approach is a simple alternative for calculation of rotor slip losses and can be performed easily.

In case of non-sinusoidal excitation the rotor harmonic losses are considered as the dominant component of losses caused by harmonic currents [Kli68] [Hon80] [IEC92]. This is mainly because of the increase of rotor resistance due to deep bar effect at harmonic frequencies.

As an example, the presence of 10% of 5th harmonic voltage can cause an increase of 10% in the rotor losses as compared with the rated rotor losses [Fuc87]. Another example has quoted a 5.7% increase in rotor $I^2R$ losses in an induction motor supplied by distorted waveform as reported in [Ric85]. A recent study has shown that the extra losses due to deep bar effect (quoted as skin effect) could be about three times larger than that if this effect is neglected [Lip92]. It has been demonstrated that rotor losses can spatially increase the rotor heating during start up [Cha94] and can causing a high thermal stress in the rotor especially in the junctions of the rotor bars and end rings.

In general, the harmonic copper losses in the rotor due to the flow of harmonic current, $I_n$, can be calculated as:

$$W_{2n} = 3R_{2n}I_n^2$$  \hspace{1cm} (2.22)

where $R_{2n}$ is the rotor resistance corresponding to the nth harmonic order. It should be noted that the harmonic order in the rotor is different from that in the stator but, as described in Section 2.3 the two can be assumed identical without a great error.

### 2.6.3 Core losses

Core losses, $W_c$, are iron losses in the stator and rotor due to space fundamental and harmonic fluxes [Cha68] [Ven82]. These losses consist of hysteresis and eddy current losses which are proportional to the frequency and square of the frequency
respectively [Cum81] [Kir84] [Gos87]. Time varying rotating flux causes the magnetic material to experience a cyclic variation resulting in hysteresis losses. They can be defined as the amount of energy lost in relation to the corresponding hysteresis loop and hence depend on the area of the loop and the frequency. Due to the complexity of the hysteresis loop, an empirical relation rather than the analytical solution, has been applied. Hysteresis losses can be estimated as a function of maximum flux density, $B_{\text{max}}$, and frequency as:

$$W_h = K_h f B_{\text{max}}^\alpha$$  \hspace{1cm} (2.23)

where $K_h$ is an empirical coefficient being a function of volume or weight of the magnetic material and the exponent $\alpha$ is in the range 1.5 to 2.5 [Sel92]. Under normal operation of the motor the slip frequency (rotor current frequency) is very small and hence hysteresis losses in the rotor can be neglected. The frequency in the stator is the same as that of the supply and hence $W_h$ in the stator core is appreciable.

The second part of the core losses is known as eddy current losses [Sel92]. Eddy currents are the circulating currents produced due to the induced emfs in the iron when is subject to a alternating flux. The direction of the eddy currents in the iron is so as to oppose the change in the flux and force the magnetic field to the outer surfaces due to the magnetic skin effect. This process produces heat in the iron and hence eddy current losses. A significant reduction in flow of eddy currents has been achieved by introducing laminated cores in both stator and rotor. The eddy current losses can be approximated by:

$$W_e = K_e f^2 B_{\text{max}}^2$$  \hspace{1cm} (2.24)

where $K_e$ is a constant and its value depends on the thickness of the laminations and is inversely proportional to the resistivity of the material [Sle92]. Equation (2.24) while gives a good estimate of the eddy current losses, is not accurate. The reason is that the flux density in the iron is assumed uniform and unaffected by the eddy
currents which is not the case in practice. Moreover, the permeability of the iron is assumed constant but in most practical cases the operation is performed under saturation and in the non-linear area of the B-H curve where the iron permeability is different.

In practice, however, core losses are not segregated into hysteresis and eddy current losses and are given under a single term of core losses. These losses are a function of secondary voltage (induced emf) and are assumed to be constant at their no-load value [Cum81]. Under no-load conditions the core losses are usually calculated as no-load input power less the stator copper losses and friction and windage losses as:

\[ W_c = W_n - W_1 - W_{fr} \]  \hspace{1cm} (2.25)

Core losses are usually modelled by a resistor \( R_c \) in parallel with the magnetising reactance \( X_m \) as shown in equivalent circuit of Figure 2.1. The value of this resistor can be experimentally calculated for different machines using Equation (2.25) and the measured input voltage. Since the flux density \( B \) is proportional to the ratio \( V/f \), the core losses are often said to be proportional to the voltage squared. Therefore, unlike the stator and rotor copper losses, core losses are influenced by the voltage waveforms rather than the current waveforms. Under normal operation, core losses in the rotor are often neglected due to low slip frequency. The stator core losses take place mostly in stator back iron (about 70%) and in stator teeth [Bou95].

The presence of time harmonics in the supply voltage of induction motors causes core losses to be increased. The increase is due to the higher peak flux density which affect both stator and rotor iron cores [Wil82].

Using Equations (2.23) and (2.24) and the relation \( V \propto B \cdot f \), the core losses due to the pu time harmonic voltage, \( V_n \), and harmonic order, \( n \), can be approximated as:

\[ W_{cn} \propto n^{\mu-\nu} V_n^\nu \]  \hspace{1cm} (2.26)
where $\mu = 1.5$ to $1.65$ and $\nu = 1.7$ to $2$ [Buc84]. Equation (2.26) neglects the core losses in the rotor due to both time fundamental and time harmonic frequencies. As reported in [Buc84], this is because the air gap flux due to time harmonics acts as a shield on the rotor side.

Apart from the values of $\mu$ and $\nu$, there are some other influencing parameters such as variation of pu leakage reactance from 0.13 to 0.23 and iron losses due to motor skewing and end-leakage reactances in the rotor and stator which affect the accuracy of Equation (2.26). Some of these effects represent the time harmonic space fundamental stray losses [Buc84]. Accordingly, a multiplying factor of has been incorporated to compensate the mentioned effects. With typical values of $\mu = 1.5$ and $\nu = 2$ and using Equation (2.14), harmonic core and space fundamental stray losses can be expressed as a function of harmonic currents rather than the harmonic voltages as:

$$W_{cn} = 0.25 W_c n^{118} I_n^2$$

where $W_c$ is the core losses at fundamental frequency [Buc84]. This type of equation implies that $W_{cn}$ can be represented as a frequency dependent series resistive component in the harmonic equivalent circuit.

Based on a non-linear magnetic field analysis, magnetising currents and iron losses for two saturated induction motors (800 W and 75 kW) have been calculated [FucI84] [FucII84]. Harmonic core losses as compared with fundamental core losses have been calculated as:

$$\frac{W_{cn}}{W_c} \propto \left(\frac{E_n}{E}\right)^2 \frac{R_c}{R_{cn}}$$

where $E$ and $E_n$ are the fundamental and harmonic components of the induced voltage respectively, $R_c$ and $R_{cn}$ are the core loss resistances at fundamental and harmonic frequencies respectively. It has also been shown that $R_{cn}$ is a parallel component of harmonic equivalent circuit and can be approximated as:
\[ R_{cn} = R_c n^{0.6} \quad n \leq 11 \] (2.29)

which gives an accurate measure for harmonic core losses but not for the harmonic component of the magnetising current [FucI84].

The effects of square wave and PWM inverter supply on core losses of an induction motor has been investigated and reported in [Bog96]. The corresponding results have been utilised to examine the inverter parameters and their effect on total machine losses. A procedure has been developed to assess the iron losses in a rotor-flux-oriented induction motor under frequencies from zero to 100 Hz [Lev96]. The corresponding iron loss resistance has been calculated for the given frequency range.

2.6.4 Stray load losses

In general, stray load losses are defined as additional fundamental and high frequency losses in the iron, losses due to the circulating currents in the stator winding, and harmonic losses in the rotor bars under load conditions [Cum81]. These losses originate from the saturation effect in magnetic materials, space harmonics, leakage flux and structural imperfections of induction motors [JimI85] [JimII85]. With purely sinusoidal supply stray load losses are about 0.5 to 1% of the rated motor power, 5 to 15% of the rated motor losses or 10 to 25% of the rated \( I^2R \) losses (neglecting skin effect) [Buc84] [San93]. Although, stray load losses have been a subject of investigations for a long time, there are still confusions in their definitions, effects and methods in which they can be accurately estimated and/or measured.

In some cases, stray no-load losses are also mentioned but they are often considered as part of the fundamental core losses since they are very difficult to be separated [Cum81]. There have been no reliable method for computing stray no-load losses as an individual part.
According to the literature, the stray load losses can be divided into the following components [Alg59] [Cha63] [Kli68] [Cum81] [JimI85] [JimII85] [Eld95]:

- Eddy current losses in stator conductors due to slot leakage flux,
- Losses in the motor end structure (copper, steel and other metallic parts) due to slot and end-region leakage flux,
- Rotor and stator surface losses due to the zigzag leakage flux,
- Eddy current losses in stator and rotor end laminations due to end leakage flux,
- Surface losses in the stator and rotor due to the zigzag leakage flux,
- Losses due to pulsation flux in the rotor teeth and losses due to slot permeance and slot mmf harmonics,
- Induced losses in the rotor due to mmf harmonics produced by stator load current,
- Induced losses in the stator due to rotor mmf harmonics,
- Extra core losses due to the skew leakage flux (in motors with skewed slots).

Approximate methods have been developed to estimate different components of the stray load losses [Alg59]. Definitions of stray load losses as well as different methods for measurement and calculation of these losses have also been reported in [JimI85] [JimII85]. The effect of phase belt, slot mmf, and slot permeance harmonics have been included in the equivalent circuit for calculation of main and stray copper losses [Jai64]. Stray iron losses due to mmf and permeance harmonics, end leakage and skew leakage have also been calculated.

Most of the standards recognise the difficulty in measuring stray losses. Some standards including IEEE Standard 112 test methods [IEEE91] assume that stray load losses are proportional to the square of load torque or rotor current squared. Being a function of load current, measurement of stray load losses can be improved by forcing them to fit the equation $K_1 (I_2)^2$ or $K_2 T^2$ where $I_2$ stands for the rotor current and $T$ is the shaft torque [Cum81]. To determine stray load losses, $W_{II}$, two methods have been recommend by [IEEE91] as direct and indirect methods. The
direct method deals with the rotor removed and reverse rotation tests. In the indirect method the stray load losses are obtained as:

\[
W_l = \text{Test losses} - \text{Conventional losses} \quad (2.30)
\]

where:

\[
\text{Test losses} = W_{\text{in}} - W_{\text{out}} \text{ (at the full load)} \quad (2.31)
\]

\[
\text{Conventional losses} = W_1 + W_2 + W_{\text{fw}} + W_c \quad (2.32)
\]

This method suffers from the lack of confidence in the results due to the inaccuracies in the measurements especially mechanical output power, \(W_{\text{out}}\). IEEE Standard 112 [IEEE91] suggests that if stray load losses are not measured, depending on the motor power rating, an assumed value between 0.9% to 1.8% of the motor rated power can be applied as rated value of the stray load losses.

IEC [IEC72] and Australian Standards [AS83] assume that stray load losses vary with the square of the stator current, \(I\), and are approximately equal to 0.5% of the input power at the rated load:

\[
W_l = 0.005P_{\text{rated}} \left( \frac{I}{I_{\text{rated}}} \right)^2 \quad (2.33)
\]

where \(P_{\text{rated}}\) and \(I_{\text{rated}}\) are the input power and the line current at rated load [San93]. However, in many practical situations the stray load losses might exceed the 0.5% figure [Eld95].

Stray load losses are highly affected by the frequency as in the case of non-sinusoidal excitation [Cum81] [Hon84] [Ric85] [San93] [Gra96]. In this case these losses are typically larger than the core losses. The rotor stray losses which are negligible under sinusoidal excitation are amplified due to the rotor harmonic frequencies [Hon80]. A number of methods including developing modified equivalent circuits [Hon80] and defining stray load loss components [Cum86] have
been suggested for determining stray load losses due to the time harmonics. Some of these already discussed in conjunction with other components of harmonic losses. As an example, stray load losses due to the time harmonics have been taken into account by introducing a stray load loss resistor, $R_{lln}$:

$$R_{lln} = R_{ll}n^{0.8}$$  \hspace{1cm} (2.34)

where $R_{ll}$ is defined as the base value of the stray load loss resistor [Cum86] but no comment has been given for its calculation. Assuming that the stray load losses are a fraction $K_{sll}$ of the total losses, $R_{ll}$ can be approximated by:

$$R_{ll} = K_{sll} \left( \frac{1 - \eta}{\eta} \right)$$  \hspace{1cm} (2.35)

where $\eta$ is the full load efficiency of the motor [Per96].

In some models such as that presented in [Buc84], part of stray load losses due to time harmonics have been included in harmonic core losses as given by Equation (2.27).

### 2.6.5 Friction and windage losses

Friction and windage losses, $W_{fw}$, are mechanical losses due to the friction of bearings and windage which are typically constant with load [Cum81]. These losses are a function of the cooling fan design, bearing losses, and the aerodynamics of the rotor structure [Ric85]. $W_{fw}$ is not influenced by the voltage waveform [Kli68] and its value is independent of any harmonic effects [Ric85]. As recognised by most standards, this part of the losses can be separated by reducing the input voltage of the unloaded machine and by plotting the input power versus voltage squared [Cum81].

A recent investigation on IEEE Standard 112 test methods [IEEE91] demonstrated that the condition of the grease in the bearing could have a large impact on the friction and windage losses and hence the efficiency test results [Gra96].
2.7. Standard Methods for Determining Losses

There are a number of standard methods for determining losses in induction motors. Three standard methods including IEEE Standard 112, IEC 34-2, and JEC-37 have been discussed and compared by Cummings et al [Cum81]. IEEE Standard 112 efficiency test methods have been discussed under two parts as direct output measurement (Methods A, B and C) and determination of losses without output measurement (Methods E and F). It has been demonstrated that the IEEE, JEC and IEC brake method (Method A in IEEE) are basically similar but differing in temperature corrections.

The IEEE Method A employs temperature corrections which is suitable for low power motors. The conventional losses are measured directly in IEEE Methods B and C where appropriate temperature corrections are required. Method C is very accurate because all readings are electrical and the meter errors are cancelled by reversing the power flow. The IEC pump-back method is similar to Method C except for temperature corrections and reversal power flow method [Cum81].

In IEEE Method E (separation of losses) \( W_{fw} \) and \( W_c \) are determined by no load test and \( W_1 \) and \( W_2 \) are determined from full load test. JEC has a loss separation method which does not include the stray load losses [Cum81].

In IEEE Method F (the exact equivalent circuit) is used where its parameters are determined by no load and impedance tests. The impedance test is conducted under 25% of the rated frequency to eliminate the skin effects. The JEC circle diagram method uses zero stray load loss and a higher frequency impedance test. The IEC specification uses no load, reduced load and reduced voltage in determining the conventional losses [Cum81].

Recently, the IEC Standards, Series 34 on Rotating Electrical Machines and the ANSI C50.41-1982 on Polyphase Induction Motors for Power Stations have been compared [Nil96]. The comparison has been made on temperature rises, overloading, minimum starting torques, external inertia, sound limits, vibration
limits, resistance correction temperature methods, dielectric tests, voltage unbalance, harmonic voltage and other miscellaneous topics. In terms of allowable harmonics, IEC allows a maximum harmonic voltage factor of 3% in the supply of the AC motors while ANSI has no comment on this issue. In general, it has been suggested that both standards are required to comment more on harmonic issues [Nil96].

Standard methods are considered as the most reliable and accurate techniques for determining machine losses and hence efficiency. However, motor losses are often evaluated under nominal rated conditions where the input voltage contains only the fundamental frequency. Measurement of motor electrical input power is possibly the easiest task due to the availability of the standard laboratory equipment. There have been some difficulties and inaccuracies due to the measurement of output shaft power. It has been reported that the precision becomes difficult when the input and output values get closer together [Cum81]. This is the case in machines where an efficiency of more than 0.9 should be calculated.

In harmonically distorted situations, however, a high precision is more difficult to achieve since the laboratory equipment have limited frequency bandwidth and their measurements are subject to inherent inaccuracies. Precise and accurate measurement of losses under distorted conditions demand development of sophisticated equipment for measurement of input and output power which is rather a costly effort. Alternatively, machine losses can be determined using calorimetric method as suggested by the IEC [IEC74] which will be described in the next section.

2.8 Calorimetric Method

The IEC Publication 34-2A [IEC74] has suggested a calorimetric method for determining losses in large electric machines. The method basically allows for the measurement of the dissipated heat through the cooling system and hence estimation of the total machine losses directly. There are two alternatives for performing
calorimetric tests and determining machine losses, namely, direct and calibration methods. In both methods accurate and continuous measurement of coolant properties such as flow rate, temperature, density and specific heat is required. This approach is originally intended for large machines but the principle can be applied to measure losses in small motors.

An alternative calorimetric method has been proposed for the measurement of induction motor losses [Tur91]. A thermally insulated container, a calorimeter, has been developed where machine losses can be estimated using a heater in two subsequent tests. Although, this method provides a relatively simple approach for loss estimation, it still requires a relatively complicated system to control the air properties. It has been reported that the method is accurate and repeatable for loss measurements with a resolution of 9 W. Further discussion on this method along with the principle of the operation of the calorimetric method will be presented in Chapter 3.

A modified version of the calorimetric calibration method suitable for loss measurement in converter-fed induction motors has been reported in IEC 34-2, Amendment 2 [IEC96]. The method requires an experimental setup to take away the generated heat by the test motor. A dissipation resistor whose input power can be easily measured is required in the exhaust path for balance type of measurement. The motor losses will be determined as the dissipated heat in the resistor in relation to the air temperature rise.

Taking the existing calorimetric methods as guides, construction of a new double chamber calorimeter (DCC) is proposed in this thesis. The DCC utilises a simple approach to conveniently perform direct measurement of motor losses without requiring critical control and measurement of the air properties. One of the advantageous of the DCC is that motor losses can be estimated regardless of the motor input voltage and current waveforms. Also, loss measurements can be performed accurately without being affected by the error involved in the measurement of output shaft power. Details of the design and construction of the
DCC, initial testing and calibration along with its application to measure motor losses are fully investigated and given in Chapters 3 and 4.

2.9 Conclusions

Classification of fundamental losses and different methods for estimation and their calculation in induction motors have been described in this chapter. Evaluation of losses in the rotor and stator windings as well as in the iron can be performed in different ways including the conventional method of equivalent circuit which is recognised by most standards. Additional losses due to the time harmonics present in the supply voltage of induction motors have been investigated. They can be estimated by developing individual harmonic equivalent circuits corresponding to each harmonic frequency. Different approaches for deriving the required parameters have also been described in this chapter.

It has been shown that both rotor and stator winding resistances increase with harmonic frequency where rotor resistance is greatly affected. Also the complex nature of the variation of motor leakage reactance with harmonic frequency has been noted. Different approaches have been given in the literature to represent these variations and some of them were discussed in detail and the important parameters have been highlighted in this chapter. A number of harmonic loss models available in the literature have been presented to evaluate the associated harmonic losses in induction motors.

A brief review of the IEEE, IEC and JEC standards has been given and different methods for evaluation of induction motor losses were compared. Application of a calorimetric method for direct measurement of electric machine heating has been discussed. Applicability of a modified version of the calorimetric method for accurate measurement of motor losses under distorted supply condition has been justified.
Chapter 3

Calorimeter Design to Measure Induction Motor Losses

3.1. Introduction

The principle of the calorimetric method originally suggested by the IEC [IEC74] can be applied for direct measurement of losses in small air-cooled induction motors. The calorimetric method is especially suitable in situations where the machine is supplied by distorted waveforms. In this case, standard methods cannot be applied to determine machine losses since they are valid under nominal operational conditions (i.e., fundamental voltage and frequency) of the machine. Under different loading conditions, the calorimetric method provides a convenient arrangement for estimation of motor losses without requiring accurate measurement of motor input and output powers.

A modified calorimetric method has been successfully applied previously to measure losses in a 5.5 kW cage induction motor by developing a single chamber calorimeter [Sha90] [Tur91]. The proposed method was based on performing two subsequent colorimetric tests in which exactly the same coolant flow conditions should be maintained. This was rather a critical task and required a relatively complicated and expensive system to accurately control the air properties during the two tests.

In order to simplify the direct measurement of motor heat loss, development of a new open type calorimeter is proposed in this chapter. Since a double chamber calorimeter (DCC) type is adopted, no critical control of the air properties (except the temperature) is required. The DCC is capable of measuring total losses of a 7.5 kW cage induction motor up to 1 kW.

The principle of the calorimetric method along with the application of a single chamber calorimeter for measurement of losses in an induction motor is given in
this chapter. Details of the design and construction of the DCC is described in this chapter and the dynamic operation of the DCC is discussed. Specifications of the temperature, voltage and current measurement systems and their calibration accuracy are investigated in this chapter.

3.2. Principle of the Calorimetric Method

Basically, in the calorimetric method a mechanism is required so that the generated heat within the machine can be carried away and measured. In large generators, this can be performed by assuming that the majority of the generated heat is transferred to the cooling medium [IEC74]. A small part of the losses which is dissipated by radiation, convection and conduction and is not transmitted to the cooling medium can be calculated separately.

Depending upon the circumstances, two approaches have been suggested to estimate the machine losses using the IEC [IEC74] calorimetric method. In the first approach, referred to as the direct method, the rate of heat transfer by the coolant, \( q(W) \), which represents the machine losses, can be determined using the following energy balance equation:

\[
q = mc_p\Delta T
\]

where the coolant properties are:

- \( m \) = mass flow rate (kg/s),
- \( c_p \) = specific heat (J/kg K), and,
- \( \Delta T \) = temperature rise (K).

This approach demands accurate and continuous measurement of the coolant properties long enough to ensure that thermal equilibrium has been achieved. Certain criteria including the duration of the tests and changes in the temperature rise have been recommended in order to assume that thermal equilibrium has been achieved [IEC74].
However, there are some situations where direct calorimetric method is difficult to implement and/or practically not applicable or economical. An alternative approach, called the calorimetric calibration method, has been recommended where predetermined calibration curves have to be used in order to estimate the machine losses [IEC74]. Calibration curves have to be determined by performing tests under conditions so that the dissipated heat can be measured electrically and within a desired accuracy. These curves indicate the relationship between the dissipated heat loss and temperature rise of the cooling medium. This approach still requires knowledge of coolant properties as well as satisfying certain conditions in order to estimate machine losses accurately.

As stated in the IEC recommendations, both these methods have been utilised for loss measurements in large generators, but they can be applied for measurement of losses in other machines with some modifications. This is particularly required for small size machines where air is the natural cooling medium, as described in the next section.

3.3 Open and Closed Type Calorimeters

The essential feature in the calorimetric method is the measurement of all the generated heat within the machine. Therefore, it is required to employ a controlled mechanism to the cooling system capable of measuring the transferred heat loss. In air-cooled machines, this can be done by surrounding the test machine by a thermally insulated enclosure, generally referred to as the calorimeter, with the minimum possible heat leakage. Depending on the construction a calorimeter can be open or closed type as shown in Figure 3.1 [Tur91].

The principle operation of the open type calorimeter is somewhat similar to the direct calorimetric method suggested by the IEC [IEC74]. A relatively simple construction along with uncomplicated measurement system can be adopted in this approach. The heat loss of the test machine is determined directly by measuring the
air properties at the inlet and outlet passages of the calorimeter and based on the energy balance expressed by Equation (3.1).

Similarly, accurate and continuous measurement of the air properties is required in order to calculate the total heat transfer within the calorimeter. Precautions have to be considered since the air specific heat and density are widely variable depending on the relative humidity and barometric pressure as well as the temperature. Accurate measurement of the air flow rate is also a difficult task to perform due to the non-uniform velocity profile across the inlet and/or outlet passages. The temperature measurement is probably the easiest task in this approach, however, care should be taken to overcome inaccuracies due to the possible non-uniform air temperature distribution, especially across the outlet passage.

In the closed type calorimeter the generated heat by the machine is removed by a predetermined primary coolant (normally air) and transferred to the outside of the calorimeter by means of a heat exchange system [Tur91]. A secondary medium (normally water or oil) along with instruments for measuring its properties are then required to determine the total transferred heat and hence total machine losses. The advantage of this approach is that properties of water or oil can be measured with more confidence due to their smaller variations for a given heat loss when compared with the air. However, the closed type calorimeter is relatively expensive due to the
extra cost of the heat exchange system and is more complicated when compared to the open type one.

In either type of calorimeter, it is essential to provide the normal operating conditions under which the test machine should be operated. In other words, the temperature at different parts of the machine should be kept within the specified limits. In the open type calorimeter this can be achieved by providing sufficient air flow through the calorimeter by means of a fan. In the closed type calorimeter, normal operating conditions can be met by ensuring adequate heat flow via the heat exchange system. However, inherent inaccuracies in the measurement of air properties and flow rate in the open type calorimeter and complexity of the heat exchange system in the closed type calorimeter have prevented wide use of either of these methods. An alternative method of calorimetry has been proposed by Turner et al [Tur91] and is discussed in the next section.

3.4 Balance Calorimetric Method

As reported in [Tur91], an open type calorimeter was first designed and constructed to determine losses of a ventilated 45 HP, 3-phase cage induction motor. The essential difference between this method and that recommended by the IEC [IEC74] was the operation of a resistive heater as an alternative source of heat. In this approach, the calorimetric tests have to be performed in two subsequent parts as main test and balance test.

In the main test, the machine is driven under desired load conditions where air with constant temperature and flow rate is forced into the calorimeter. The test is continued until the steady state condition is achieved, ie when the temperature rise between the inlet and outlet air is constant. In the balance test, the cooling medium, air, should be maintained with the same flow rate as in the main test. The unexcited motor has to be driven by an auxiliary machine outside the calorimeter at the same speed as in the main test. Alternatively, the air should be heated by a heater located inside the calorimeter. The heater electrical input power is then adjusted so that the
same temperature difference between the inlet and outlet air is achieved as in the main test. The total motor losses (excluding the windage and friction losses) are then assumed equal to the electrical power into the heater provided thermal equilibrium has been achieved.

It has been reported in [Tur91] that the method was accurate and suitable for induction motor loss measurement but better results could be achieved by improving the calorimeter design and temperature measurement system. As an improved version, another open type calorimeter was designed and constructed by Turner et al [Tur91] to measure losses of a 5.5 kW TEFV squirrel cage induction motor. Since it was essential to maintain exactly the same conditions in both parts of the test, improvements were made to the calorimeter construction, control and measurement systems. The calorimeter enclosure was constructed using panels of 25 mm thickness expanded polystyrene having a thermal conductivity of less than 0.03 W/m K. A simple air conditioning system was used to maintain the inlet air at 20°C during both parts of the test. The air temperature was measured using platinum resistance thermometers having a resolution of 0.1°C.

In order to monitor the air flow rate through the calorimeter a water manometer indicator was fitted at the input duct. By this, the air barometric pressure was continuously monitored in order to make sure that the air flow rate remains the same during both parts of the test. A feedback voltage regulator was used to supply a constant balanced voltage to the test motor during the first part of the test. Also a closed loop control system was provided to achieve a constant load current. A DC generator was coupled to the test induction motor for loading purposes.

The full load loss of 1 to 1.5 kW was measured using the calorimeter with a typical error of 20 W. The accuracy of the loss measurements using this method was reported to be 4.7% at no-load and 1.45% at full load. It has also been reported that the method is repeatable and capable of measuring losses with a resolution of 9.4 W. However, a number of modifications were suggested to further improve the resolution of the loss measurement. Use of temperature sensors with higher
resolution and averaging the air temperature particularly at the outlet were suggested.

Conducting calorimeter tests in two parts required each calorimetric test to be completed in about 6-8 hours. The possibility of variations in the air properties during the operation of the calorimeter is relatively high. Therefore, careful precautions are required to ensure that those conditions remain reasonably unchanged during both parts of the test. The 8-hour figure is more than twice the time required for conducting either the main or the balance test. Requiring the precise knowledge of the air properties and a relatively complicated measurement and control system as well as the long testing time are considered to be the main disadvantages of this approach. A significant improvement can be achieved by developing a double chamber calorimeter as described in the next section.

3.5 Double Chamber Calorimeter (DCC)

Both calorimeters discussed in Section 3.4 had only a single chamber to house the motor under test and the balance heater. In order to simplify the induction motor calorimetric loss measurement, design and construction of a new open type calorimeter, a double chamber calorimeter (DCC), is proposed. The DCC has a major difference compared with the single chamber type proposed by Turner et al [Tur91], that is, being capable of performing both induction motor and balance tests at the same time.

With the DCC approach, duration of each calorimetric test is about half of that required for the single chamber type at a cost of doubling the calorimeter size. With the availability of relatively cheap insulation material, the DCC is found to be economical. Based on the author's experience, a series of 5 to 6 tests can be performed during a day which is a considerable time saving as compared to the calorimetric setup of [Tur91].

Development of the DCC approach has resulted in a significant simplicity of the calorimeter construction, instrumentation, calorimetric operation and loss
measurement. Unlike the single chamber calorimeter, critical control of the air flow rate through the calorimeter is not required in this approach. Also there is no need to have precise knowledge of the air density, barometric pressure or humidity. Moreover, having a control system to maintain the inlet air at a constant temperature is not essential. In general, it is assumed that any changes in the air properties, except for the temperature, will affect the dynamic operation of the calorimeter in both chambers equally and has no significant effect on the loss measurement procedure. Experimental results confirmed that the same order of accuracy as in the single chamber type can be achieved with the relatively simple and convenient setup of the DCC.

In principle, this method is compatible with a modified version of the calorimetric calibration method suggested by the IEC as reported in the Amendment 2 of IEC publication 34-2 [IEC94] (The author had access to the draft copy of the reference document before the final version was released). As far as the author is aware, the present work is the first contribution in design, construction and implementation of a modified calorimetric method in relation to the IEC proposed method.

3.5.1 Heat transfer mechanism within the DCC

The schematic diagram of the proposed DCC along with the heat transfer paths are shown in Figure 3.2. The generated heat by the excited test motor and the reference heater can be transferred by three conventional methods of heat transfer as convection, conduction and radiation. Therefore, under thermally steady state condition, it can be written that:

\[
P_{\text{motor}} = q_{\text{conv}1} + q_{\text{cond}1} + q_{\text{rad}1} \tag{3.2a}
\]

\[
P_{\text{heater}} = q_{\text{conv}2} + q_{\text{cond}2} + q_{\text{rad}2} \tag{3.2b}
\]

where

\[
P_{\text{motor}} = \text{total losses of the test motor},
\]

\[
P_{\text{heater}} = \text{heater electrical input power},
\]
Chapter 3: Calorimeter Design to Measure Induction Motor Losses

\[ q_{\text{conv}} = \text{heat transfer via convection}, \]
\[ q_{\text{cond}} = \text{heat transfer via conduction}, \]
\[ q_{\text{rad}} = \text{heat transfer via radiation}, \]

and subscripts 1 and 2 refer to the first and the second chambers respectively.

Figure 3.2: Conventional heat transfer mechanism within the double chamber calorimeter (DCC)

The objective of the calorimetric method is to achieve the maximum possible convection heat transfer, \( q_{\text{conv}} \), within the calorimeter. This is done by maintaining sufficient air flow through the calorimeter to take away most of the generated heat by both test motor and the reference heater. By careful design of the calorimeter, this figure can be up to 95% of the total heat transfer within the calorimeter. It must also be noted that the calorimeter should be designed to be large enough to allow sufficient air flow around the test motor and reference heater but not too large to cause a wide temperature distribution in each chamber.

Depending on the air flow through the calorimeter there will be an air temperature rise across each chamber proportional to the dissipated heat within that chamber. Using the principle of the calorimetric method already described in Section 3.2 and by applying energy balance, the convection heat transfer corresponding to each chamber can be estimated as:
Since the calorimeter chambers are sealed and airtight from the outside, the mass flow rate is exactly the same in both chambers (ie $m_1 = m_2$). It must be noted that this quantity can be changed by varying the fan speed.

Air specific heat, $c_p$, however, does change with temperature and relative humidity as illustrated in Figure B.1 of Appendix B. It can be seen that for an increase of 50% in relative humidity (from 50% to 100%) the air specific heat increases only by 0.5% and 3.3% for an average air temperature of 20°C and 50°C respectively. These figures are small enough to give the confidence that the air specific heat is not sensitive to the relative humidity for a temperature range considered in this application. However, with a relative humidity of 50%, the specific heat increases by about 4% when the average air temperature increases from 20°C to 50°C. Therefore, this variation has to be taken into account while using the DCC for evaluation of machine heat loss.

Heat transfer via conduction to the outside of the calorimeter, $q_{cond}$, basically takes place through the calorimeter insulation material, mounting bolts, motor shaft extended to the outside of the calorimeter, and the connection wires. It is realised that for more accurate estimation of machine losses, the calorimeter should be designed so that $q_{cond}$ is kept to the minimum possible value and can be estimated accurately. Experimental tests are conducted to measure the heat leakage through the calorimeter walls and develop a simple thermal model which is described in Chapter 4. Calculation of the conducted heat leakage through the extended shaft is also given in Chapter 4. A preliminary calculation showed a negligible conduction heat leakage through the mounting bolts and the connection wires.

Radiation heat transfer from any surface is proportional to the fourth power of the absolute temperature of that surface ($T^4$). The radiated heat transfer between the
test induction motor and calorimeter walls is estimated to be very small since their surface temperature difference is relatively small. The radiated heat from the heater element is relatively high and therefore care should be taken in the design of the resistive heater. Precautions are considered to reduce the effective heater surface temperature to the same level as the induction motor.

In order to further minimise the radiated heat transfer, distribution of thin layers of insulation material around the reference heater and induction motor has been proposed. It is assumed that these shields absorb the radiated heat which, eventually, will be dissipated into the cooling medium and transferred by convection. Therefore, any possible radiated heat will contribute to the convection heat transfer part, $q_{\text{conv}}$, and hence terms $q_{\text{rad1}}$ and $q_{\text{rad2}}$ in Equations (3.2a) and (3.2b) can be neglected. These shields also served an important role in directing and adequately mixing the air in both chambers.

Accordingly, by combining Equations (3.3a) and (3.3b), and under thermally steady state conditions, the convection heat transfer in the first chamber can be estimated as:

$$q_{\text{conv1}} = q_{\text{conv2}} \frac{c_{p1}}{c_{p2}} \frac{\Delta T_1}{\Delta T_2}$$

(3.4)

where

$$q_{\text{conv2}} = P_{\text{heater}} - q_{\text{cond2}}$$

(3.5)

Total machine losses, $P_{\text{motor}}$, can be calculated as:

$$P_{\text{motor}} = q_{\text{conv1}} + q_{\text{cond1}} = \left( P_{\text{heater}} - q_{\text{cond2}} \right) \left( \frac{c_{p1}}{c_{p2}} \frac{\Delta T_1}{\Delta T_2} \right) + q_{\text{cond1}}$$

(3.6)

where all the terms on the right hand side of Equation (3.6) can be measured and/or estimated.

The conducted heat leakage is assumed to be a small portion of the total heat loss (about 5%). Therefore, the accuracy of the heat loss measurement is directly
dependant on the accuracy $P_{heater}$ as well as the air temperature measurement across each chamber, $\Delta T_1$ and $\Delta T_2$. Consequently, these quantities have to be measured carefully in order to obtain a desired accuracy. Details of the measurement system are given in Section 3.5.5 and the accuracy of the loss measurement is investigated and discussed in Chapter 4.

### 3.5.2 Design and construction of the DCC

The DCC is designed based on the discussion given in previous section and schematic diagram shown in Figure 3.3. The calorimeter is made using plane slabs of insulation material, class VH expanded polystyrene (EPS) (Appendix B), having an average thermal conductivity of 0.035 W/m K. This type of material is commercially available at low cost. It is light but strong enough to allow the calorimeter enclosure to stand alone. A water clean-up building adhesive (Selleys Supa Nails) is applied to join the EPS slabs. The external dimensions of the DCC are 1300 mm × 750 mm × 600 mm and the thickness of the EPS slabs is 100 mm.

Another plane slab of EPS is used to separate the calorimeter into two adjacent chambers to house the induction motor under test and the reference heater respectively. One of the side panels is arranged to be removable in order to have access to the inside of the calorimeter. Two plastic pipes of 150 mm diameter are mounted on both ends as air inlet and outlet ducts. A variable speed fan is mounted at the inlet duct to force the air through the calorimeter chambers. Specifications of the fan is given in Appendix C.

The test machine is a 7.5 kW high efficiency cage induction motor with nominal line-to-line voltage of 415 V. Assuming a 90% efficiency for the test motor, a motor loss of up to 850 W is to be measured. In this application, motor losses are expected to be more since distorted voltages are to be applied to the test motor resulting in extra harmonic losses. Therefore, the DCC is designed to have a rating of 1 kW for loss measurement.
The test motor is mounted on two pieces of hard wood of 32 mm thickness to allow adequate air flow under the motor. A bakelised canvas plate of 400 mm × 400 mm × 6 mm is placed under the motor as shown in Figure 3.3. Both wooden bases and the plate are chosen strong enough to sustain the induction motor weight without causing destruction to the insulation material. This arrangement is quite adequate to perform no-load tests on the machine. Further modifications are made to the DCC in order to perform loaded machine tests as described in Section 3.5.4.

![Figure 3.3: Schematic diagram of the constructed double chamber calorimeter (DCC) housing the test motor and the reference heater](image)

### 3.5.3 Reference heater

The power rating for the reference heater is selected to be 1 kW, of the same order as the induction motor losses. Commercially available 1 kW heater elements have surface temperatures in excess of several 100°C when operating at full power. This would result in a high proportion of radiation heat transfer between the heater element and calorimeter walls.

In order to reduce the effective heater surface temperature the heater element was mounted within a cylindrical metal enclosure of 150 mm diameter and 10 mm thickness. Having a fan in front of the heater resulted in more air flow around it and further reduction of the surface temperature to the same level as the induction motor. Both the cylindrical enclosure and the fan facilitated a closer match between the geometry and thermal conditions for the reference heater and the induction
motor under test. The reference heater was also mounted on a wooden base of 50 mm thickness via two metal bars. In addition to the mechanical support, the wooden base served a proper insulation between the heater body and calorimeter floor.

3.5.4 Loaded machine mechanism

In order to perform experimental work under loaded conditions some modifications were made to the calorimeter after performing no-load tests on the test motor. A schematic diagram of the calorimeter modified for performing loaded machine tests is shown in Figure 3.4. A test bed was designed and constructed so that the test machine can be coupled to a DC generator outside the calorimeter. The insulation material under the test motor was mechanically supported by applying two plates of bakelised canvas inside and outside the calorimeter in a sandwich form as shown in Figure 3.4.

Two wooden bases already mounted under the test motor were bolted to the metal test bed through the calorimeter floor using four bolts. This arrangement has two advantages: (i) prevents a direct contact between the motor base and the metal test bed and (ii) provides a simple arrangement to replace different frame size machines to be tested. The first advantage is to reduce the conducted heat leakage through the mounting bolts due to the lower temperature of the wooden base as compared with the motor base. This has resulted in a negligible conducted heat leakage through the mounting bolts due to the small temperature difference across each bolt.

The most challenging aspect in the loaded machine arrangement was to take the shaft of the test motor out of the calorimeter without having air leakage around the shaft. For this purpose, a stuffing box was designed and fitted into the one of the side walls of the calorimeter. The stuffing box was made using screw cap PVC pipes with the diameter of 120 mm. Two holes were cut on the caps to allow the extended shaft to go through. The stuffing box was filled with soft polyester fibre
material, enough to prevent air leakage from the inside to the outside of the calorimeter.

Preliminary tests confirmed an extra 20 W as friction losses due to the presence of the stuffing box. It was done by performing two reduced voltage tests on the unloaded machine and comparing windage and friction losses under two conditions, once with the stuffing box full of the material and once without the material. This loss is basically caused by friction between the outer surface of the extended shaft and the stuffing box material and is assumed to be reasonably constant during a test. Heat leakage through the cross section of the stuffing box is assumed to be negligible because of the relatively small cross sectional area as compared to the whole calorimeter. This assumption is valid since the thermal conductivity of the stuffing box material is relatively low and is of the same order as the EPS.

![Schematic diagram of the calorimeter for loaded machine tests](image)

Figure 3.4: Schematic diagram of the calorimeter for loaded machine tests

Since the motor shaft temperature is relatively high, a flexible coupling was used inside the calorimeter for torque transmission to the DC generator via an extended shaft. The coupling has a high thermal resistance and hence prevents excessive heat leakage through the shaft. The two machines were coupled via a flexible universal coupling as shown in Figure 3.4.
3.5.5 Instrumentation and measurement system

3.5.5.1 Temperature measurement system
A variety of sensors are available for air temperature measurements. In this project two types of measuring sensors, namely, Resistance Temperature Detectors (RTDs) and Thermocouples are used. A data acquisition (DA) system in conjunction with a computer software package (Labtech Notebook) is also employed for data logging/analysis. Specifications of the DA system is provided in Appendix C.

A. Absolute temperature measurement using RTDs
The RTDs have a platinum sensor (PT100) located in a metal enclosure at the tip. The PT100 sensors have a resistance of about 100 Ω at 0°C which increases linearly with the temperature within a certain range. In this application eight RTD sensors capable of working in 0-200°C range are used. Each RTD sensor is connected to an RTD module for signal conditioning via a 4-connection wire. Corresponding to 0-200°C, each RTD module then produces a 0-5 V signal, giving a resolution of 25 mV per °C. All RTD modules are mounted on an ISO-RACK and connected to an eight channel I/O computer board for A/D conversion and data logging with DA system. Initial calibration of the RTDs in conjunction with the DA system indicated an accuracy of ±0.1°C for each individual RTD within the range 0-100°C.

RTD sensors are used to measure the absolute temperature at different points inside the calorimeter, calorimeter walls, induction motor frame, heater chassis and outside wall. Temperature distributions inside the calorimeter are examined by locating RTD sensors at different points inside the calorimeter chambers. The collected temperatures can be used to calculate the conducted heat leakage corresponding to each chamber, \( q_{\text{cond1}} \) and \( q_{\text{cond1}} \), as required by Equation (3.6).

A simple thermal model is developed to relate the calorimeter conducted heat leakage and the temperature difference between the inside and outside of the calorimeter. For the size of the constructed DCC a heat leakage of 1.7 W is estimated for one degree Celsius of temperature difference between the inside and
outside the calorimeter. Details of the thermal model as well as the experimental tests are presented in Chapter 4.

B. Relative temperature measurement using thermopiles

According to Equation (3.6), it is desirable to measure the air temperature differences, $\Delta T_1$ and $\Delta T_2$, across each chamber rather than the absolute air temperatures. For this purpose, the use of thermocouple wires in the form of a thermopile has attracted attention. Applying this simple technique has the following advantages:

(i) measurement of the air temperature difference across each chamber directly,
(ii) being able to average the non-uniform temperature distribution across the inlet and outlet throats, and,
(iii) requiring no compensation circuit to give the absolute temperature and consequently providing a simple temperature measurement circuit.

In this particular application T type thermocouples (Copper vs Constantan) are used to form two separate thermopiles. A thermopile can be considered as a number of floating variable voltage sources connected in series where their voltages are proportional to the junctions temperature. Choosing the number and position of the junctions depends on the temperature distribution of the subject whose temperature is to be measured.

In this application, the moving air through the calorimeter has a turbulent flow at the inlet and outlet ducts. Therefore, it is assumed that the temperature distribution across the inlet and outlet ducts are fairly uniform. However, slight differences in temperature distribution was detected and hence it was decided to place a number of thermocouple junctions across the inlet and outlet air throats. Considering the cross sectional area of the throats, five thermocouple junctions were placed at each side. By this, the measured thermoelectric voltage of each thermopile is proportional to one fifth of the average temperature difference across each chamber. A simple drawing for a thermopile having five junctions on each side is shown in Figure 3.5.
The thermoelectric voltage of each thermopile was measured using a PC-based data acquisition (DA) system having a 12 bit resolution within a range ±5 V. Since the thermopile thermoelectric voltage is relatively small (about 0.2 mV for a temperature difference of 1°C) the gain of the related channel on signal conditioning board was set to the maximum available value of 800 to obtain the highest resolution. For ease of calculations a software scale factor of 1.25 was also applied to the measured signal resulting in an overall magnification of 1 V per mV.

![Diagram of thermocouples](image)

**Figure 3.5:** A simple arrangement of thermocouples to form a thermopile

Anti-aliasing RC filters with cutoff frequency of 4 Hz were also connected at the output terminals of the thermopiles to by-pass high frequency noise as well as the 50 Hz noise from the measured signals. In order to eliminate the EMI on the thermopiles' signal the thermocouple wires were twisted and covered by metal shields which were grounded to the system earth on one end. These precautions served a significant reduction in noise and EMI in the thermopile's thermoelectric voltages.

**C. Calibration of thermopiles**

Variation of thermoelectric voltage of a thermocouple junction with temperature is non-linear. For different thermocouples, different lookup tables are available to convert the thermoelectric voltage to temperature. Typical curves and lookup tables
for different type thermocouples are shown in Appendix B. There are also some different order polynomials suggested by standards for voltage to temperature conversion available for different type thermocouples [NI93]. However, depending on the application, thermocouples need to be calibrated in conjunction with the associated circuits, such as zero compensation circuits, within the required accuracy and the specified working temperature range.

T type thermocouples have a working temperature in the range of -270 to +400°C. According to the data sheet with reference junction at 0°C, the generated thermoelectric voltage is about 40 μV/°C for an average temperature of 25°C which increases to about 60 μV/°C for an average temperature of 300°C. It must also be noted that the standard tolerance for T type thermocouples, being due to their material, is about ±1°C [Guy89].

Since the working temperature in this application is between the room temperature and 50°C and since the thermocouples are used in a particular form of thermopile, it was decided to calibrate them in a specified temperature range in conjunction with the DA system. For this purpose, two thermally insulated water baths were used. The temperature of the water in each bath was adjusted using a thermostat control resistive heater. In order to achieve uniform water temperature, stirring pumps were used to mix the water in each bath. Two laboratory grade mercury-in-glass thermometers with resolution of 0.1°C were used as references for water temperature measurements in each bath, $T_1$ and $T_2$. The absolute accuracy of the thermometers was not known, however, they were calibrated against each other for the temperature range 10-50°C. Accordingly, using these thermometers, a maximum uncertainty of ±0.1°C was obtained in the measurement of temperature difference $\Delta T (= T_1 - T_2)$.

Temperature measurements were made with the two sides of the thermopiles inserted into water baths. The calibration process was started by adjusting the water temperature in one bath at about 50°C and the other at about 47°C. The latter was reduced to about 40°C in a few steps to cover a temperature difference in the range
of 3-10°C. Using the DA system, the output voltage of the thermopiles were measured at a sampling rate of 100 Hz for appropriate duration. For each test enough time was allowed to achieve a stable water temperature in each bath. This was assumed to be met when the changes in the thermopile voltage was constant at a value equivalent to 0.1°C. This process was repeated by reducing the water temperature down to about 25°C.

The absolute water temperatures in both baths, $T_1$ and $T_2$, measured by thermometers, along with the calculated average temperature, $T_{avg}$, and temperature difference, $\Delta T = T_1 - T_2$, are shown in Table 3.1. The corresponding values for the thermopiles' thermoelectric voltages, $\Delta V_1$ and $\Delta V_2$ measured by DA system, are also shown where a negligible discrepancy can be seen between the two. This is basically within the given range of accuracy and is due to the resolution of the measurement system. The last column in Table 3.1 shows the slope $\Delta V / \Delta T$ which is in the range 0.210-0.238 mV/°C for an average temperature changing from 25 to 50°C. Such a variation is expected since the thermocouple thermoelectric voltage is a non-linear function of the average working temperature. Therefore, in calibration of the thermopiles, the average working temperature, $T_{avg}$, should be taken into account.

<table>
<thead>
<tr>
<th>$T_1$ (°C)</th>
<th>$T_2$ (°C)</th>
<th>$T_{avg}$ (°C)</th>
<th>$\Delta T$ (°C)</th>
<th>$\Delta V_1$ (mV)</th>
<th>$\Delta V_2$ (mV)</th>
<th>Slope ($\Delta V / \Delta T$) (mV/°C)</th>
</tr>
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<tbody>
<tr>
<td>51.7</td>
<td>49.7</td>
<td>50.7</td>
<td>2.0</td>
<td>0.468</td>
<td>0.472</td>
<td>0.2340</td>
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<tr>
<td>51.7</td>
<td>46.0</td>
<td>48.9</td>
<td>5.7</td>
<td>1.275</td>
<td>1.276</td>
<td>0.2234</td>
</tr>
<tr>
<td>51.6</td>
<td>43.6</td>
<td>47.6</td>
<td>8.0</td>
<td>1.826</td>
<td>1.815</td>
<td>0.2283</td>
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<tr>
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<td>46.3</td>
<td>10.7</td>
<td>2.393</td>
<td>2.385</td>
<td>0.2236</td>
</tr>
<tr>
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<td>40.6</td>
<td>44.1</td>
<td>7.0</td>
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<td>1.561</td>
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</tr>
<tr>
<td>44.1</td>
<td>40.5</td>
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<td>3.6</td>
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<td>0.852</td>
<td>0.2383</td>
</tr>
<tr>
<td>42.2</td>
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<td>40.5</td>
<td>3.4</td>
<td>0.805</td>
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<tr>
<td>39.7</td>
<td>35.9</td>
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<td>0.813</td>
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<tr>
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<td>6.8</td>
<td>1.429</td>
<td>1.432</td>
<td>0.2101</td>
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<tr>
<td>39.7</td>
<td>30.0</td>
<td>34.9</td>
<td>9.7</td>
<td>2.049</td>
<td>2.046</td>
<td>0.2112</td>
</tr>
<tr>
<td>34.9</td>
<td>30.3</td>
<td>32.6</td>
<td>4.6</td>
<td>1.002</td>
<td>0.996</td>
<td>0.2178</td>
</tr>
<tr>
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<td>31.2</td>
<td>1.8</td>
<td>0.419</td>
<td>0.415</td>
<td>0.2328</td>
</tr>
<tr>
<td>31.4</td>
<td>28.6</td>
<td>30.0</td>
<td>2.8</td>
<td>0.616</td>
<td>0.616</td>
<td>0.2200</td>
</tr>
<tr>
<td>31.0</td>
<td>27.3</td>
<td>29.2</td>
<td>3.7</td>
<td>0.807</td>
<td>0.801</td>
<td>0.2181</td>
</tr>
<tr>
<td>30.9</td>
<td>25.4</td>
<td>28.2</td>
<td>5.5</td>
<td>1.198</td>
<td>1.192</td>
<td>0.2178</td>
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<tr>
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<td>27.0</td>
<td>3.2</td>
<td>0.696</td>
<td>0.689</td>
<td>0.2175</td>
</tr>
</tbody>
</table>

Table 3.1: Calibration of thermopiles using mercury-in-glass thermometers
By curve fitting to the collected data two expressions are derived to estimate the slope $\Delta V / \Delta T$ as a function of the average working temperature corresponding to each thermopile. As shown in Figure 3.6, a straight line is fitted to the experimental data, $\Delta V / \Delta T$, to relate the average working temperature with the thermopile slope. This slope will be used to calculate the temperature difference for a given thermopile voltage.

![Graph of experimental data and fitted line for thermopile slope](image)

**Figure 3.6:** Experimental data and fitted line for thermopile slope (mV/°C) as a function of average working temperature, $T_{avg}$

It should be noted that variations of thermopile voltage with a maximum temperature difference of 10°C was assumed linear. This assumption is subject to a negligible error since the non-linearity only becomes significant when higher temperature differences are applied. A comparison between the experimental data and estimated values using the fitted line demonstrated a maximum discrepancy of 0.13°C and an average difference of 0.05°C. The calibration procedure confirmed that the temperature difference, $\Delta T$, can be measured with maximum uncertainty of 0.1°C using the calibrated thermopiles.
3.5.5.2 Voltage and current measurement circuits

Measurement of electrical quantities such as voltage, current and power under sinusoidal supply conditions can be performed using standard laboratory instruments. However, under distorted supply conditions such instruments may not provide reliable results due to the limited frequency bandwidth. For this reason and in order to have access to the induction motor and reference heater voltage and current waveform data via the DA system, voltage and current measurement circuits were designed and implemented. For voltage measurement circuits appropriate voltage dividers along with the isolation amplifiers were used. Hall effect current transducers were employed for induction motor and reference heater line current measurements. The circuit diagram of the motor voltage and current measurement system is shown in Figure 3.7. Specifications of the voltage isolation amplifiers and Hall effect current transducers are given in Appendix C.

![Figure 3.7: Three phase voltage and current measurement circuits](image)

Both voltage and current measurement circuits were calibrated in conjunction with the DA system using known input quantities. A high precision digital multimeter was employed as the reference. Accordingly, accuracy of both voltage and current
measurement circuits were calculated to be ±1%. The calibration accuracy of these circuits was checked from time to time.

During each calorimetry test, the required data for current and voltage waveforms were captured and saved in data files for further analysis. Averaging of the data, calculation of the rms values of voltages and currents, active power calculations, FFT on the current and voltage waveforms etc are some examples of post data analysis required in this application.

3.6. Conclusions

The suitability of a double chamber calorimeter (DCC) capable of measuring induction motor losses up to 1 kW has been justified in this chapter. The calorimeter is especially applicable under harmonically distorted conditions where motor losses can be estimated directly and regardless of the motor input voltage and current waveforms. The calorimetric method also provides a convenient setup for measurement of motor losses under various loading conditions without being affected by inaccuracies due to the output power measurement.

Critical control and measurement of the air properties has always been a major concern in previously developed calorimeters. It has been demonstrated that how the use of two adjacent chambers in the DCC results in a significant simplicity of the heat loss measurements. The DCC is particularly advantageous since no critical control of the air properties, except the temperature, is required in the process of heat loss estimations. The dynamic operation of the DCC has been described in this chapter by developing the heat transfer mechanism in both chambers of the calorimeter.

A mechanism has been designed for performing loaded machine tests while the test motor operates inside the calorimeter. The motor shaft has been extended to the outside of the calorimeter through a stuffing box installed on one of the side walls. Precautions have been considered to prevent air leakage around the shaft and also to minimise heat leakage through the stuffing box, extended shaft and mounting bolts.
The importance of the air temperature measurement across each chamber on the accuracy of the heat loss measurement is highlighted. A simple but efficient technique has been developed for measurement of air temperature rise using T type thermocouples in the form of a thermopile. Calibration of thermopiles has been described and it is demonstrated that the air temperature rise can be measured with an accuracy of 0.1°C. Temperature distributions inside the DCC has been examined by measurement of absolute temperature using RTDs with an accuracy of ±0.1°C. Voltage and current measurement circuits have been calibrated using a high precision multimeter and demonstrated to have a 1% accuracy.
4.1 Introduction

The suitability of the calorimetric method for the measurement of induction motor losses has been justified and a DCC has been developed as described in Chapter 3. There are a number of interactive parameters which affect the thermodynamic conditions of the DCC in the process of the motor heat loss measurements. The air flow rate thorough the calorimeter, the rate of temperature rise within the calorimeter, the level of reference heater power, and the amount of heat leakage through the insulation material are among these parameters which should be carefully investigated before starting the calorimeter tests on the induction motor.

The rate of air temperature rise within the calorimeter depends on the dissipated heat loss and the air flow rate through the calorimeter. The higher the heat loss, the larger the temperature rise can be within the calorimeter. The temperature rise can be reduced by increasing the air flow rate, but a high air flow rate, on the other hand, can cause a small air temperature rise reducing the accuracy of the loss estimation.

This chapter deals with effects of various parameters on heat loss measurement using the DCC. The rate of the conducted heat leakage through the calorimeter insulation material is examined by performing some basic thermal tests. The dynamic operation of the DCC is investigated by performing a series of tests using a known heater instead of the test motor. The experimental results are utilised to carefully determine limits for DCC parameters such as air flow rates through the calorimeter chambers, maximum temperature rise inside and across each chamber and the reference heater input power level. Accuracy of the heat loss measurement using the DCC is also examined both experimentally and analytically.
4.2 Conducted Heat Leakage

4.2.1 Heat Conduction through a Plane Slab

Heat flow through an insulation material due to a certain temperature difference is basically similar to the flow of current I in a resistor R due to a potential difference V (Ohm's Law). Under thermal steady state conditions and assuming one dimensional heat flow the heat conduction, q in Watts (W), through a plane slab can be calculated as:

\[ q = \frac{\Delta T}{R} \]  

(4.1)

where \( \Delta T \) is temperature difference across the slab in Kelvin (K) and R is the thermal resistance of the slab in Kelvin per Watts (K/W) [Inc88]. Thermal resistance of the slab, R, is calculated in the same manner as the electrical resistance of a conductor:

\[ R = \frac{t}{kA} \]  

(4.2)

where \( t \) is the thickness of the slab in meters (m), \( k \) is the thermal conductivity of the slab material in Watts per meters-Kelvin (W/m K) and \( A \) is the area of the slab in meter squared (m\(^2\)). For the constructed calorimeter \( t = 0.1 \) m and \( A = 4.41 \) m\(^2\) which represents the internal area of the calorimeter enclosure formed by six plane slabs of insulation material (EPS). Thermal conductivity of the EPS varies linearly with the average working temperature, \( T_{\text{avg}} \) (°C), between 0.03 and 0.04 W/m K as:

\[ k(T_{\text{avg}}) = a + bT_{\text{avg}} \]  

(4.3)

where the values for \( a \) and \( b \) have been derived using the graph given in the EPS data sheet (Appendix B) as:

\[ a = 0.0321 \]

\[ b = 0.00013. \]
For instance, at a room temperature of 25°C and temperature difference of 20°C across the EPS plane slab, the average temperature will be 35°C (25+20/2=35) and hence \( k = 0.0367 \text{ W/m K} \). Therefore, using Equations (4.1) and (4.2), the rate of the conducted heat leakage through the calorimeter walls from interior to exterior can be calculated as:

\[
q = \frac{\Delta T}{R} = \frac{20}{0.1/(0.0367 \times 4.41)} \approx 32 \text{ W} \tag{4.4}
\]

### 4.2.2 Heat Conduction through the extended shaft

The procedure discussed in section 4.2.1 can be applied to calculate the conducted heat leakage through the motor shaft which is extended from the inside to the outside of the calorimeter through the stuffing box. The effective length of the extended shaft is 120 mm and its diameter is 40 mm giving a cross sectional area of \( 1.257 \times 10^{-3} \text{ m}^2 \). The thermal conductivity of the steel is 58 \( \text{W/m K} \) [Inc90] and hence the one-dimensional conducted heat leakage through the cross sectional area of the extended shaft, \( q_{\text{shaft}} \), can be calculated as:

\[
q_{\text{shaft}} = \frac{\Delta T}{R_{\text{shaft}}} = \frac{1}{0.12/(58 \times 1.257 \times 10^{-3})} \approx 0.6 \text{ W/°C} \tag{4.5}
\]

where \( R_{\text{shaft}} \) stands for the effective thermal resistance of the extended shaft which is subject to a known temperature difference. For instance, a heat leakage of 6 W can be calculated if the average temperature difference of 10°C is applied across the extended shaft.

Under thermal steady state condition, it can be assumed that the shaft temperature inside the calorimeter is the same as the average air temperature in chamber 1. Similarly, the shaft temperature outside the calorimeter can be assumed to be the same as the ambient temperature. Therefore, the average temperature difference between the inside and outside of the chamber 1 can be utilised for calculation of \( q_{\text{shaft}} \). However, experimental results confirmed that the shaft temperature outside the calorimeter is more than the ambient temperature by a few °C due to the friction
losses of the plummer block which is located very close to the stuffing box. This resulted in a situation where the temperature difference across the extended shaft was very small and hence the conducted heat leakage via the shaft was neglected without introducing a great error.

4.3 Measurement of Calorimeter Heat Leakage

In order to measure the calorimeter heat leakage through the walls a series of tests was performed using a low power heater. The heater was placed inside the calorimeter (without induction motor, reference heater and separating barrier) as illustrated by Figure 4.1. In order to avoid stratification a small DC fan was also placed inside the calorimeter to allow adequate air mixing. The fan input power was assumed to be converted to heat and transferred to the air inside the calorimeter and hence was considered as part of the total input power, $P_{in}$ (ie $P_{in}=P_{heater}+P_{fan}$).

The whole calorimeter was sealed and made airtight by introducing two circular pieces of EPS fitted into the plastic pipes of the inlet and outlet air ducts. In this case, the generated heat inside the calorimeter is dissipated only through the calorimeter walls. Under thermal equilibrium the conducted heat leakage through the calorimeter enclosure is assumed to be exactly equal to $P_{in}$. For a fairly constant ambient temperature, thermal equilibrium is achieved when there is no significant change in the temperature inside the calorimeter for a reasonable period of time (eg a temperature changes of ±0.1°C for a period 20-30 minutes).

![Figure 4.1: Experimental setup for measurement of calorimeter conducted heat leakage through the walls](image)
In order to measure the air temperature, a number of RTDs were distributed in different positions inside the calorimeter as well as the ceiling, the side walls and the floor as shown in Figure 4.1. Since the calorimeter was initially placed on a wooden bench in a large room, the temperature of outside walls was assumed uniform and constant at the ambient temperature. This was justified by measuring the air and wall temperatures at different positions outside the calorimeter where a maximum temperature difference of $0.5^\circ\text{C}$ was detected. Therefore, having one RTD in a position to measure the average outside wall temperature was considered to be sufficient.

### 4.3.1 Test Procedure and test results

There are two alternatives for running this test. In the first approach, a test was started with $P_{in}$ adjusted to about 30 W. According to the heat loss model described in Section 4.2, after reaching steady state, this amount of dissipated heat causes a temperature rise of more than $18^\circ\text{C}$ for the air inside the calorimeter. In this particular case the test was conducted for about 8 hours to reach steady state conditions. This was mostly due to the use of a small heat flow rate (30 W) to warm up about $0.6\text{ m}^3$ of air by $18^\circ\text{C}$.

However, it was realised that there is a quicker way of performing these tests without affecting the dynamics of the heat leakage measurement process. In the second approach, a higher power can be supplied to the heater, but for a short time enough to allow the air to reach a temperature close to the final one. This can be interpreted as the pre-heating process which takes 5-15 minutes to be completed for a heater input power of 100-200 W. The heater input power then has to be reduced to the original set value and the test should be continued until the thermally steady state condition is achieved. Using this approach, the previous test was completed within 2 hours where similar results were achieved and hence this method was adopted for further tests.
Two more tests were conducted with input power, $P_{in}$, adjusted to 15 and 50 W to cover the entire operating range of the calorimeter in this application. The corresponding temperature rise of about 10°C and 30°C for the air inside the calorimeter was achieved. The measured values of $P_{in}$ and the difference between the spatially averaged temperatures inside and outside the calorimeter, $\Delta T$, are shown in columns 2 and 3 of Table 4.1 respectively. A preliminary calculation demonstrated that the measured input power, $P_{in}$, was about 13% higher than the calculated conducted heat leakage through the calorimeter walls using the method described in Section 4.2.1. It was concluded that the heat leakage through the edges and corners of the calorimeter must be appreciable and needed to be taken into account. This issue was further investigated as described in the next sub-section.

4.3.2 Heat leakage through the calorimeter edges and corners

To calculate the heat leakage through the edges and corners of the calorimeter, a parameter known as conduction shape factor, $S$, should be used [Inc90]. It is a function of geometry and temperature distributions of the material and can be calculated analytically or numerically. Thus, Equations (4.1) and (4.2) can be combined and modified to give the conducted heat leakage as:

$$q = k S \Delta T$$  \hspace{1cm} (4.6)

For a plane slab with an area $A$ and thickness $t$, two-dimensional edge sections with length $L$ and three-dimensional corners formed by walls having a thickness $t$, $S$ is given in [Inc90] as:

$$S_{plane \ slab} = \frac{A}{t}$$
$$S_{edge} = 0.54L$$
$$S_{corner} = 0.15t.$$

(4.7)
Table 4.1: Measured and calculated values for the calorimeter conducted heat leakage in different tests

|       | P<sub>in</sub> (W) | ΔT (°C) | q<sub>walls</sub> (W) | q<sub>edges</sub> (W) | q<sub>corners</sub> (W) | q<sub>total</sub> (W) | |P<sub>in</sub> - q<sub>total</sub>| |
|-------|---------------------|---------|------------------------|-----------------------|-----------------------|----------------------|----------------|---|
| Test 1 | 51                  | 28.8    | 44.5                   | 6.1                   | 0.2                   | 50.8                 | 0.2           |   |
| Test 2 | 32                  | 18.5    | 28.2                   | 3.9                   | 0.1                   | 32.2                 | 0.2           |   |
| Test 3 | 16                  | 9.6     | 14.3                   | 2.0                   | 0.1                   | 16.4                 | 0.4           |   |

Accordingly, the heat leakage through the calorimeter walls, edges and corners were calculated for the previously described tests (being in the range 15-50 W) as shown in Table 4.1. The maximum discrepancy between the measured input power, P<sub>in</sub>, and the calculated total heat leakage, q<sub>total</sub>, is 0.4 W. This confirms that the conducted heat leakage through the calorimeter edges is significant since it contributed to more than 10% of the total heat leakage. However, it can be seen that the heat leakage through the calorimeter corners is very small and can be neglected.

The variation of the calculated total heat leakage through the whole calorimeter and the measured input power as a function of ΔT is illustrated in Figure 4.2. It can be seen that the calorimeter heat leakage increases linearly with temperature difference, ΔT, within the temperature range considered here. The gradient of the graph demonstrated an average heat leakage of 1.7 W per one degree Celsius of temperature difference, ΔT. Considering the resolution of ±0.1°C for RTDs, the conducted heat leakage can be estimated with a resolution of ±0.2 W. However, the accuracy of the wattmeter used to measure the heater input power was ±2% giving a maximum uncertainty of ±1W in the estimated conducted heat leakage.

The heat leakage characteristics of the calorimeter determined in this work is required to be inferred during future calorimetric tests on the induction motor. Based on the described model, a spreadsheet was developed to take the inside and outside temperatures as input parameters, calculate the thermal conductivity of the
EPS at the average temperature, and finally calculate the total heat leakage for each chamber of the calorimeter separately.

![Graph showing conducted heat leakage through the calorimeter vs temperature difference (ΔT)](image)

**Figure 4.2:** Measured ($P_{in}$) and calculated ($q_{total}$) conducted heat leakage through the calorimeter vs temperature difference (ΔT)

### 4.4 Calorimeter Calibration using Two Identical Heaters

In order to investigate the operation of the DCC, a series of experiments was performed using a heater (test heater) instead of the test induction motor. The main objective was to have more control on the heat generated in the first chamber without involving the test induction motor. This arrangement allows the effect of various air flow rates through the calorimeter and different input power levels supplied to the heaters to be investigated.

The calorimeter accuracy can be examined by comparing the predicted heat loss for the test heater and its electrical input power measured by a wattmeter. The repeatability of the heat loss estimation can also be investigated by repeating similar tests under different environmental conditions. For convenience and in order to provide the closest possible match between the thermodynamic conditions in the
two chambers the test heater was constructed exactly the same as the reference heater already described in Chapter 3.

An illustration of the DCC housing both reference and test heaters surrounded by radiation shields/air mixing baffles is shown in Figure 4.3. RTD sensors are placed at different positions inside and outside the calorimeter for absolute air temperature measurement. Air temperature rise across the two chambers, $\Delta T_1$ and $\Delta T_2$, are measured using calibrated thermopiles.

Using Equation (3.6), the dissipated heat by the test heater, $P_{th}$, can be estimated as:

$$P_{th} = (P_{rh} - q_{cond_2}) \left( \frac{c_{p1}}{c_{p2}} \frac{\Delta T_1}{\Delta T_2} \right) + q_{cond_1}$$

(4.8)

where $P_{rh}$ stands for the electrical input power into the reference heater. The other quantities in Equation (4.8) have already been defined in Chapter 3.

![Figure 4.3: Double chamber calorimeter (DCC) housing the test and reference heaters for calibration](image)

4.4.1 Test Procedure

A series of tests was carried out with different heater input power levels and different air flow rates through the calorimeter. In all the tests both DC fans were operating and their input power, being about 12 W, were calculated by measuring their supply DC voltage and current and added to the heaters' input power. In the early stages it was realised that the maximum temperature in each chamber will
occur on the wall in front of the heater due to direct air flow produced by the DC fan mounted on each heater. Therefore, another layer of EPS was placed in front of each heater to reduce the wall temperature and allow for better mixing of the air and hence more uniform temperature distribution inside the calorimeter chambers.

Heater input power levels were selected to be between 200-800 W to cover a wide range of heat loss measurement using the DCC. Both heaters were independently supplied by single-phase variacs and their electrical input power levels were measured using the EMTEK-6000 single-phase digital wattmeters. Since both wattmeters were exactly the same (their calibration was checked in a separate test), the relative error due to the wattmeter readings was neglected. Three different air flow rates of 30, 55 and 80 L/s were maintained through the DCC while conducting these tests. Air volumetric flow rates were estimated by averaging the measured air velocity profile across the outlet duct using a hot probe anemometer.

Each test was continued long enough to ensure that the thermally steady state condition is achieved. This condition was met when the changes in the temperature rise inside and across each chamber was within ±0.1°C for a period of 10-15 minutes. During each test, the average input power supplied to the heaters, DC fans input power, absolute temperatures inside the calorimeter chambers and thermopiles' thermoelectric voltages (corresponding to the air temperature rise across each chamber) were measured at appropriate time intervals and recorded separately for further analysis.

4.4.2 Temperature distribution and heat leakage calculation

Air temperature distribution inside each chamber is a function of heater input power and the air flow rate through the calorimeter. However, for a given heater input power and air flow rate, some differences in the order of couple of degrees Celsius were observed between the RTD readings in each chamber. The measured temperatures were then averaged and used to calculate the conducted heat leakage through each calorimeter chamber for different tests. A preliminary analysis
showed that the estimated conducted heat leakage using the averaged temperatures in each chamber is subject to a negligible error when compared to that calculated using individual temperatures at different points inside the calorimeter chambers. However, at lower air flow rates this error might be appreciable and should be corrected if required.

Figure 4.4: Estimated heat leakage through the calorimeter chambers at different heater input power levels and various air flow rates
The conducted heat leakage through each calorimeter chamber, $q_{\text{cond1}}$ and $q_{\text{cond2}}$, estimated at different heater input power levels and various air flow rates are shown in Figure 4.4. It can be seen that, for a known heater input power, the heat leakage through chamber 2 is relatively higher than that for chamber 1. The reason is the higher average temperature in chamber 2 due to the higher inlet air temperature. The conducted heat leakage from chamber 2 to chamber 1 was also estimated to be relatively small and negligible when compared to $q_{\text{cond1}}$ and $q_{\text{cond2}}$.

The variation of the conducted heat leakage in each chamber is almost linear with the corresponding heater input power. The ratio of the conducted heat leakage to the heater input power significantly decreases (from 7% to 1.5% for $q_{\text{cond2}}$) with increasing the air flow rate (from 30 L/s to 80 L/s). Although, the conducted heat leakage can be estimated with an accuracy of ±1 W, it is not appropriate to allow for a large amount of conducted heat leakage through the calorimeter walls by maintaining a low air flow rate. This issue will be discussed in conjunction with the maximum average temperature inside the calorimeter in Section 4.4.5.

### 4.4.3 Air temperature rise across each chamber

Under thermally steady state conditions, the thermopiles' thermoelectric voltages were collected at a sampling rate of 100 Hz for a duration of 100 seconds using the DA system. These readings were averaged over the given period and converted to temperature according to the calibration curves derived for each thermopile as previously discussed in Chapter 3. The averaged temperatures corresponding to each thermopile, $\Delta T_1$ and $\Delta T_2$, were then used for calculation of the dissipated heat by the test heater, $P_{\text{th}}$, based on Equation (4.8).
Using the experimental data, variation of the air temperature rise across chamber 1 calculated at different heater input power levels and different air flow rates are shown in Figure 4.5. It can be seen that, for a given air flow rate, $\Delta T_1$ increases almost linearly with increasing the heater input power. One of the reasons for any non-linearity seen could be the changes of the air specific heat, $c_p$, with air temperature which becomes appreciable at lower air flow rates (ie higher air temperatures). Therefore, in calculating the heat transfer using Equation (4.8), a correction factor due to the variation of the air specific heat with temperature can be taken into account if a high order of accuracy is required.

The resolution of the loss measurement using DCC can be determined based on the resolution of the air temperature measurement, $\Delta T_1$ and $\Delta T_2$, which is 0.1°C. For a given heat loss, the best resolution can be achieved at the lowest air flow rate where a relatively high temperature rise occurs across each chamber. According to the gradient of the curves shown in Figure 4.5 a resolution of 3 W can be achieved at the lowest air flow rate of 30 L/s. This figure increases to 6 W and 8 W when higher air flow rates of 55 L/s and 80 L/s are maintained through the calorimeter.
chambers respectively. However, the accuracy of the loss measurement should be separately investigated in conjunction with the other quantities required for the heat loss measurement as discussed in the next sub-section. A theoretical error analysis is also presented in Section 4.5.

4.4.4 Estimation of dissipated heat by the test heater

As a first approximation, the dissipated heat by the test heater, \( P_{th} \), is estimated according to Equation (4.8) with the assumption of unchanged specific heat in both chambers (ie \( c_{p1} = c_{p2} \)). The estimated values are then compared with the electrical input power supplied to the test heater which was measured by a wattmeter. A maximum absolute discrepancy of 40 W (11%) was observed between the estimated and measured values of \( P_{th} \) at an air flow rate of 30 L/s for heater input power levels in the range 200-500 W. The error reduces to 8% when a correction factor due to the change of air specific heat is applied. This figure is still high and implies that the majority of the generated heat cannot be taken away to the outside of the calorimeter. This issue is discussed in further detail in the next sub-section where the limits for the air flow rate are derived.

Using the experimental results, the absolute discrepancy between the estimated and measured losses in the test heater, error (W), is calculated as a function of air temperature rise across one chamber as shown in Figure 4.6. The air flow rate was constant at 55 L/s, the heater input power levels were in the range of 200-500 W and the air temperature rise, \( \Delta T \), was in the range of 3-9°C. The absolute error changes within a range of 7-14 W and increases slightly with temperature rise as demonstrated by the fitted line to the experimental data.
The percentage error, however, significantly decreases with the increase of the temperature rise across each chamber as demonstrated by the fitted line to the experimental data. A similar trend for the percentage error can be achieved if the absolute error is assumed to be constant, e.g., at an average value of 11 W. Although, the trend shown in the bottom graph of Figure 4.6 indicates a point at which the
percentage error could be zero, this cannot happen due to the limited precision of the employed equipment. Instead, a minimum percentage error of 2% can be seen from the graph.

Experimental test results at a higher air flow rate of 80 L/s also confirmed that the absolute error remains reasonably constant at about 12 W for the heater input power of 200-800 W where the corresponding percentage error is 6%-1.5% respectively. Further discussion is given in Section 4.5 where a theoretical error analysis is presented.

4.4.5 Deriving limits for the DCC

The experimental test results using two heaters in separate chambers can be used to determine limits for the air flow rate through the calorimeter, temperature distribution inside each chamber, air temperature rise across each chamber and heater input power levels. These limits are the important aspects which should be known prior to the application of the DCC for the measurement of the test induction motor losses.

The air temperature rise in each chamber is a function of the dissipated heat loss in each chamber as well as the air flow rate. As an example, for an air flow rate of 30 L/s and heater input power levels of 400 W, the average air temperature in chamber 2 was measured to be about 45°C some 24°C more than the ambient temperature. This results in a reasonable proportion of conducted heat leakage through the calorimeter walls. An average figure for such a temperature difference related to chamber 2 is about 25 W which is about 5% of the total dissipated heat in chamber 2.

More importantly, the low air flow rate causes a non-uniform temperature distribution inside the calorimeter chambers which results in a conducted heat leakage different from the estimated values. Meanwhile, it was demonstrated that the discrepancy between the estimated and measured heat loss (% error) becomes significant when a low air flow rate is maintained through the calorimeter chambers.
Therefore, it was concluded that the air flow rate of 30 L/s through the calorimeter is definitely not sufficient when a heat loss of more than 200 W is to be estimated.

Maintaining normal operating conditions for the test motor is another criterion which should be considered in determining air flow rate through the calorimeter. Depending on the motor heat loss, a sufficient air flow rate should be maintained through the calorimeter to prevent the motor from overheating. To start with, the average air flow rate established by the fan of the test motor was estimated to be in the range of 50-60 L/s. This was determined by measuring the air velocity profile across the motor case. This figure, however, can be interpreted as the minimum air flow rate through the calorimeter since it satisfies those conditions under which the test motor needs to be operated without being overheated. The higher limit for the air flow rate is determined according to the limits for temperature rise inside and across the two chambers as discussed hereafter in this section.

With a higher air flow rate through the calorimeter the average temperature reduces in calorimeter chambers. For instance, with an air flow rate of 55 L/s and the heater input power levels of 500 W, the average temperatures in chambers 1 and 2 were measured to be about 27°C and 34°C respectively, being about 6°C and 13°C more than the ambient temperature. This resulted in a lower conducted heat leakage and a more uniform temperature distribution in the two chambers.

According to Figures B.1 and B.2 (Appendix B), the variation of the air specific heat can be predicted with a reasonable accuracy and without requiring knowledge of the relative humidity and barometric pressure if the average temperature does not exceed 50°C. This condition can be achieved by limiting the average air temperature in chamber 2 to 50°C. In this case, the average air temperature in chamber 1 can be up to 40°C depending on the inlet air temperature and dissipated heat loss in chamber 1. A correction factor up to 4% due to the variation of specific heat within the two chambers can be applied for loss estimation using Equation (4.8). Error due to neglecting the effect of relative humidity and barometric pressure is considerably small and can be ignored.
The limits for the air temperature rise across each chamber, $\Delta T_1$ and $\Delta T_2$, can be determined using the limits for the air flow rate and the maximum average temperature in the calorimeter chambers which are proportional to the heat loss to be estimated. According to Figure 4.5 and with an air flow rate of 55 L/s a maximum heat loss of 800 W can cause a temperature rise of more than 12°C across chamber 1. Similarly, the same air temperature rise will occur across chamber 2. Assuming an inlet air temperature of 25°C, the average temperature in chamber 2 exceeds the upper limit of 50°C. At a higher air flow rate of 80 L/s a maximum air temperature rise of 10°C across chamber 1 was achieved due to a heat loss of 800 W as shown in Figure 4.5. This figure was found to be adequate as the upper limit for $\Delta T_1$. A similar limit may be applied to the air temperature rise across chamber, $\Delta T_2$, provided that the average temperature in chamber 2 remains under 50°C.

With an air flow rate of 80 L/s and assuming a minimum heat loss of 200 W, a minimum temperature rise of 2°C can be measured across each chamber. Considering the resolution of the temperature measurement system (thermopiles in conjunction with the DA system), this figure is found to be adequate as the lower limit for the air temperature rise across each chamber. However, in this case, the heat loss measurement could be subject to an error greater than 6% as shown in Figure 4.6. Therefore, care should be taken to use the maximum possible air temperature rise across the calorimeter chambers to obtain the highest possible accuracy without violating the specified limits.

According to the experimental results, an air flow rate of 80 L/s was found to be suitable for conducting loaded machine tests with the confidence that the operational conditions of the test motor remain within the specified limits. However, in the case that measurement of a higher heat loss is required or the average temperature in chamber 2 exceeds 50°C, the air flow rate can be increased up to 100 L/s.

In the single chamber type calorimeter approach [Tur91], the heater input power should be of the same order as the induction motor losses. However, in the DCC
approach, satisfying this condition is not essential since the heat loss estimation is performed based on the ratio of the air temperature rise across each chamber. Therefore, it is possible to supply the reference heater with an input power different from the level of motor losses. This argument is justified by referring to the experimental tests using two heaters. As an example, the test heater input power of 500 W was estimated while an input power of 200 W was applied to the reference heater. The maximum discrepancy between the estimated heat loss using Equation (4.8) and the losses measured by the wattmeter was less than 15 W. This figure is in good agreement with similar test results conducted with both test and reference heater power levels of the same order.

Investigations on the test results confirmed that the reference heater input power could be adjusted to the same level as the dissipated heat in chamber 1 (practically equal to the motor losses) up to 500 W. When the dissipated heat in chamber 1 is more than 500 W, as in the case of the loaded machine losses, the reference heater input power can be kept constant at 500 W. This limit also ensures that the maximum temperature rise inside and across chamber 2 stays within the specified limits. A lower limit of 200 W for the reference heater input power was found to be appropriate with regard to the resolution of the heat transfer measurement.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Lower Limit</th>
<th>Upper Limit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air flow rate</td>
<td>55 L/s</td>
<td>100 L/s</td>
</tr>
<tr>
<td>$T_{avg2}$*</td>
<td>-</td>
<td>50°C</td>
</tr>
<tr>
<td>$\Delta T_1$</td>
<td>2°C</td>
<td>10°C</td>
</tr>
<tr>
<td>$\Delta T_2$</td>
<td>2°C</td>
<td>10°C**</td>
</tr>
<tr>
<td>$P_{rh}$</td>
<td>200 W</td>
<td>500 W</td>
</tr>
</tbody>
</table>

* Average air temperature in chamber 2
** Provided $T_{avg2} < 50^\circ C$

Table 4.2: Limits for different parameters derived for the DCC
The limits for different parameters of the DCC described in this section are summarised as shown in Table 4.2.

4.5 Error Analysis and Accuracy of the Loss Measurement

Dissipated losses, $q$, due to the operation of any heat source in the first chamber of the DCC can be estimated using Equation (4.8) as:

$$q = \left( P_{rh} - q_{\text{cond}2} \right) \left( \frac{c_{p1}}{c_{p2}} \frac{\Delta T_1}{\Delta T_2} \right) + q_{\text{cond}1} \quad (4.9)$$

where all the terms on the right hand side are already defined. The error analysis should be performed by investigating the accuracy of the different terms of Equation (4.9).

Since the conducted heat leakage through the calorimeter chambers, $q_{\text{cond}1}$ and $q_{\text{cond}2}$, are a small proportion of the total dissipated heat within the DCC, their accuracy has a negligible effect on the overall accuracy of the loss measurement. Also, as demonstrated in Chapter 3, specific heat is subject to 4% increase when air temperature changes from 20-50°C as in this application. Therefore, the relative error due to the estimation of specific heat in the two chambers, $c_{p1}$ and $c_{p2}$, would be a small fraction of 4% and can be neglected. Accordingly, Equation (4.9) can be written in a simplified form of:

$$q = P_{rh} \left( \frac{\Delta T_1}{\Delta T_2} \right) \quad (4.10)$$

where the accuracy of the loss measurement is directly proportional to the accuracy of the reference heater power and temperature measurements.

Assuming $\Delta T_1 = \Delta T_2 = \Delta T$, the corresponding error can be calculated as:

$$\frac{\delta_q}{q} = \frac{\delta_{P_{rh}}}{P_{rh}} + \frac{2\delta_{\Delta T}}{\Delta T} \quad (4.11)$$
where $\delta$ represents the absolute probable error due to the measurement of the related quantities. The relative accuracy of the single-phase wattmeter used for measuring $P_{th}$ is 2% as given in its specifications. The same order of accuracy is achieved if the measurement of the heater input voltage and current is considered using the DA system. As described in Chapter 3, air temperature difference across each chamber, $\Delta T$, can be measured with a maximum absolute uncertainty of 0.1°C. This figure is assumed to be constant over the specified range of 2-10°C for $\Delta T$ giving the relative accuracy of 5% to 1% respectively for the temperature measurement across each chamber.

Using Equation (4.11), the total relative error of the loss measurement, $\delta q/q$, will be in the range 12% to 4% depending on the level of heat loss to be estimated. The larger figure is applicable at low power estimation, ie 200 W, where $\Delta T \approx 2 ^\circ C$. In higher loss estimation of 1 kW, ie when $\Delta T \approx 10 ^\circ C$, the loss measurement is subject to an absolute error of 40 W giving an accuracy of 4%. Obviously, an intermediate figure needs to be worked out when $\Delta T$ lies between the specified limits. It should be noted that the absolute error may reduce if a lower air flow is applied but the relative error stays almost the same. Therefore, it can be said that the best accuracy can be achieved when the maximum temperature rise of 10 °C is established.

Experimentally, however, an average accuracy of 2% was achieved for the measurement of heat loss in the range 200-500 W with an air flow rate of 55 L/s as discussed in Section 4.4.4. This figure is within the specified range of accuracy described above.

### 4.6 Conclusions

Thermal behaviour of the double chamber calorimeter (DCC) designed for estimation of losses of a 7.5 kW cage induction motor has been investigated and verified in this chapter. The one-dimensional conducted heat leakage through the calorimeter insulation material has been measured by performing some basic thermal tests. The heat leakage has been theoretically evaluated by developing a
simple heat leakage model as a function of both absolute temperature and temperature difference related to the DCC. The significance of the heat leakage through the calorimeter edges has also been highlighted by incorporating a conduction shape factor into the model. It has been shown that the conducted heat leakage through the whole calorimeter can be estimated with an accuracy of ±1 W.

Performance, reliability and accuracy of the DCC have been evaluated by performing experimental tests using a known heater (a test heater similar to the reference heater) instead of the test motor. The reason was to have more control on the dissipated heat loss and hence on the operation of the DCC. Experimental tests have been performed under various air flow rates through the calorimeter and with different input power levels supplied to both test and reference heaters. The test results have been carefully investigated and several important outcomes such as the limits for the air flow rate, maximum temperature rise inside and across each chamber and reference heater input power levels have been derived. The importance of deriving these limits have been highlighted and their influences on the heat loss estimation have been discussed.

The experimental results confirmed that an air flow rate of 55-100 L/s is required through the calorimeter in order to measure a heat loss in the range 200 W to 1 kW. The lower limit ensures sufficient air convection around the test motor to prevent it from overheating. The temperature rise across each chamber has been limited within 2-10°C. It has been demonstrated that the heat loss can be estimated with a precision as high as 10 W. Also a machine loss of up to 1 kW can be estimated using the DCC with a maximum uncertainty of 4%. The DCC approach allowed for the same order of accuracy as with the single chamber type calorimeter [Tur91] while a significant simplicity is achieved in calorimeter construction, control and measurement system.
Chapter 5

Induction Motor Harmonic Tests

5.1 Introduction

The accurate measurement of induction motor harmonic losses has been a difficult task due to the limited frequency bandwidth of standard measuring equipment. Another problem is the inaccuracy involved in the measurement of motor output power. The development of the double chamber calorimeter (DCC) made it possible to estimate total machine losses, including harmonic losses accurately regardless of the supply voltage conditions. The approach is quite useful, even when the machine is loaded, since the motor losses can be estimated directly under any desired loading conditions.

The difficulty in providing a flexible and controllable source of harmonics has resulted in a limited access to experimental data for harmonic analysis of induction motors. However, with the aid of the harmonic generator [Gos93], it is possible to examine the effects of different time harmonics on performance of the test induction motor. Using the DCC, the corresponding additional losses can be estimated accurately under various distorted supply conditions as well as different loading levels. In general, the experimental data will be utilised to investigate the harmonic behaviour of the test motor.

In this chapter, the specifications of the test motor are given and the equivalent circuit parameters are derived based on the standard test results. The influencing parameters on the variation of motor losses are described. The methodology of performing experimental harmonic tests on the test motor is discussed and the estimation of motor losses using the DCC is described. The total estimated machine losses along with the segregated harmonic losses corresponding to different tests are tabulated and presented in this chapter.
5.2 Test Induction Motor

The test machine is a high efficiency 3-phase cage induction motor with the specifications shown in Table 5.1. A copy of the motor's manufacturer data sheet is given in Appendix D.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power rating</td>
<td>7.5 kW</td>
</tr>
<tr>
<td>Nominal voltage/frequency</td>
<td>415 V, 50 Hz</td>
</tr>
<tr>
<td>Connection</td>
<td>Δ</td>
</tr>
<tr>
<td>Full load current</td>
<td>14.5 A</td>
</tr>
<tr>
<td>Full load speed</td>
<td>1440 rpm</td>
</tr>
<tr>
<td>Full load efficiency</td>
<td>0.88</td>
</tr>
<tr>
<td>Full load power factor</td>
<td>0.82</td>
</tr>
</tbody>
</table>

Table 5.1: Specifications of the test induction motor

5.2.1 Initial tests

In order to calculate the motor parameters, several tests were conducted on the unloaded test motor supplied by mains via a variac. The motor line-to-line voltages, line currents and input power were measured using a 3-phase digital AC meter. In the early stages it was realised that there are some variations of the order ±15 W in the measured values of motor input power at no-load from time to time. There are several reasons for this as discussed below:

Imperfect supply: Based on the measurements, it was noticed that the mains voltage in the laboratory, where all the experiments were conducted, is distorted, unbalanced and subject to some random fluctuations. A typical line-to-line supply voltage waveform is captured as illustrated in Figure 5.1. It is not pure sinusoidal and contains low order harmonics, mainly 3rd (1%) and 5th (1.6%), with a voltage THD of 2.1%. A flat top waveform having similar distortion levels was observed in the line-neutral supply voltage.
Depending on the time of the day variations of the order ±10 V rms was observed in the line-to-line supply voltages. Also a discrepancy of up to 1% was detected between the rms values of the line-to-line supply voltages indicating that the supply is unbalanced. Most of the time, all of these imperfections existed in the supply voltage to some extent and caused variations in the motor input power.

**Temperature effect:** The unloaded machine is required to operate for several hours in order to allow the temperature to be distributed and settled in the different parts of the machine. Temperature changes, in turn, will affect the motor parameters, especially the stator winding resistance and hence the corresponding losses. The ambient temperature is also another influencing parameter which might indirectly cause variations in motor losses from time to time.

![Un-scaled supply line-to-line voltage waveform](image)

*Figure 5.1: Un-scaled supply line-to-line voltage waveform*

**Low power factor:** Since unloaded induction motors operate with a small power factor, about 0.1 for the test motor, the phase shift between the motor voltage and current is close to 90°. In such a situation, any small error in the estimation of the phase angle results in a significant change in the calculation of the active power.

The 3-phase digital AC meter employed in this application is capable of working in a low power factor situation of 0.2. According to the specifications of the meter, as
shown in Appendix C, the voltage, current and active power at nominal frequency and unity power factor can be measured with an accuracy of 1%. However, the error in active power measurements using the given meter could increase up to 3% at low power factors as in the case of unloaded induction motors.

**Solutions:** Different approaches were considered to reduce the above effects on the variation of motor input power measurement. In order to minimise the effect of temperature changes, the test motor was allowed to run for at least 3 hours under nominal voltage to reach a thermally stable condition. Adjustment of the variac greatly reduced the supply voltage fluctuations almost to zero. The output line-to-line voltages were monitored continuously and the variac was manually adjusted to the nominal value where necessary. However, there was no simple solution to overcome the distortion and imbalance in the supply voltage. Alternatively, the HG was employed to perform experimental tests on the motor as discussed in Sections 5.3 and 5.4.

### 5.2.2 Motor Parameters

For various tests conducted using the variac, an average no-load input power of 400 W was obtained for the test motor. The variac was also used to conduct several locked rotor tests on the machine. Based on the no-load and locked rotor test results and the manufacturer's data sheet, the motor parameters corresponding to the single phase approximate equivalent circuit shown in Figure 5.2 are derived as:

\[
\begin{align*}
S_{\text{base}} &= 3 \times V_{\text{base}} \times I_{\text{base}} = 3 \times 240 \times 14.5 = 10440 \text{ VA} \\
Z_{\text{base}} &= V_{\text{base}} / I_{\text{base}} = 240 / 14.5 = 16.55 \Omega \\
R_1 &= 0.47 \text{ to } 0.70 \Omega = 0.028 \text{ to } 0.042 \text{ pu depending on the operating temperature} \\
R_2 &= 0.72 \Omega \text{ at low slips (using manufacturer's data sheet)} = 0.044 \text{ pu} \\
R_2 &= 1.36 \Omega \text{ (using locked rotor test results at } 50 \text{ Hz)} = 0.082 \text{ pu} \\
X &= X_1 + X_2 = 3.05 \Omega \text{ (using locked rotor test results)} = 0.18 \text{ pu} \\
R_c &= 580 \Omega \text{ (using no-load test results)} = 35.0 \text{ pu} \\
X_m &= 41 \Omega \text{ (using no-load test results)} = 2.5 \text{ pu}
\end{align*}
\]
where $R_2$ and $X_2$ are the rotor parameters referred to the stator side.

![Approximate single phase equivalent circuit for the test induction motor](image)

**Figure 5.2:** Approximate single phase equivalent circuit for the test induction motor

The stator winding DC resistance was also measured at room temperature and immediately after some of the no-load and loaded tests where an increase of about 40% was observed.

### 5.3 Harmonic Generator (HG)

The HG is a 3-phase 10 kVA programmable inverter with the capability of producing harmonically distorted voltages. It can produce individual or a combination of different order harmonics, up to 1 kHz (20th) as well as the fundamental frequency. The magnitude of the individual harmonics can be as high as 30% of the fundamental component provided that the maximum voltage THD does not exceed the upper limit of 40%. These figures have been selected large enough to cover the maximum allowable distortion level specified for the test motor.

In addition to the magnitude, the phase shift of each harmonic with respect to the fundamental component can also be specified. This may particularly change the peak value of the produced voltage and can be applied to investigate the corresponding effect on the equipment under test. A brief description of the HG operation is given in Appendix A.

In order to eliminate the switching frequency from the HG output voltage, a second order low pass LC power filter is connected to the output terminals of the HG. As shown in Figure 5.3, the series inductors are 750 µH per phase and the delta
connected capacitors are 2.2 μF each. The resonance frequency of the filter is about 2.3 kHz.

![Diagram of LC power filter](image)

Figure 5.3: LC power filter connected at the output terminals of the HG

Experimental results showed that the filter satisfactorily eliminates the unwanted high frequencies giving a reasonably pure sinewave at its output. For instance, a typical fundamental voltage waveform produced by the HG and measured at the output of the filter is shown in Figure 5.4. The voltage THD and the most significant harmonic components are given as: THD = 3.2%, 5th (1.2%) and 50th (0.9%). The rest of the harmonic components have a distortion level less than 0.5% and hence are neglected. It can be seen that the level of distortion is almost the same as that corresponding to the mains voltage as discussed in Section 5.2.1. The presence of high frequency component (50th harmonic) can be justified since it is close to the resonance frequency of the filter.

![Un-scaled fundamental voltage waveform](image)

Figure 5.4: Un-scaled fundamental voltage waveform produced by the HG and measured at the output of the filter
Experimental results confirmed that the HG output voltage measured at the output of the filter is affected by variations of the 3-phase voltage supplied to the HG input. Therefore, it was decided to use the variac to manually adjust the input voltage of the HG to a desired value. More importantly, the level of unbalance in the HG output voltage was negligible as compared with the unbalance that exist in the mains voltage. These issues along with the effect of inverter DC link voltage level and effect of filter on the output voltage are discussed in Appendix A.

In terms of distorted waveforms, the voltage measured at the output of the filter was in a reasonable agreement with the requested data. For instance, the HG was setup to produce a distorted waveform containing 10% of the 11th harmonic. The voltage waveform captured at the output of the filter is shown in Figure 5.5. The spectrum analysis confirmed that the given waveform contains 7.6%, (rather than 10% requested), of 11th, 1.1% of the 5th and a THD of 8.1%. Some of the possible reasons for the described discrepancies are discussed in Appendix A in relation to the HG operation. Further investigations confirmed that these mismatches are unimportant as long as the produced voltages can be measured with sufficient accuracy.

Figure 5.5: Un-scaled voltage waveform containing 10% of 11th harmonic produced by the HG and measured at the output of the filter
5.4 Induction Motor Calorimetric Tests

5.4.1 Verification of the DCC

Investigations on the heat loss measurement procedure using the DCC has been presented in Chapter 4. In order to further verify the DCC operation several calorimetric tests were performed on the unloaded test induction motor supplied by mains via a variac. The motor voltage, current and input power were continuously monitored using the 3-phase digital AC meter and the DA system. Using a variac, the reference heater input power was adjusted to 400 W (of the same order as the motor no-load losses) and measured using the single-phase wattmeter and the DA system. The DCC was sealed and the fan speed was adjusted to force the air through the calorimeter with a constant flow rate of around 75 L/s. During the course of these tests, any possible fluctuations in the motor voltage and heater input power were compensated manually using the variacs.

The tests were continued long enough to make sure that the thermal equilibrium for the test motor, the reference heater and the DCC itself has been achieved. This was checked by monitoring the air temperature rise inside and across the calorimeter chambers. The steady state condition was assumed to be achieved when the air temperature difference across each chamber was constant at a certain value ±0.1°C for the last hour of the test. According to the experimental results the 3-hour figure already derived as the settling time for the test induction motor was found to be sufficient for steady state thermal establishment of the whole calorimetric setup as well.

Under steady state conditions all the relevant measurements were recorded for the appropriate duration for the last hour of the test. The collected data were then used for estimation of motor losses according to the procedure described in Chapter 3. Among the measured quantities, the heater input power and the temperature rise across each chamber (measured by thermopiles) have a critical role in the estimation of motor losses. For comparison, the electrical motor input power measured by the
digital AC meter was also recorded carefully. The motor voltage and current data had no direct role in the estimation of losses using the DCC, however, they were also recorded separately to estimate the motor impedance under different conditions. The repeatability of the DCC approach was checked by performing calorimetric tests on the test motor at different times/days. In some cases, different air flow rates within the limits specified for the DCC were maintained through the calorimeter to recheck the calorimeter performance and accuracy in measuring motor losses.

For all of the tests, the calculated motor losses using the DCC were compared with the electrical motor input power measured by the digital meter. In the worse case, a maximum discrepancy of 15 W (< 4%) and an average difference of ±5 W was observed between the two. This amount of error is quite acceptable since it is well within the range of accuracy specified for the DCC and AC meter. The experimental results demonstrated sufficient accuracy and gave enough confidence to utilise the DCC for estimation of motor harmonic losses under different supply conditions.

5.4.2 Experimental tests using the harmonic generator

The first series of main calorimetric tests was conducted on the unloaded test motor supplied by the harmonic generator (HG) at different supply conditions. The tests were planned to cover nominal supply (ie 415 V, 50 Hz) as well as the harmonically distorted voltages. For harmonic tests, a combination of fundamental nominal voltage and one of the non-triplen odd harmonics up to 1 kHz (ie n = 5, 7, 11, 13, 17 and 19) were applied to the test motor separately. The magnitude of the individual harmonic voltages were chosen to be 10% of the fundamental voltage with no phase shift. The reason for choosing the given harmonics was that they are the most common and harmful harmonics that could be present in the mains. Even order harmonics are normally cancelled out due to the symmetrical supply waveforms. Balanced triplen harmonics cannot be produced by the HG.
The tests were started by applying the nominal voltage again for three hours to achieve steady state conditions. The corresponding measurements were taken and recorded for further analysis. According to the theoretical analysis, with the same distortion level, the low order harmonics cause more losses in the motor when compared to the higher order ones [Jal94]. Therefore, it was decided to continue the harmonic tests in the presence of the highest order harmonic (ie 19th) followed by the lower order ones. This procedure was found useful since the tests were conducted in a way that the machine temperature increases in a natural way from the lowest value (at nominal supply) to the highest one (at nominal supply plus 10% of 5th harmonic). Also, during the consecutive harmonic tests, thermal equilibrium was achieved in a shorter time, practically about one hour, since the temperature distribution was already established in different parts of the machine.

Total machine losses under fundamental and distorted supply conditions, $W_{t1}$ and $W_{tn}$ respectively, were measured using the DCC approach as shown in Table 5.2. A comparison between the values of these quantities shows that harmonic losses, $W_n$, contribute a relatively small part of the total losses. Therefore, care should be taken in order to separate them from the total losses with sufficient accuracy. The original intention was to calculate the harmonic losses as:

$$W_n = W_{tn} - W_{t1}$$

(5.1)

corresponding to each test. This simple approach is quite acceptable if the losses due to the fundamental component of the applied distorted voltages remain the same and equal to $W_{t1}$ in all of the consecutive harmonic tests. This assumption is valid due to the fact that all the harmonic tests were performed while a nominal value of 415 V was given as the desired magnitude for the fundamental component of the produced waveform.

However, as mentioned before, there was still a possibility of alteration in the fundamental losses from time to time due to the various reasons such as motor operating conditions. Therefore, it was decided to perform at least two tests under
nominal conditions one at the beginning and one at the end of each series of harmonic tests. The estimated losses using these two tests were then averaged and considered as \( W_{t1} \) to be used in Equation (5.1). Experimental results confirmed that the discrepancies are often less than the resolution of the loss measurement and can be disregarded. Conducting each series of tests in one day was also beneficial since the environmental conditions remained reasonably unchanged allowing the same operational conditions for all the tests.

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>Total losses, ( W_{tn} ) (W)</th>
<th>Harmonic losses, ( W_n = W_{tn} - W_{t1} ) (W)</th>
<th>( V_n ) (pu)</th>
<th>( I_n ) (pu)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>400</td>
<td>30</td>
<td>0.074</td>
<td>0.148</td>
</tr>
<tr>
<td>7</td>
<td>385</td>
<td>15</td>
<td>0.078</td>
<td>0.103</td>
</tr>
<tr>
<td>11</td>
<td>380</td>
<td>10</td>
<td>0.073</td>
<td>0.072</td>
</tr>
<tr>
<td>13</td>
<td>375</td>
<td>5</td>
<td>0.069</td>
<td>0.058</td>
</tr>
<tr>
<td>17</td>
<td>375</td>
<td>5</td>
<td>0.064</td>
<td>0.050</td>
</tr>
<tr>
<td>19</td>
<td>375</td>
<td>5</td>
<td>0.064</td>
<td>0.047</td>
</tr>
</tbody>
</table>

Table 5.2: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{t1} = 370 \) W

Harmonic losses related to each test, \( W_n \), were then calculated according to Equation (5.1) as shown in Table 5.2. It can be seen that, with almost the same distortion level, harmonic losses due to the lower order harmonics are higher than losses due to the higher order harmonics. For instance, with presence of about 7% of 5th harmonic voltage a harmonic loss of 30 W was estimated which is about 8% of the no-load and about 3% of full load losses. On the other hand, no appreciable harmonic loss can be estimated due to the presence of about 7% of 11th, 13th, 17th and 19th order harmonic voltages in this particular induction motor.

The motor line-to-line voltages and two of the line currents were also captured using the voltage and current measurement circuits and the DA system with a sampling
rate of 6.4 kHz. This will provide 128 points for one complete cycle of the voltage and current. Figure 5.6 illustrates the un-scaled waveforms for the motor voltage and current containing primarily the 5th harmonic. Using MATLAB spectrum analysis, the pu values for harmonic voltages and currents calculated for each test are shown in Table 5.2. The error due to the inaccuracy of the actual sampling frequency and the leakage effect [Gir82] is estimated to be up to 1%. This figure should be added to the voltage and current measurement error (1% as described in Section 3.5.5.2) giving an overall uncertainty of 2% in the estimation of harmonic voltages and currents. The corresponding absolute error could be as high as 0.001 pu in the case that a harmonic component is calculated to be 0.05 pu (ie 5% of the fundamental component).

It can be seen that the maximum harmonic current occurs due to the presence of the lowest order harmonic (ie 5th) in the supply voltage of the test motor. The calculation shows that the 5th harmonic current is about 15% of the motor rated current and more than 40% of the no-load input current. Further analysis of the test results are given in Chapter 6.

![Figure 5.6: Un-scaled motor voltage and current waveforms containing the 5th harmonic](image)
The harmonic tests were repeated by applying a higher level of voltage distortion (eg 20%) to the test motor. Since the harmonic losses due to the pair harmonics (eg 17th and 19th) were very close together, only 3 harmonics (ie 5th, 11th and 17th) were applied this time. Similarly, the tests were started and finished by applying the nominal voltage/frequency where an average value of $W_{t1} = 385$ W was estimated due to the fundamental supply voltage of 415 V. Harmonic tests were conducted in the same manner described before and total losses, $W_{tn}$, corresponding to each test were estimated using the DCC. Similarly, total estimated machine losses were utilised to calculate the harmonic losses for each test according to Equation (5.1). Total and harmonic losses corresponding to different tests along with the calculated pu harmonic voltages and currents are shown in Table 5.3.

According to the given data, a maximum harmonic loss of 80 W occurred due to the presence of 5th harmonic voltage having a 16% distortion level. In this case, a 5th harmonic current, equivalent to 30% of the motor rated current, was drawn from the supply (HG). This figure reduces to about 10% in the presence of about 14% of the 17th harmonic voltage while no significant increase can be seen in the machine losses. Further analysis of the given results are presented in Chapter 6.

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>Total losses, $W_{tn}$ (W)</th>
<th>Harmonic losses, $W_n = W_{tn} - W_{t1}$ (W)</th>
<th>$V_n$ (pu)</th>
<th>$I_n$ (pu)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>465</td>
<td>80</td>
<td>0.162</td>
<td>0.302</td>
</tr>
<tr>
<td>11</td>
<td>415</td>
<td>30</td>
<td>0.153</td>
<td>0.147</td>
</tr>
<tr>
<td>17</td>
<td>400</td>
<td>15</td>
<td>0.138</td>
<td>0.105</td>
</tr>
</tbody>
</table>

Table 5.3: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, $W_{t1} = 385$ W

5.4.3 Loaded machine test results

In order to investigate the effect of distorted supply voltage on the test induction motor under different load conditions two series of main tests were performed one
under half load and one under full load. The test motor was coupled to a DC generator which was supplying a load bank. The input power supplied to the load bank was calculated by measuring its DC voltage and current. The variation of the motor load level was done by adjusting the DC generator field current.

The first series of loaded machine tests was conducted at half of the full load and under different supply conditions. The DC generator output power was kept constant at about 3 kW during all tests in this series. The motor speed was also monitored and recorded to be constant at 1476±2 rpm (s = 0.016) during the tests. Similar to the no-load situation, the tests were started with the fundamental voltage/frequency using the HG long enough to achieve thermal stability. The corresponding average motor losses, $W_{tl}$, estimated using the DCC approach was 500W.

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>Total losses, $W_{tn}$ (W)</th>
<th>Harmonic losses, $W_{tn} - W_{tl}$ (W)</th>
<th>$V_n$ (pu)</th>
<th>$I_n$ (pu)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>650</td>
<td>150</td>
<td>0.162</td>
<td>0.289</td>
</tr>
<tr>
<td>7</td>
<td>585</td>
<td>85</td>
<td>0.171</td>
<td>0.214</td>
</tr>
<tr>
<td>11</td>
<td>560</td>
<td>60</td>
<td>0.152</td>
<td>0.144</td>
</tr>
<tr>
<td>13</td>
<td>545</td>
<td>45</td>
<td>0.149</td>
<td>0.130</td>
</tr>
<tr>
<td>17</td>
<td>540</td>
<td>40</td>
<td>0.133</td>
<td>0.105</td>
</tr>
<tr>
<td>19</td>
<td>535</td>
<td>35</td>
<td>0.131</td>
<td>0.094</td>
</tr>
</tbody>
</table>

Table 5.4: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, $W_{tl} = 500$ W at half load conditions.

Thereafter, distorted voltages containing non-triplen odd harmonics with 20% distortion level were applied to the motor. Under thermally steady state conditions, the relevant data were recorded and used to calculate the total machine losses corresponding to different supply conditions. The estimated total and harmonic losses along with the calculated values for pu harmonic voltages and currents are
shown in Table 5.4. It can be seen that both fundamental and harmonic losses are increased as compared with the no-load test result. However, the harmonic currents remained almost the same as in the no-load tests having a decreasing trend with harmonic order. Further analysis of the results is presented in Chapter 6.

The second series of loaded machine tests was conducted on full load machine supplied by distorted waveforms. The output power of the DC generator was constant at about 6 kW. The motor speed was also constant at 1440±2 rpm during all the tests in this series. Machine losses under fundamental supply conditions was estimated using the DCC to be 1015 W. Total losses under different supply conditions along with the pu values for harmonic voltages and currents are shown in Table 5.5. It can be seen that the difference between the total and fundamental losses increases as low order harmonics is applied.

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>Total losses, ( W_{tn} ) (W)</th>
<th>Harmonic losses, ( W_{tn} - W_{tl} ) (W)</th>
<th>( V_n ) (pu)</th>
<th>( I_n ) (pu)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>1225</td>
<td>210</td>
<td>0.167</td>
<td>0.325</td>
</tr>
<tr>
<td>7</td>
<td>1160</td>
<td>145</td>
<td>0.164</td>
<td>0.244</td>
</tr>
<tr>
<td>13</td>
<td>1100</td>
<td>85</td>
<td>0.150</td>
<td>0.145</td>
</tr>
<tr>
<td>19</td>
<td>1055</td>
<td>40</td>
<td>0.142</td>
<td>0.106</td>
</tr>
</tbody>
</table>

Table 5.5: Motor losses and pu harmonic voltages and currents under different distorted supply conditions, \( W_{tl} = 1015 \) W at full load conditions

A comparison between the results given in Tables 5.2 and 5.5 indicates that, for a given pu voltage distortion due to a particular harmonic order, the additional losses in the motor increases with load. For instance, additional losses due to the presence of about 16% of the 5th harmonic voltage at full load is about 40% higher than that in half load. In terms of harmonic currents, no significant difference is observed between the no-load, half load and full load results. Further analysis of the given test results will be presented in Chapter 6.
5.4.4 Separation of fundamental losses

Using the estimated fundamental losses, $W_{t1}$, different components of the machine losses calculated for different loading conditions are shown in Table 5.6. As expected, both stator and rotor losses are increased with load while core losses and windage and friction losses are almost constant. Stray load losses are contributed a small portion of the total fundamental losses.

<table>
<thead>
<tr>
<th>Losses (W)</th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_{t1}$ (from tests)</td>
<td>370</td>
<td>385</td>
<td>500</td>
<td>1015</td>
</tr>
<tr>
<td>$W_1 = 3R_1I_1^2$</td>
<td>46</td>
<td>48</td>
<td>115</td>
<td>383</td>
</tr>
<tr>
<td>$W_c = 0.9V^2/R_c$</td>
<td>261</td>
<td>262</td>
<td>260</td>
<td>267</td>
</tr>
<tr>
<td>$W_{fw} =$ constant</td>
<td>50</td>
<td>50</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>$W_{II} = 0.005 P_{rated} (I_1/I_{rated})^2$</td>
<td>5</td>
<td>5</td>
<td>11</td>
<td>33</td>
</tr>
<tr>
<td>$W_2 = W_{t1} - (W_1 + W_c + W_{fw} + W_{II})$</td>
<td>8</td>
<td>20</td>
<td>64</td>
<td>283</td>
</tr>
</tbody>
</table>

Table 5.6: Separation of fundamental losses in the test motor under different loading conditions

5.5 Conclusions

The effects of the supply conditions on the variation of test motor no-load losses has been investigated in this chapter. It has been shown that the mains voltage is subject to imperfections such as voltage variation (±2%), unbalance (up to 1%) and is harmonically distorted (having a 2% THD). This has resulted in a variation of ±5% in the motor no-load losses when supplied by the mains. Motor temperature settling time and motor low power factor are other influencing parameters. Precautions have been suggested to overcome and/or minimise these problems.
The applicability of the DCC in measurement of motor losses has been verified and its calibration accuracy has been checked using a digital AC meter. It has been shown that the machine losses can be estimated with a maximum uncertainty of 4% using the DCC. Using the harmonic generator (HG), motor losses have been determined under various distorted supply conditions and at different loading levels. Preliminary analysis confirmed that the low order harmonics cause more pronounced losses in the motor when compared to the high order harmonics. Also, it has been demonstrated that, for a given harmonic frequency with a known voltage distortion level, additional machine losses increase with load. The flow of harmonic currents due to a particular distorted voltage is almost independent of the motor loading conditions. Further analysis of the test results will be given in Chapter 6.
Chapter 6

Induction Motor Harmonic Loss Models

6.1 Introduction

Typical frequency variation of stator and rotor resistance as well as the leakage reactance of induction motors have been presented in the literature and were discussed in Chapter 2. In some cases, harmonic loss models have been developed to estimate harmonic losses in induction motors as described in Chapter 2. Harmonic losses have been assumed to be almost independent of the motor loading, however, this has not been clearly justified.

Using the experimental results given in Chapter 5, the variation of motor parameters with harmonic order at different loading conditions will be examined and discussed in this chapter. Wherever applicable, these variations will be compared with the existing harmonic models. Also the variation of harmonic losses with the motor load will be investigated using the experimental results.

Most standards specify limits for THD as well as for individual harmonics in the utility power supply. The suitability of these criteria will be discussed in relation to induction motors operating under distorted supply conditions. Accordingly, a derating factor will be defined for induction motors to allow for extra heating due to the presence of distorted voltages.

6.2 Analysis of the Test Results

Harmonic losses due to the presence of different levels of harmonic voltages in the supply of the test induction motor were measured using the DCC approach as presented in Chapter 5. The pu values of motor harmonic voltages and currents were also calculated for the given tests as shown in Tables 5.2 to 5.5. Analysis of the test results are presented in the following subsections.
Most of the analytical calculations presented in this chapter was performed using Microsoft Excel. Curve fitting was done using Solver to work out the correlations between one or more parameter/s and a series of data based on a given function. The best estimate for the experimental data is achieved by forcing the parameter/s to give a minimum value for sum of the square of errors between the estimated and experimental data.

### 6.2.1 Variation of total machine resistance $R_n$

Using the experimental results given in Tables 5.2 to 5.5, the test motor effective series resistance, $R_n$, corresponding to each harmonic order can be calculated as:

$$R_n = \frac{W_n}{I_n^2}$$  \hspace{1cm} (6.1)

where $W_n$ and $I_n$ are pu total harmonic losses and harmonic currents respectively. The calculated values of $R_n$ corresponding to different tests are shown in Table 6.1.

<table>
<thead>
<tr>
<th>n</th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>0.15</td>
<td>0.084</td>
<td>0.17</td>
<td>0.22</td>
</tr>
<tr>
<td>7</td>
<td>0.18</td>
<td>-</td>
<td>0.18</td>
<td>0.27</td>
</tr>
<tr>
<td>11</td>
<td>0.28</td>
<td>0.13</td>
<td>0.28</td>
<td>-</td>
</tr>
<tr>
<td>13</td>
<td>0.29</td>
<td>-</td>
<td>0.25</td>
<td>0.45</td>
</tr>
<tr>
<td>17</td>
<td>0.38</td>
<td>0.13</td>
<td>0.34</td>
<td>-</td>
</tr>
<tr>
<td>19</td>
<td>0.43</td>
<td>-</td>
<td>0.38</td>
<td>0.40</td>
</tr>
</tbody>
</table>

Table 6.1: Calculated values for $R_n$ corresponding to different harmonic tests

A graphical presentation of total machine resistance vs harmonic order is also illustrated in Figure 6.1. It can be seen that, in most cases, the machine resistance increases with harmonic order. This is basically due to the increase of the effective resistance of the rotor bars caused by deep bar effect. However, the calculated data
is subject to measurement error, due to both $W_n$ and $I_n$, which will be discussed next.

The estimated values of $W_n$ are subject to an average absolute error of $\pm 5$ W to $\pm 15$ W (depending on the level of the loss measurement). Therefore, the calculated values of $R_n$ as given in Table 6.2 are subject to an error of 5\% to 100\% depending on the value of $W_n$ where the worst case error occurs in relation to the smaller values of $W_n$. The larger the $W_n$ more accurate is the calculated $R_n$ which is achieved in loaded machine tests and when low order harmonics are applied. For clarity, error bars corresponding to the calculated values of $R_n$ using half load test results with $W_n \pm 10$ W are shown in Figure 6.1. For a given harmonic order, similar error bars can be applied to the rest of the data. In other words, the calculated values of $R_n$ at lower order harmonics are more accurate as compared with $R_n$ at higher order harmonics.

![Figure 6.1: Calculated values of total machine resistance for different harmonic tests and under different loading conditions](image)

It is also evident from the graph that, for a given harmonic order, the total effective resistance increases with load. There is a good reason behind this, that is the
temperature effect. It is known that, at nominal voltage/frequency, the motor temperature increases with machine loading due to the increased fundamental losses (e.g., from 400 W at no-load to about 1 kW at full load). This, in turn, causes the stator and rotor effective resistances to increase resulting in a higher fundamental copper losses.

In the presence of harmonics, there will be additional harmonic losses due to the increased stator and rotor winding resistances as from no-load to full load conditions. Also, for a given loading condition, harmonic losses cause more heating in the machine, on top of the fundamental losses, which again could affect the resistances and hence variation of fundamental copper losses. This issue needs to be investigated carefully in order to segregate additional fundamental losses due to the machine temperature rise as a result of a distorted voltage.

In order to investigate the variation of machine total resistance with harmonic order as well as with load, the calculated data for $R_n$ are fitted to a curve of the form:

$$R_n = K_R n^a$$  \hspace{1cm} (6.2)

where $K_R$ represents total machine resistance at fundamental frequency which is only a function of the motor conductor temperature. Assuming a constant conductor temperature, exponent $a$ describes the variation of motor resistance, particularly rotor bars due to deep bar effect, with harmonic order.

Using Equation (6.2), four different curves were fitted to the experimental data to give the best estimate for $R_n$ as shown in Figure 6.2. The corresponding values for $K_R$ and the exponent $a$ are shown in the second and the third rows of Table 6.2. It can be seen that, apart from data for no-load test 1, $K_R$ increases with machine loading which can be attributed to the temperature effect. The exponent $a$, however, demonstrates an inconsistent trend for increase of $R_n$ with load.
Chapter 6: Induction Motor Harmonic Loss Models

Figure 6.2: Experimental data for total machine resistance $R_n$ and the best fitted curves using Equation (6.2)

Considering the upper and lower limits for the $R_n$ due to the measurement error, further curve fitting were conducted to obtain a range of acceptable values for $K_R$ and $a$ as shown in Table 6.2. It can be seen that both $K_R$ and $a$ could take a wide range of values to estimate $R_n$ within its accuracy limits. For instance, under half load conditions, $K_R$ could take any value between 0.045 and 0.074 with $a$ varying from 0.80 down to 0.43 as illustrated in Figure 6.3. Similar graphs can be given for the rest of the data using the limits for $K_R$ and $a$ as given in Table 6.2. Therefore, it can be concluded that the values of $K_R$ and $a$ as given in rows two and three of Table 6.2 are not the only values, but among the values, which can give a reasonable estimate for $R_n$.

<table>
<thead>
<tr>
<th>Best fitted values</th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_R$</td>
<td>0.074</td>
<td>0.047</td>
<td>0.055</td>
<td>0.117</td>
</tr>
<tr>
<td>$a$</td>
<td>0.33</td>
<td>0.38</td>
<td>0.65</td>
<td>0.45</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Range of acceptable values</th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_R$</td>
<td>0.039 to 0.098</td>
<td>0.028 to 0.071</td>
<td>0.045 to 0.074</td>
<td>0.101 to 0.145</td>
</tr>
<tr>
<td>$a$</td>
<td>0.80 to 0</td>
<td>0.73 to 0.12</td>
<td>0.80 to 0.43</td>
<td>0.56 to 0.30</td>
</tr>
</tbody>
</table>

Table 6.2: Calculated values of $K_R$ and $a$ to give the best estimate for $R_n$ in different tests using Equation (6.2)
The curve fitting was then repeated so that only one common value for exponent $a$ is incorporated to estimate $R_n$ under different loading conditions. In this case, the calculated values of $R_n$ will be presented by different curves which have the same trend (i.e., one value for $a$) but different starting points (i.e., different values for $K_R$ corresponding to each series of data). For simplicity only one curve was fitted to both series of no-load test data without introducing a great inaccuracy. The derived values were: $a = 0.5$ and $K_R = 0.044, 0.080$ and $0.104$ corresponding to no-load, half load and full load conditions respectively. The fitted curves along with the experimental data are shown in Figure 6.4.

A comparison between the estimated and experimental values of $R_n$ demonstrated an average discrepancy of 14% between the two which mostly reflects the large measurement error of no-load data. However, using the estimated $R_n$, the corrected values for $W_n$ were calculated and compared with the measured values of $W_n$ where a maximum difference of 17 W and an average discrepancy of 8 W was observed. These figures are in good agreement with the possible error of the loss measurement showing the suitability of the given curves to estimate $R_n$. 

Figure 6.3: Experimental values for $R_n$ at half load and the fitted curves according to the error bars
According to locked rotor test results, a value of $R = 0.102$ pu was calculated as the total machine resistance at the rated current and fundamental frequency, i.e., at $n = 1$. This is compared with $K_R=0.104$, corresponding to the fitted curve at full load (with $n = 1$), where a difference of about 2% is calculated. Since the difference is within the specified accuracy of the measurement, it can be said that the model is reasonably accurate. However, no experimental data was available to assess the values of $K_R$ at no-load and half load conditions.

An expression similar to Equation (6.2) has been given in [Cum86] with $a = 0.6$. For a given $K_R$ and using $a = 0.5$, the estimated values of $R_n$ at $n = 19$ could be subject to a maximum discrepancy of more than 25% as $a$ changes from 0.5 to 0.6.

### 6.2.2 Variation of total impedance $Z_n$

In order to investigate the variation of total machine impedance with harmonic order as well as with load, the ratio $Z_n=V_n/I_n$ is calculated for each test as shown in Table 6.3. These values are subject to 5% measurement error due to both $V_n$ and $I_n$ as described in Chapter 3.
Chapter 6: Induction Motor Harmonic Loss Models

Table 6.3: Calculated values for $Z_n(=V_n/I_n)$ corresponding to different tests

<table>
<thead>
<tr>
<th>n</th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>0.50</td>
<td>0.54</td>
<td>0.56</td>
<td>0.55</td>
</tr>
<tr>
<td>7</td>
<td>0.76</td>
<td>-</td>
<td>0.80</td>
<td>0.72</td>
</tr>
<tr>
<td>11</td>
<td>1.02</td>
<td>1.04</td>
<td>1.05</td>
<td>-</td>
</tr>
<tr>
<td>13</td>
<td>1.18</td>
<td>-</td>
<td>1.14</td>
<td>1.11</td>
</tr>
<tr>
<td>17</td>
<td>1.28</td>
<td>1.32</td>
<td>1.27</td>
<td>-</td>
</tr>
<tr>
<td>19</td>
<td>1.35</td>
<td>-</td>
<td>1.38</td>
<td>1.44</td>
</tr>
</tbody>
</table>

As illustrated in Figure 6.5, it can be seen that the total machine impedance increases non-linearly with harmonic order mostly due to the deep bar effect occurring in the rotor bars. The calculated values of $Z_n$ are fitted in a curve of the form:

$$Z_n = K_Z n^b$$

where $K_Z$ is a constant representing the total machine impedance under fundamental frequency, i.e. $n = 1$. The exponent $b$ describes the frequency variation of total machine impedance.

Curve fitting to the given experimental data individually demonstrated that $K_Z$ could take any value between 0.18 and 0.24 with $b$ varying from 0.7 to 0.6 for different loading conditions. The best curve which describes all the experimental data gives $K_Z = 0.21$ and $b = 0.65$ as plotted in Figure 6.5. An average error of 5% was observed between the estimated and experimental values for $Z_n$. It can be seen that the experimental data almost evenly distributed around the fitted curve indicating that the machine total impedance does not change with load. In other words, a given harmonic voltage, $V_n$, causes a fixed harmonic current, $I_n$, to flow through the machine regardless of the motor loading conditions. Further investigations are carried out by considering the variation of total leakage reactance with harmonic order and load as in the next subsection.
6.2.3 Variation of total leakage reactance $X_n$

In order to examine the variation of total leakage reactance with harmonic order, the calculated values of $Z_n$ and $R_n$ (uncorrected) are used to determine $X_n$ as:

$$X_n = \sqrt{Z_n^2 - R_n^2} \quad (6.4)$$

The calculated values of $X_n$ corresponding to each series of data are illustrated in Figure 6.6. Subsequently, different curves of the form:

$$X_n = K_X n^c \quad (6.5)$$

were fitted to the experimental data corresponding to each test. In Equation (6.5), $K_X$ represents the total leakage reactance at nominal frequency, i.e. at $n=1$, and exponent $c$ describes the variation of motor leakage reactance with harmonic order.

The values for $K_X$ and $c$ which gave the best estimate for the experimental data individually are given in Table 6.4. Another curve was fitted to give the best estimate for all data gave $K_X = 0.20$ and $c = 0.64$ as shown in the last column of Table 6.4. The corresponding fitted curve is shown by the solid line in Figure 6.6 where the calculated $X_n$ are evenly distributed around the fitted curve. This
indicates that total machine reactance is almost independent of the motor loading conditions and varies only with harmonic order. It should be mentioned that an average error of 7% was observed between the experimental data and the estimated values of $X_n$ using the fitted curve to all data.

<table>
<thead>
<tr>
<th></th>
<th>no-load test 1</th>
<th>no-load test 2</th>
<th>half load</th>
<th>full load</th>
<th>all data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_X$</td>
<td>0.21</td>
<td>0.18</td>
<td>0.23</td>
<td>0.18</td>
<td><strong>0.20</strong></td>
</tr>
<tr>
<td>$c$</td>
<td>0.63</td>
<td>0.71</td>
<td>0.54</td>
<td>0.68</td>
<td><strong>0.64</strong></td>
</tr>
</tbody>
</table>

Table 6.4: Calculated values of $K_X$ and exponent $c$ to give the best estimate for $X_n$ in different tests using Equation (6.5)

![Graph showing experimental and estimated values for $X_n$ at different tests](image)

Figure 6.6: Experimental and estimated values for $X_n$ at different tests

As given in Section 5.2.2, the machine nominal leakage reactance was calculated using locked rotor test results as $X = 0.18$ pu. This figure is in total agreement with the derived curve for $X_n$ at full load conditions where $K_X = 0.18$. However, a discrepancy of more than 10% was observed between the nominal leakage reactance and the estimated value of $K_X = 0.20$. For comparison, another curve was fitted to the experimental data including a value of $X = 0.18$ pu at $n = 1$ where $K_X = 0.20$ and
Chapter 6: Induction Motor Harmonic Loss Models

0.65 were derived. This indicates that the fitted curve is subject to an error of more than 10% when estimating X at n = 1.

The exponent \( c = 0.8 \) has been defined [Cum86] in for machines with a wide range of power ratings. Assuming X = 0.2 pu, a maximum discrepancy of more than 50% was observed in the calculation of \( X_n \) for n = 19 and \( c = 0.8 \).

6.2.4 Variation of leakage inductance \( L_n \)

As discussed in Chapter 2, the non-linear variation of the machine leakage reactance is due to the deep bar effect which leads to a reduction of the effective nominal leakage inductance with frequency. For the given test results, the calculated ratio \( L_n = X_n/n \) demonstrates a decreasing trend with harmonic order as shown in Figure 6.7. It order to investigate the variation of \( L_n \) with both harmonic order and motor loading condition, the given data is compared with the equation:

\[
L_n = K_L n^d
\]

(6.6)

where \( K_L \) represents the nominal leakage inductance at n = 1 and the exponent \( d \) is a negative number which describes the variation of \( L_n \) with harmonic order. By curve fitting to each series of data individually four different values were obtained for \( K_L \) (from 0.16 to 0.19) and exponent \( d \) (from -0.26 to -0.34). As illustrated by the solid line in Figure 6.7, the best curve describing all data gives \( K_L = 0.17 \) and \( d = -0.28 \) and an average error of 6% with the experimental data.

Another curve was fitted by including the nominal value of \( L = 0.18 \) pu at n = 1 to the experimental data. The resultant values were \( K_L = 0.18 \) and \( d = -0.32 \) giving no error at n=1 and an average error of 5% as compared with the experimental values. This curve then can be considered as a good model describing the frequency variation of leakage inductance for the test machine.

An expression similar to Equation (6.6) has been presented in [Mal92] where the exponent \( d \) can take any value between -0.1 to -0.27 depending on the value of n. It appears that the exponent \( d \) takes a value closer to -0.27 at low values of n.
Therefore, $d = -0.28$ to $-0.32$ as derived for the given test results can be justified for the range of harmonics applied here (ie from $n = 1$ up to 19). An empirical model similar to Equation (6.6) has been given in [Buc84] where the exponent $d = -0.16$. The model has been derived by performing locked rotor tests on a number of induction motors for harmonic order up to 200 as described in Chapter 2 indicating that a smaller absolute value should be considered for $d$ at higher frequencies.

![Figure 6.7: Experimental values and fitted curve for total leakage inductance vs harmonic order at different tests](image)

6.3 Estimation of Harmonic Losses

The variation of stator and rotor resistances, $R_{1n}$ and $R_{2n}$, as a function of harmonic order has been presented in [Buc84] and given by Equations (2.7) and (2.9) as:

$$R_{1n} = R_{1dc}(1 + C_1 h^4 n^2)$$  \hspace{1cm} (6.7)

$$R_{2n} = R_{2dc}(1 + C_2 h n^{0.5})$$ \hspace{1cm} $n>1$  \hspace{1cm} (6.8)

where $R_{1dc}$ and $R_{2dc}$ are the DC stator and rotor resistances respectively. The values for coefficients are, $C_1 = 1.58 \times 10^{-5}$, $C_2 = 0.18$ to $0.35$, $h$ is the stator/rotor slot depth in cm and $n$ is the order of the harmonic frequency in the rotor and/or stator.
respectively. For simplicity $h$ in the rotor and stator are assumed to be the same as already discussed in Chapter 2.

According to Equation (2.8) the slot depth for the test motor is estimated to be about 1.9 cm. A preliminary calculation shows that the stator winding resistance defined by Equation (6.7) doubles with respect to its DC value for harmonic order $n \approx 70$. Therefore, $R_{1n}$ of the test motor can be assumed to be constant for the given harmonic frequencies which are less than 1 kHz (i.e., $n \leq 20$). Consequently, the stator winding resistance for all harmonic orders is assumed to be equal to its DC value, $R_{1dc}$, which only varies with temperature.

The rotor resistance, however, varies more rapidly with harmonic frequency due to deep bar effect as previously described in Chapter 2. According to Equation (6.8) the rotor effective resistance for the test motor doubles at $n \approx 4$, as compared with its DC value, which demonstrates the sensitivity of the rotor resistance to the harmonic order.

The stator and rotor harmonic copper losses, $R_n I_n^2$, can be estimated using Equations (6.7), (6.8) and the harmonic current $I_n$. In addition to the measured values, harmonic currents can be approximated as a ratio $V_n/X_n$ ignoring the effect of $R_n$. As described in previous sections, the estimated values of $X_n$ using models given in [Buc84] and [Cum86] are different from the experimental data. Therefore, the estimated harmonic currents using any of these two models will be different from the measured values of harmonic currents. A comparison confirmed a 40% difference between the measured values of harmonic currents and the estimated values using the model given in [Buc84].

According to [Buc84], an expression can be derived to predict harmonic iron and stray losses, $W_{cn}$, as:

$$W_{cn} = W_c (5.5 n^{-0.5} V_n^2) \quad \text{for } n>2 \quad (6.9)$$
where $W_c$, is the iron losses under nominal conditions and $V_n$ is the pu harmonic voltage.

In order to compare the experimental results with the results estimated by the loss model given in [Buc84], a spreadsheet was developed. For the test motor, harmonic losses are estimated for the stator, rotor and for the iron based on the measured values of harmonic currents and added to give the total harmonic losses. A comparison between the measured and estimated results confirmed a negligible discrepancy between the two indicating that the model presented in [Buc84] adequately describes the frequency variation of the stator and rotor resistances of the test motor. However, the test motor leakage reactance, $X_n$, for different harmonic orders and hence harmonic currents, $I_n$, are not estimated accurately using the models given in [Buc84] and [Cum86].

These models can be modified to predict the test motor parameters as close as possible to the experimental results. The modifications have to be in the estimation of harmonic currents. This, in turn, requires modification of the model to represent an accurate prediction for variation of leakage reactance/inductance with harmonic order. As discussed in Sections 6.2.3, the best estimation for the test motor leakage reactance, $X_n$, can be made by using:

$$X_n = 0.20n^{0.65} \quad (6.10)$$

where the exponent is different from that given in the described loss models. As a result, the harmonic currents, $I_n$, can be estimated so that a minimum difference with the measured values is achieved.

Equation (6.9) also needs to be modified according to the modified model given for $X_n$ as:

$$W_{cn} = W_c(3.5n^{-0.3}V_n^2) \quad n>2 \quad (6.11)$$
Finally, $C_2$ in Equation (6.8) is also adjusted so that a minimum difference is achieved between the measured and estimated harmonic losses. The new value of $C_2 = 0.046$ is within the specified range for machines with $P < 10$ kW as suggested in [Buc84].

The modified loss model is employed to estimate the individual harmonic losses in the test motor. The total harmonic losses corresponding to each test are calculated as the sum of individual losses and compared with the measured values. There were some discrepancies between the estimated and measured losses mostly due to the measurement error.

### 6.4 Variation of Harmonic Losses with THD

Total pu harmonic losses can be expressed as:

$$W_n(pu) = R_n I_n^2 = (R_n^{0.5})(\frac{V_n}{X_n^{0.65}})^2$$

$$= \frac{R}{X^2} V_n^2 n^{-0.8}$$

(6.12)

which suggests that total harmonic losses are almost inversely proportional to the harmonic order $n$. Therefore, for a constant $V_n$, low order harmonics cause more pronounced harmonic losses in induction motors. This confirms that the THD cannot be considered as the most appropriate criteria in applying supply distortion to estimate induction motor harmonic losses. However, using Equation (6.12), a more appropriate figure, a weighted THD (WTHD), can be defined as:

$$WTHD = \sqrt{\sum V_n^2 n^{-0.8}}$$

(6.13)

which gives a larger weighting to the lower order harmonics.

The quantity $R/X^2$ varies for machines with different power ratings and can be used to specify the motor harmonic limits. The larger the machine the lower the $R/X$ ratio is and hence the higher WTHD that can be specified.
6.5 Specifying Derating for Induction Motors

Equations (6.12) and (6.13) can be utilised to specify a derating factor for induction motors when supplied by distorted waveforms. The derating factor should be determined so that the machine heating does not exceed the allowable limit while harmonics are present. In other words, the extra losses produced due to the presence of harmonics should be compensated by reducing the rated load according to a suitable derating factor.

Based on an analysis presented in Appendix E total fundamental losses, \( W_{\text{total}} \), can be approximated by:

\[
W_{\text{total}} = W_{\text{const}} + W_{\text{load}}
\]

(6.14)

where \( W_{\text{const}} \) represents that part of the losses which is independent of load and \( W_{\text{load}} \) is the load dependent part of the losses which can be expressed as:

\[
W_{\text{load}} = (1 + 2I_m \Phi_0)I_2^2R
\]

(6.15)

where \( I_m \) is the magnetising current, \( I_2 \) is the rotor current representing the load, \( R \) is the pu total machine resistance (ie \( R = R_1 + R_2 + R_h \)) and \( \Phi_0 \) is the phase angle between the input voltage and \( I_2 \) at full load. It is shown in Appendix E that \( \Phi_0 \) can be approximated by:

\[
\Phi_0 = \tan^{-1}\left( \frac{X}{R_1 + R_2/s_0} \right)
\]

(6.16)

where \( s_0 \) is the full load slip.

In the presence of harmonic losses, \( W_n \), a new value should be defined for \( I_2 \) so that the sum of \( W_n \) and \( W_{\text{load}} \) due to new \( I_2 \) is equal to \( W_{\text{load}} \) due to the rated \( I_2 = 1 \) pu. The new value for \( I_2 \) then can be specified by defining a derating factor, \( DF \), as:

\[
DF = \sqrt{1 - \frac{W_n}{(1 + 2I_m \Phi_0)R}}
\]

(6.17)
By substituting Equation (6.12) in Equation (6.17), the DF can be re-expressed as:

\[ DF = \sqrt{1 - \frac{WTHD^2}{X^2(1 + 2I_m\Phi_0)}} \]  

(6.18)

having a value between 0 and 1. When \( WTHD = 0 \) then \( DF = 1 \), and when:

\[ WTHD = WTHD_{\text{max}} = X \sqrt{(1 + 2I_m\Phi_0)} \]  

(6.19)

then \( DF = 0 \) indicating motor should operate at no-load so that it can tolerate extra heating. It should be noted that the derived DF is valid under conditions close to full load, therefore, \( WTHD_{\text{max}} \) only indicates a limit for \( WTHD \) and cannot be applied practically.

**Example:** Typical parameters corresponding to four induction motors, including the test motor, having different power ratings [Per96] [IEEE87] are given in Table 6.5. In order to specify derating factors for the given motors three different distorted waveforms as described in Table 6.6 are chosen. All three waveforms have 10% THD, however, waveforms I and III contain 5th and 19th harmonic respectively while waveform II contains different order/magnitude harmonics. The calculated THD and WTHD for each waveform are also shown in Table 6.6.

<table>
<thead>
<tr>
<th></th>
<th>Test motor</th>
<th>Motor A</th>
<th>Motor B</th>
<th>Motor C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>7.5 kW, 415 V</td>
<td>3.7 kW, 460 V</td>
<td>300 kW, 415 V</td>
<td>1.645 MW, 11 kV</td>
</tr>
<tr>
<td></td>
<td>4-pole</td>
<td>2-pole</td>
<td>4-pole</td>
<td>4-pole</td>
</tr>
<tr>
<td>( R_1 ) (pu)</td>
<td>0.04</td>
<td>0.052</td>
<td>0.006</td>
<td>0.0063</td>
</tr>
<tr>
<td>( R_2 ) (pu)</td>
<td>0.044</td>
<td>0.041</td>
<td>0.009</td>
<td>0.0063</td>
</tr>
<tr>
<td>( X ) (pu)</td>
<td>0.18</td>
<td>0.123</td>
<td>0.188</td>
<td>0.22</td>
</tr>
<tr>
<td>( S_0 )</td>
<td>0.04</td>
<td>0.039</td>
<td>0.0089</td>
<td>0.0055</td>
</tr>
<tr>
<td>( I_m ) (pu)</td>
<td>0.35</td>
<td>0.4</td>
<td>0.3</td>
<td>0.3</td>
</tr>
<tr>
<td>( WTHD_{\text{max}} )</td>
<td>0.190</td>
<td>0.128</td>
<td>0.198</td>
<td>0.232</td>
</tr>
</tbody>
</table>

Table 6.5: Typical parameters for three induction motors with different power ratings
The calculated values for derating factor due to the given distorted waveforms are shown in Table 6.7. It can be seen that derating becomes more appreciable at higher values of WTHD. The given data confirms that THD is not an appropriate criterion to specify harmonic limits for induction motors. The derating factor particularly becomes significant when smaller machines such as motor A experience distorted voltages.

<table>
<thead>
<tr>
<th>Waveform</th>
<th>WTHD</th>
<th>Test motor</th>
<th>Motor A</th>
<th>Motor B</th>
<th>Motor C</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>0.053</td>
<td>0.96</td>
<td>0.91</td>
<td>0.96</td>
<td>0.97</td>
</tr>
<tr>
<td>II</td>
<td>0.046</td>
<td>0.97</td>
<td>0.93</td>
<td>0.97</td>
<td>0.98</td>
</tr>
<tr>
<td>III</td>
<td>0.031</td>
<td>0.99</td>
<td>0.97</td>
<td>0.99</td>
<td>0.99</td>
</tr>
</tbody>
</table>

Table 6.6: Different distorted waveforms having the same THD but different WTHD

Table 6.7: Derating factor due to distorted waveforms for the different machines
heating a parameter known as service factor, SF, typically equal to 1.15 has been defined where the motor temperature rise could be 6-10°C more than the allowable limit. For the given motors, the maximum derating factor can be specified according to the 1.15 service factor as:

\[
DF_{\text{max}} = \frac{1}{\sqrt{1.15}} = 0.93
\]  

(6.20)

This figure confirms that the given distorted waveforms (I, II and III), except one case, can be safely applied to all motors. Also it can be said that the 5% limit given for the THD in the utility power system network [AS91] [Bai82] is a conservative value in relation to induction motors. The limits for individual harmonics, 4% for odd harmonics and 2% for even harmonics, are even more tight since induction motors can tolerate a higher distorted voltage. Specifying the harmonic order is also required when determining standard harmonic limits for induction motors.

Using \(DF_{\text{max}} = 0.93\) and Equation (6.18), the maximum allowable value for WTHD calculated for the different induction motors are shown in Table 6.8. It can be seen that the larger machines can tolerate extra losses as compared with smaller machines.

<table>
<thead>
<tr>
<th>WTHD</th>
<th>Test motor</th>
<th>Motor A</th>
<th>Motor B</th>
<th>Motor C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.069</td>
<td>0.046</td>
<td>0.071</td>
<td>0.084</td>
</tr>
</tbody>
</table>

Table 6.8: Maximum allowable WTHD to give a DF based on SF = 1.15

6.6 Conclusions

Experimental results obtained from harmonic tests on a 7.5 kW cage induction motor operating under different supply and loading conditions has been analysed in this chapter. It has been demonstrated that the machine total resistance increases with both load and harmonic order. The increase due to the load is mostly because
of the temperature increase in the stator windings and rotor bars. This, in turn, leads to the increase of harmonic losses in the stator and rotor as the machine is loaded up. In other words, apart from the temperature effect, load has no significant influence on machine harmonic losses. The variation of motor total resistance with harmonic order has also been examined where a consistent trend has been found under different loading conditions.

It has been demonstrated that the machine total impedance is nearly independent of motor load but varies non-linearly with harmonic order. This implies that, for a given distorted voltage, harmonic currents flowing through the machine are the same regardless of the motor load level. This has been justified by comparing the harmonic currents measured under different load levels.

The variation of total leakage reactance/inductance with harmonic order has been investigated and appropriate expressions have been derived. Wherever applicable, the estimated parameters have been compared with the existing harmonic loss models. For a given distorted voltage, the effect of motor load on machine leakage reactance/inductance was found to be insignificant. However, these parameters significantly change with harmonic order mainly due to deep bar effect.

It is shown that the contribution of low order harmonics to motor harmonic losses is more than that for higher order ones and hence, THD is not an adequate measure to specify the motor harmonic limits. A Weighted THD (WTHD) has been defined to highlight the significance of the harmonic order in the estimation of induction motor harmonic losses.

A derating factor (DF) has been suggested to allow for additional harmonic losses in induction motors when supplied by distorted voltages. Applicability of the derived DF on several induction motors with different power ratings, from 3.7 kW to more than 1 MW has been justified. It is demonstrated that larger machines can tolerate more harmonic losses as compared with the smaller machines. The WTHD limits are defined for different machines based on a parameter known as the motor service
factor (SF). It has been demonstrated that with a typical value of 1.15 for the service factor, most commonly available induction motors can tolerate extra losses due a THD more than 10%. This suggests the conservative nature of the standards in allowing a maximum THD of 5% in the utility power supply in relation to induction motors.
Chapter 7

Conclusions and Recommendations

7.1 Double Chamber Calorimeter (DCC)

As a major contribution from this project, a new open type double chamber calorimeter (DCC) has been developed to accurately measure losses of a 7.5 kW cage induction motor. The DCC is capable of measuring machine losses up to 1 kW with a resolution of 10 W and with an accuracy of 4%. A simple loss model has been developed to estimate the conducted heat leakage through the calorimeter insulation material with an accuracy of ±1 W. The model utilised a conduction shape factor to include the conducted heat leakage through the calorimeter edges.

The thermodynamic performance of the DCC has been examined using two heaters in the separate chambers. Important aspects such as limits for the air flow rate through the DCC, temperature rise inside and across each chamber have been derived. A minimum air flow rate of 55 L/s ensures the normal operation of the test motor inside the calorimeter. In order to measure a heat loss around 1 kW with acceptable temperature rise an air flow rate of about 100 L/s is required. The limits for the air temperature rise across each chamber was determined to be in the range 2-10°C. It has been demonstrated that a high accuracy can be achieved when the air flow rate is adjusted to obtain the maximum air temperature rise across each chamber.

The DCC is compared with the single chamber type calorimeter whose application has been reported in the literature. It has been justified that the DCC is much simpler and cheaper than the single chamber type calorimeter where the same order of accuracy can be achieved in loss measurement. The DCC approach provides a situation where no critical measurement of air properties (except for the temperature) is required during the course of loss estimation. This is considered to
be the most important advantage of the DCC when compared with the single chamber calorimeter. Finally, it has been demonstrated that the DCC provides a reliable, convenient and accurate facility for measurement of motor heat loss regardless of the motor input voltage waveforms.

7.2 Induction Motor Harmonic Tests

One objective of this thesis was to study the effects of time harmonics on the behaviour of mains-connected induction motors when fed from distorted voltage waveforms. Among the associated unwanted effects, additional losses which increase the machine heating have been investigated in this thesis. Various harmonic tests have been conducted on the test induction motor using the harmonic generator (HG) as a controllable distorted voltage source. The tests were performed under no-load, half load and full load conditions by applying balanced distorted waveforms (therefore containing no triplen harmonics) up to 1 kHz. Analysis of the experimental data obtained from the harmonic tests on 7.5 kW induction motor has resulted in the followings conclusions.

For a given distorted voltage containing a known harmonic frequency, additional motor losses significantly increase with machine load. In the worst case as in full load condition and in the presence of 5th harmonic this figure could be as high as 2 times larger than for no-load. It has been demonstrated that the extra losses are mainly due to the increased motor resistance (stator and rotor windings) because of increased temperature. This, in turn, yields an interesting and useful conclusion that harmonic losses in the motor are independent of motor load and only increase due to the temperature effect.

The outcome discussed above can be particularly utilised for determining harmonic losses in large induction motors as supplying them with a controlled distorted voltage is practically impossible especially under full load conditions. Alternatively, machine overheating under full load conditions affected by voltage
distortion can be obtained from harmonic tests at no-load in conjunction with a temperature correction as for full load condition.

For a given distortion level, low order harmonics cause more pronounced losses in the motor as compared with the low order harmonics. This results in a straightforward conclusion that the THD is not an adequate concept to specify motor harmonic limits. Instead, a weighted THD (WTHD) has been recommended to highlight the significance of the harmonic order in the estimation of induction motor harmonic losses.

The variation of test motor parameters with harmonic order as well as the variation of additional losses with WTHD has led to specification of a derating factor (DF) for induction motors. The DF is basically a function of motor parameters as well as the supply distortion level (both magnitude and harmonic order). Depending upon the supply WTHD, a DF can be determined which suggests the fraction of machine loading under which the additional losses due to the distorted supply can be safely tolerated by the machine. This figure has been found for several machines with various power ratings from 3.7 kW to 1.6 MW. The results confirmed that a higher WTHD can be applied to larger machines as compared with smaller machines. In other words, larger machines are more capable of handling additional harmonic losses due to the supply distortion. It has also been demonstrated that most of induction motors can afford a WTHD up to 8% if a service factor of 1.15 is applied. The 8% figure corresponds to an average THD of about 15% which is much larger than the 5% limit for THD in the utility power network as specified by standards.

### 7.3 Further Work

Several suggestions are given for future work in relation to this project:

1. Hot spot temperatures due to distorted voltages can be investigated by developing experimental setup to measure temperature at different parts inside the machine. Minor modifications are required to install temperature sensors in the stator. However, temperature measurement of rotor bars, which become very significant
under harmonic conditions, is not an easy task. During the early stages of this project, development of an FM transmission technique was suggested for rotor temperature detection. However, due to the limited funding, this idea did not come to practice but it could be an action to be taken in the future.

2. Performing calorimetric tests in the presence of multiple number of harmonic frequencies is suggested. This would be required in order to investigate the overall effect of different order harmonics and their interaction on the performance of induction motors and the resulting additional losses.

3. The effect of an unbalanced distorted supply voltage on the performance of induction motors seems to be an interesting idea to be investigated. Theoretically, a minor unbalance in the supply can produce a relatively large negative sequence component. Under normal operational conditions, the negative sequence components of the supply introduce a frequency of twice of the supply frequency to the rotor of an induction motor. The corresponding effects can be analysed based on the second order harmonic which can cause a significant additional losses in the rotor. This will be complicated when unbalanced harmonics are present in the supply voltage of an induction motor.
References


References


References


Appendix A: Harmonic Generator [Gos93]

The harmonic generator (HG) is a three phase 10 kVA 415 V controllable inverter capable of producing balanced harmonically distorted waveforms. In addition to fundamental frequency (50 Hz) it is possible to produce distorted waveforms containing individual or a combination of harmonics up to 20th (1 kHz) and with a maximum THD of 40%.

The HG input power circuit is supplied by 3-phase 415 V 50 Hz mains via a step up autotransformer giving line-to-line voltage of 580 V. Therefore, the inverter DC bus voltage is about 800 V which is about 40% higher than that in most of the standard inverter circuits. The AC/DC conversion takes place through a 3-phase diode rectifier bridge via a 1 kΩ soft charge resistor followed by the DC bus capacitors having 1200 V voltage rating. In the output circuit six high power IGBTs are utilised which can operate with a maximum switching frequency of 10 kHz. These are controlled by a 80C196KB programmable microcontroller (µC) connected to a PC for data transfer. A computer software program has been developed to accept the desired input data from the user as:

- DC bus voltage
- Number of harmonics required
- Order of the requested harmonic/s (up to 20th)
- Phase of the requested harmonic/s compared to the fundamental
- Fundamental magnitude
- Percentage magnitude of the requested harmonic/s (up to 40%).

The program then calculates the switching intervals accordingly and displays the three phase voltage waveforms on the screen. The PWM switching pattern for one
complete cycle along with the voltage spectrum will also be displayed respectively. Finally, the calculated values for the switching intervals will be downloaded to the µC via the PC serial port. The program then confirms few protection precautions before allowing the high frequency switching to take place.

Similar to most inverters, the output voltage of the HG contains a high frequency of twice the switching frequency, i.e., 20 kHz. The high frequency has been eliminated using a second order LC filter with a resonance frequency of 2.3 kHz as described in Chapter 5. Experimental results confirmed that the filter eliminated high frequency signals to a desired level as discussed in Chapter 5.

According to the experimental results, the magnitude of the produced voltage by the HG was different from the requested data by a maximum factor of ±5%. The reason was found to be the slight difference between the entered value for the DC link capacitor voltage and the actual value. The switching intervals are calculated according to the given value of DC bus voltage but produced according to the actual value. Therefore, any difference between these two results in a waveform different (in magnitude) from the requested one. If the value of DC bus voltage supplied by the user is smaller than the actual one, the produced waveforms will have a higher magnitude than the requested data and vice versa. However, choosing a value for the DC bus voltage as close as possible to the actual value helped to overcome this problem.

The output voltage of the HG measured after the filter was also subject to some variation due to the HG loading level. At full load, some 10% voltage drop was detected at the load input. This is assumed to be partly due to the voltage drop across the series inductances of the filter. The voltage drop across the switching components was found to be another source of output voltage drop. Reduction of the DC link capacitor voltage could also be another reason. As a quick and efficient solution to overcome this shortage, the requested voltage was set to be higher than the desired one (say by 10%). Further investigation is required to find a more appropriate and long-term solution for this problem.
In general, it should be noted that, in producing a waveform, the entered data are just some initial estimations for the requested waveform and cannot be used for analytical purposes. Instead, the voltage of the HG at the output of the filter (load voltage) was measured accurately using the voltage measurement circuit as described in Chapter 3 and used for data analysis. Therefore, most of the problems described before, in spite of being treated, had no essential impact on the actual harmonic tests on the motor.
Appendix B: Air properties, thermocouples and insulation material

B.1: Air specific heat and density

The variation of air specific heat, \( c_p \), and air density with temperature and relative humidity is illustrated in Figures B.1 and B.2 respectively.

B.2: Thermocouples

The temperature range of different type thermocouples are shown in Table A and Table B. Typical curves (thermoelectric voltage vs temperature) for different types of thermocouples are shown in Figure B.3. Also, a lookup table for conversion of thermoelectric voltage to temperature for T type thermocouples (Copper vs Constantan) is shown in Table B.1.

B.3 Specifications of the Insulation Material

Specifications of the class VH expanded polystyrene (EPS) insulation material is given in pages 147-149.
Figure B.1: Variation of air specific heat vs temperature and relative humidity
[IEC74]
Figure B.2: Variation of air density vs temperature and relative humidity [IEC74]
### TABLE A

**USEFUL RANGES OF THERMOCOUPLES**

<table>
<thead>
<tr>
<th>BASE METALS</th>
<th>DEGREES F</th>
<th>EMF (MV)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Copper/Constantan</td>
<td>-300 to 750</td>
<td>-5.284 to 20.805</td>
</tr>
<tr>
<td>Iron/Constantan</td>
<td>-300 to 1600</td>
<td>-7.52 to 50.05</td>
</tr>
<tr>
<td>Chromel/Alumel</td>
<td>-300 to 2300</td>
<td>-5.51 to 51.05</td>
</tr>
<tr>
<td>Chromel/Constantan</td>
<td>32 to 1800</td>
<td>0 to 75.12</td>
</tr>
<tr>
<td>Platinum 10% Rhodium / Platinum</td>
<td>32 to 2800</td>
<td>0 to 15.979</td>
</tr>
<tr>
<td>Platinum 13% Rhodium / Platinum</td>
<td>32 to 2900</td>
<td>0 to 18.636</td>
</tr>
<tr>
<td>Platinum 30% Rhodium / Platinum 6% RH</td>
<td>100 to 3270</td>
<td>.007 to 13.499</td>
</tr>
<tr>
<td>Platinel 1813 / Platinel 1503</td>
<td>32 to 2372</td>
<td>0 to 51.1</td>
</tr>
<tr>
<td>Iridium / Iridium 60% Rhodium 40%</td>
<td>2552 to 3326</td>
<td>7.30 to 9.56</td>
</tr>
<tr>
<td>Tungsten 3% Rhenium / Tungsten 25% Rhenium</td>
<td>50 to 4000</td>
<td>.064 to 29.47</td>
</tr>
<tr>
<td>Tungsten / Tungsten 26% Rhenium</td>
<td>60 to 5072</td>
<td>.042 to 43.25</td>
</tr>
<tr>
<td>Tungsten 5% Rhenium / Tungsten 26% Rhenium</td>
<td>32 to 5000</td>
<td>0 to 38.45</td>
</tr>
</tbody>
</table>

### TABLE B

**SHEATH MATERIAL TEMPERATURE CHARACTERISTICS**

<table>
<thead>
<tr>
<th>METAL</th>
<th>RECOMMENDED MAXIMUM OPERATING TEMPERATURE</th>
<th>MELTING TEMPERATURE</th>
</tr>
</thead>
<tbody>
<tr>
<td>COPPER</td>
<td>300 F</td>
<td>1980 F</td>
</tr>
<tr>
<td>ALUMINUM</td>
<td>700</td>
<td>1220</td>
</tr>
<tr>
<td>MONEL²</td>
<td>1000</td>
<td>2450</td>
</tr>
<tr>
<td>LOW CARBON STEEL</td>
<td>1200</td>
<td>2775</td>
</tr>
<tr>
<td>CUFRO-NICKEL 30%</td>
<td>1400</td>
<td>2260</td>
</tr>
<tr>
<td>430 STAINLESS STEEL</td>
<td>1550</td>
<td>2600</td>
</tr>
<tr>
<td>347 STAINLESS STEEL</td>
<td>1650</td>
<td>2550</td>
</tr>
<tr>
<td>316 STAINLESS STEEL</td>
<td>1650</td>
<td>2500</td>
</tr>
<tr>
<td>304 STAINLESS STEEL</td>
<td>1650</td>
<td>2600</td>
</tr>
<tr>
<td>446 STAINLESS STEEL</td>
<td>2000</td>
<td>2550</td>
</tr>
<tr>
<td>310 STAINLESS STEEL</td>
<td>2000</td>
<td>2550</td>
</tr>
<tr>
<td>309 STAINLESS STEEL</td>
<td>2000</td>
<td>2550</td>
</tr>
<tr>
<td>INCONEL³</td>
<td>2100</td>
<td>2600</td>
</tr>
<tr>
<td>HASTELLOY X⁴</td>
<td>2300</td>
<td>2350</td>
</tr>
<tr>
<td>NICKEL</td>
<td>2300</td>
<td>2650</td>
</tr>
<tr>
<td>INCONEL 702²</td>
<td>2400</td>
<td>2570</td>
</tr>
<tr>
<td>PLATINUM</td>
<td>3050</td>
<td>3223</td>
</tr>
<tr>
<td>NIOBIUM (COLUMBIUM)</td>
<td>3600</td>
<td>4474</td>
</tr>
<tr>
<td>MOLYBDENUM</td>
<td>4000</td>
<td>4752</td>
</tr>
<tr>
<td>TANTALUM</td>
<td>4500</td>
<td>5425</td>
</tr>
</tbody>
</table>

---

¹Hoskins Manufacturing Company
²Englehard Industries
³International Nickel Company
⁴Union Carbide Corporation
## TABLE J

TEMPERATURE-MILLIVOLT GRAPH
FOR THERMOCOUPLES

<table>
<thead>
<tr>
<th>Temperature (°F)</th>
<th>Millivolts</th>
</tr>
</thead>
<tbody>
<tr>
<td>500</td>
<td>0</td>
</tr>
<tr>
<td>1000</td>
<td>20</td>
</tr>
<tr>
<td>1500</td>
<td>40</td>
</tr>
<tr>
<td>2000</td>
<td>60</td>
</tr>
<tr>
<td>2500</td>
<td>80</td>
</tr>
<tr>
<td>3000</td>
<td>100</td>
</tr>
<tr>
<td>3500</td>
<td>120</td>
</tr>
<tr>
<td>4000</td>
<td>140</td>
</tr>
<tr>
<td>4500</td>
<td>160</td>
</tr>
</tbody>
</table>

### ANSI SYMBOL

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>T</td>
<td>Copper vs. Constantan</td>
</tr>
<tr>
<td>E</td>
<td>Chromel vs. Constantan</td>
</tr>
<tr>
<td>J</td>
<td>Iron vs. Constantan</td>
</tr>
<tr>
<td>K</td>
<td>Chromel vs. Alumel</td>
</tr>
<tr>
<td>G</td>
<td>Tungsten vs. Tungsten 26% Rhenium</td>
</tr>
<tr>
<td>C</td>
<td>Tungsten 5% Rhenium vs. Tungsten 26% Rhenium</td>
</tr>
<tr>
<td>R</td>
<td>Platinum vs. Platinum 13% Rhodium</td>
</tr>
<tr>
<td>S</td>
<td>Platinum vs. Platinum 10% Rhodium</td>
</tr>
<tr>
<td>B</td>
<td>Platinum 6% Rhodium vs. Platinum 30% Rhodium</td>
</tr>
</tbody>
</table>

*Not ANSI Symbol

Figure B.3: Typical curves (thermoelectric voltage vs temperature) for different types of thermocouples
### Appendix: Conversion of thermoelectric voltage to temperature for T type thermocouples (Copper vs Constantan)

<table>
<thead>
<tr>
<th>DEG C</th>
<th>0</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
<th>DEG C</th>
</tr>
</thead>
<tbody>
<tr>
<td>THERMOELECTRIC VOLTAGE IN ABSOLUTE MILLOUTS</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

#### Table B.1: Conversion of thermoelectric voltage to temperature for T type thermocouples (Copper vs Constantan)
Holotia is the brand name for RMAX block moulded flame retardant modified grade of EPS (expanded polystyrene).

It is a closed cell, resilient, lightweight rigid cellular plastics material produced in a range of densities between 10 kg/m$^3$ and 35 kg/m$^3$.

The main applications for products manufactured from Holotia are thermal insulation systems (wall, roof and sub-floor), ceiling panels and other decorative surfaces, void forms and blockouts, pipe insulation, protection packaging, flotation and buoyancy applications, and stage sets.

The manufacturing process
Pre Expansion
Expandable polystyrene is supplied as plastic beads in which an expanding agent, usually pentane, has been dissolved. In the presence of steam the thermoplastic polystyrene softens and the increasing vapour pressure of the expanding agent causes the beads to expand up to 40 times their original volume. During this stage the degree of expansion is controlled to achieve the desired density. Expanded polystyrene does not contain any ozone depleting substances and none is used in its manufacture.

Conditioning
From the pre-expander the beads are gently transported to large hoppers for ageing. The time of ageing is set to cool and stabilise the beads and allow for infusion of air to replace the expanding spent in the cells.

Moulding
After conditioning the beads are charged into a closed mould where they are further expanded and fused together by steam heating.

Finishing
The freshly moulded blocks of Holotia are passed through temperature controlled ovens to remove moisture and the final traces of expanding agent.

Manufactured to a standard
Holotia EPS is manufactured to AS 1366, Part 3-1992, Rigid Cellular Plastic Sheets for Thermal Insulation, Rigid Cellular Polystyrene, in six classes. The standard designates a colour to identify each of the six classes:

- Class L: Blue
- Class SL: Yellow
- Class S: Brown
- Class M: Black
- Class H: Green
- Class VH: Red

The standard specifies the minimum physical property limits for each of the six classes (see Table 1) and methods for determination of compliance.

Quality Control
To meet the compliance requirements of the standard, the RMAX quality control system monitors and controls each stage of the manufacturing process and assures that Holotia conforms to AS 1366.3 within 95% confidence limits by on site testing of density and key physical properties.

Comprehensive physical testing for product development and quality assurance is carried out in the company's central laboratory, which is NATA approved.

Properties of Holotia
The physical properties are primarily determined by the moulded density for well made-oven cured EPS. See Figures 1 to 4.

However, these properties will be affected by raw material and manufacturing variations, and for this reason Australian Standard 1366-3-1992 specifies the classes in terms of performance.

<table>
<thead>
<tr>
<th>Physical Property</th>
<th>Class</th>
<th>Test Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Compressive stress at 10 percent deformation (min.)</td>
<td>L SL S M H VH</td>
<td>AS 2498.3</td>
</tr>
<tr>
<td>Cross-breaking strength (min.)</td>
<td>kPa</td>
<td>50 70 85 105 135 165</td>
</tr>
<tr>
<td>Rate of water vapour transmission (max.) measured parallel to rise at 23°C</td>
<td>kg/m²s</td>
<td>710 610 510 470 420 400</td>
</tr>
<tr>
<td>Dimensional stability of length, width, thickness (max.) at 70°C, dry condition 7 days</td>
<td>percent</td>
<td>1 1 1 1 1 1</td>
</tr>
<tr>
<td>Thermal resistance (min.) at a mean temperature of 25°C (50mm sample)</td>
<td>m².K/W</td>
<td>1.13</td>
</tr>
<tr>
<td>Flame propagation characteristics: median flame duration (max.)</td>
<td>s</td>
<td>2 2 2 2 2 2</td>
</tr>
<tr>
<td>eighth value (max.)</td>
<td>s</td>
<td>3 3 3 3 3 3</td>
</tr>
<tr>
<td>median volume retained</td>
<td>percent</td>
<td>15 18 22 30 40 50</td>
</tr>
<tr>
<td>eighth value (min.)</td>
<td>percent</td>
<td>12 15 19 27 37 47</td>
</tr>
</tbody>
</table>

1 W/m.K = 3.34 Btu in/hr. °F
Figure 7: Thermal conductivity at 10°C vs density.

Concrete: 0.04
Brick: 0.043
Glass: 0.048
EPS Concrete: 0.12
Wood: 0.35
Compressed Wood: 0.83
Fiberglass: 1.0
EPS - Class SL: 1.13
EPS - Class VH: 1.28

Figure 8: Typical R values, various insulating materials, 50mm thick.

loss of structural integrity of physical properties; core specimens taken from 20 year old freezer rooms show no deterioration.

Unlike some other insulating materials, the k value of Isolite decreases at lower average mean temperatures. (See Figure 9).

Figure 9: Indicative thermal conductivity versus temperature.

High temperature operation

The effect of elevated temperatures on the mechanical properties is an accelerating decline in the values shown in Figures 1 to 5 until at approximately 85°C the so-called zero strength is reached. See Figure 10.

Isolite should not be continuously exposed to temperatures in excess of 80°C as expansion and blistering may occur.

Effect of moisture on k value

The dimensional stability and mechanical properties of Isolite are not affected by water but because absorbed water will increase the k value, as with all insulating materials, care should be taken in designing insulated structures to take account of water and water vapour that may be present.

While Table 3 shows that certain amounts of water are absorbed by EPS under various conditions, Table 4 demonstrates that the loss of R values in EPS as a result of this moisture absorption is minimal. Overseas research has also revealed that the decay in thermal resistance caused by moisture is considerably less for EPS than for either extruded polystyrene foam or cellular glass (see Figure 11).

Table 3: Moisture gain of EPS by liquid water absorption.

<table>
<thead>
<tr>
<th>Test conditions</th>
<th>% by volume</th>
</tr>
</thead>
<tbody>
<tr>
<td>ASTM C-272</td>
<td>2.5</td>
</tr>
<tr>
<td>Submersion</td>
<td>3.0</td>
</tr>
<tr>
<td>10 metre submersion</td>
<td>3.0</td>
</tr>
<tr>
<td>Submersion</td>
<td>6.0</td>
</tr>
<tr>
<td>Submersion</td>
<td>7.8</td>
</tr>
</tbody>
</table>

Table 4: Typical thermal performance by EPS thickness after vapour induced moisture gain.

<table>
<thead>
<tr>
<th>Moisture gain (% by volume at 25mm)</th>
<th>R-value retention %</th>
</tr>
</thead>
<tbody>
<tr>
<td>25mm</td>
<td>50mm</td>
</tr>
<tr>
<td>2</td>
<td>96</td>
</tr>
<tr>
<td>4</td>
<td>92</td>
</tr>
<tr>
<td>6</td>
<td>89</td>
</tr>
<tr>
<td>8</td>
<td>86</td>
</tr>
<tr>
<td>10</td>
<td>84</td>
</tr>
<tr>
<td>12</td>
<td>82</td>
</tr>
<tr>
<td>14</td>
<td>80</td>
</tr>
</tbody>
</table>

As with other building materials care should always be taken to keep Isolite dry before and during installation.
properties rather than density.

The standard lists Nominal Density for each class (see Table 2), but these densities should be regarded as a guide only as the physical properties shown in Table 1 may be achieved by EPS of other densities.

Table 2
Nominal density, kg/m$^3$.

<table>
<thead>
<tr>
<th>Class</th>
<th>L</th>
<th>SL</th>
<th>S</th>
<th>M</th>
<th>H</th>
<th>VH</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>11</td>
<td>13.3</td>
<td>16</td>
<td>19</td>
<td>24</td>
<td>28</td>
</tr>
</tbody>
</table>

Mechanical properties

The density dependency of the main physical properties of Isolite can be seen in Figures 1 to 4: Compressive strength, Cross Breaking strength (flexural strength) Tensile strength and Shear strength.

Compressive creep

It is common to report only the compressive stress at 10% deformation but the latter is often taken from complete Stress-Strain curves as shown in Figure 5. Although it appears to deform elastically over a range of comprehensive loads, Isolite that has been stressed will, with the release of all stress, retain some permanent deformation.

Flotation properties

The density of Isolite is low compared with water, with a nominal density range from 10 to 25 kg/m$^3$ compared with water at 1000 kg/m$^3$. The water buoyancy per cubic metre of Isolite is determined by subtracting its kg/m$^3$ density from 1000. The result is the weight in kilograms which a cubic metre of Isolite can support when fully submerged in water.
Appendix C: Specifications of the measurement system

C.1: Specification of the variable speed fan employed to force the air through the DCC is given in Figure C.1.

C.2: Specifications of the single phase wattmeter is given in page 152.

C.3: Specifications of the three phase digital AC meter is given in pages 153-155.

C.4: Specifications of the isolation amplifiers used for voltage measurement circuits is given in pages 156-159. Also specification of the Hall effect current transducers is shown in page 160.

C.4: Two computer data acquisition boards have been used in this project, CIO-DAS 08 and CIO-DAS 16. The specifications of these boards and a signal conditioning board are given in pages 161-165.
T-Series In-Line Models

In-Line Models in 4 sizes with reversible airflow.
TX—with fully automatic integral Shutter.
TL—without Shutter.
Two lume grey.

Technical Description

- Manufactured in weather-resistant polymeric materials. Integrated component design for maximum aerodynamic efficiency.
- Motor overhung designed. Suitable for running at any angle. Quiet running, totally enclosed. Will not interfere with TV or radio reception.
- Suitable for ambient temperatures from -40°C to +50°C. Fitted with self resetting thermal cut-out.
- Unique Speed Control Pack enables high, medium, and low speed for any air flow. Can be run at any date to suit air change requirements.
- Low density 5-blade impeller and short sleeve section ensure optimum pressure characteristic with minimum sound level.
- Component design allows all parts to be dismantled for cleaning easily and quickly, without the use of tools or fittings.
- Can be used to comply with international safety standards.
- Supplied Voltage 220-240V/50Hz.

Performance

<table>
<thead>
<tr>
<th>Unit Size</th>
<th>Stock Ref. No.</th>
<th>Extract Performance (or similar)</th>
<th>Watts (High)</th>
<th>Shutter (TX only)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LOW m³/h</td>
<td>MEDIUM m³/h</td>
<td>HIGH m³/h</td>
<td>T-Serie Controls Stock Ref. No.</td>
</tr>
<tr>
<td>6</td>
<td>15.17</td>
<td>31.20</td>
<td>63.75</td>
<td>36 11 19</td>
</tr>
<tr>
<td>7</td>
<td>16.21</td>
<td>42.60</td>
<td>85.20</td>
<td>36 11 19</td>
</tr>
<tr>
<td>9</td>
<td>16.37</td>
<td>70.35</td>
<td>141.70</td>
<td>36 11 19</td>
</tr>
<tr>
<td>12</td>
<td>16.47</td>
<td>167.45</td>
<td>324.90</td>
<td>36 11 19</td>
</tr>
</tbody>
</table>

Figure C.1: Specifications of the variable speed fan used to force air through the DCC
2. SPECIFICATIONS

2-1 GENERAL SPECIFICATIONS

1) Display: 0.5" LCD (Liquid Crystal Display) Max. Indication 1999 to -1999.
2) Measurement: DCV/ACV, DCA/ACA, WATTS.
3) Polarity: Bi-polar by a automatic switching, "−" indicates reverse polarity.
4) Zero Adjust: External adjustment for zero of the display is only for watt ranges, this is limited to +30 to −30 digits (ACV/DCV, ACA/DCA: Automatic adjustment).
5) Over-input: Indication of "1" or "−1".
6) Operating Temp: 0°C to 50°C (32°F to 122°F).
7) Operating Humidity: Less than 80% RH.
8) Power Supply: 006 DC 9V battery (heavy duty or alkaline battery).
9) Power Consumption: About 6 mA.
10) Weight: 500g (including battery).
11) Standard Accessories: Instruction manual...1 pcs, Test Lead TL-01...1 pair.

ID Display: 0.5" LCD (Liquid Crystal Display) Max. Indication 1999 to -1999, DCV/ACV, DCA/ACA, WATTS.

2-2 ELECTRICAL SPECIFICATIONS

WATT (true power)

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>2000W</td>
<td>± (1.5% + 1d)</td>
<td>1W</td>
</tr>
<tr>
<td>6000W</td>
<td>± (1.5% + 1d)</td>
<td>10W</td>
</tr>
</tbody>
</table>

Remark:
Input voltage: 0 to 600V AC (Overload protection 1000V)
Input current: 0 to 10 ACA.
Frequency characteristic: 45HZ to 65HZ.
Accuracy Spec: Tested on input voltage over 60V ACV (60 HZ).

AC VOLTAGE

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>200V</td>
<td>±(0.8%+1d)</td>
<td>0.1V</td>
</tr>
<tr>
<td>750V</td>
<td>±(0.8%+1d)</td>
<td>1V</td>
</tr>
</tbody>
</table>

Remark:
Frequency characteristic: 45 HZ to 65HZ.
Converter Response: Average responding, calibrated to display RMS value of sine wave.

DC VOLTAGE

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>200V</td>
<td>±(0.8%+1d)</td>
<td>0.1V</td>
</tr>
<tr>
<td>1000V</td>
<td>±(0.8%+1d)</td>
<td>1V</td>
</tr>
</tbody>
</table>

Remark:
Converter Response: Average responding, calibrated to display RMS value of sine wave.

DC CURRENT

<table>
<thead>
<tr>
<th>Range</th>
<th>Accuracy</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>10A</td>
<td>±(1%+1d)</td>
<td>10mA</td>
</tr>
</tbody>
</table>

Remark:
Converter Response: Average responding, calibrated to display RMS value of sine wave.

3. FRONT PANEL DESCRIPTION

4. PRECAUTIONS AND PREPARATIONS FOR MEASUREMENTS

1) Ensure that 9V battery is connected correctly to its snap terminal and placed in the battery compartment.
2) Depress the correct Function and range PUSH buttons before marking measurements.
3) Place the Test Lead into the proper input terminal before marking measurements.
4) Select the proper measurement range by starting at the highest anticipated value.
5) Remove either of the test leads from the circuit under test while changing the measurement range.

Figure C.2: Specifications of the single phase wattmeter
Type 2503
Digital AC Power Meter
(For 3-phase, 3-wire circuit)
Appendices 154

GENERAL

Voltage and current waveforms of a motor, transformer, thyristor-controlled circuit in a 3-phase line are unavoidably distorted, and the RMS value and power of such distorted waves are difficult to accurately measure by conventional methods. But the Type 2503 Digital AC Power Meter employs an entirely new principle of operation, which permits accurate measurements of the RMS value of distorted waves through simple operation. It is designed to best meet the requirements of general laboratories and on-line applications.

- A single unit of Type 2503 can measure the true RMS values of AC voltage from 3 to 600 V also DC-superimposed voltage), AC current from 100 mA to 30 A, and 3-phase 3-wire wattage from 600 mW to 36 kW. The three different functions are freely selectable by a simple one-touch operation of pushbutton. Wired as indicated on the rear panel, it measures the power in three-phase three-wire circuit by means of two-wattmeter method.

Unlike the conventional thermoelement type instruments, it features a quick response, high protection against overload and reduced instrument loss.

By using photocouplers, the input circuit is completely insulated from the controls and data output terminals. In addition, with sufficient withstand voltage and noise immunity, Type 2503 assures safe of operation and high reliability as an on-line instrument.

Type 2503 uses a LED display which assures long life and high reliability.

This instrument can also be used for data processing with BCD output and remote control as standard features and analog output as an option.

- True RMS value of voltage and current Measurement
  YEW's unique steepest-descent method allows measurement of true RMS values of various types of distorted waves, such as triangular and square waves. In addition, it also measures DC voltage or DC-superimposed voltage.

- Unique power measurement system
  YEW's original "feedback type time division multiplier" allows power measurement at high accuracy.

- ±0.1% accuracy
  Voltage, current and power are measured with a high accuracy of ± (0.1% of reading + 0.02 of range + 1 digit).

- High sensitivity
  Maximum sensitivity is 1 mV/digit, 10 μA/digit and 0.1 mW/digit.

- High withstand voltage
  Voltage input, current input, data output terminals and case are all insulated from each other:
  2200 V: Between input terminals and case, and between input and output terminals.
  1500 V: Between case and power source, and between output terminals and power source.

- Stable measurement with outstanding noise immunity
  Deliberate design minimizes the effect of common mode noise and power line superimposed noise on measurement, so the operation is stable for on-line use.

- BCD output and remote control (standard) and analog output (option)
  Equipped with BCD output and remote control, the instrument can be combined with a digital printer for data processing.
  In addition, by using the optional analog output, it can convert the upper three or lower three digits on the digital display to an analog signal to record slight variations of the input.

- Negligible instrument loss
  Type 2503 is best suited for power measurement of low power factor without disturbing the circuit being measured.

- Wide measurement range
  Voltage from 3 to 600 V, current from 100 mA to 30 A and power from 600 mW to 36 kW can be measured with use of the built-in voltage divider and current transformers. The voltage and current input terminals are provided on the rear panel and the range selectors on the front. The range is freely switchable without requiring any change in wiring at the terminals.

- LED (light-emitting diode) display
  The display uses easy-to-read LED of long life and high reliability.
## Appendices

### Measuring function

<table>
<thead>
<tr>
<th>Operating principle</th>
<th>Measuring range</th>
<th>Effective measuring range</th>
<th>Feedback time division multiplier</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Range</td>
<td>Voltage</td>
<td>Current</td>
</tr>
<tr>
<td></td>
<td>Voltage</td>
<td>Current</td>
<td></td>
</tr>
<tr>
<td>Voltage</td>
<td>3V</td>
<td>0.1A</td>
<td>30.00~10.00mA</td>
</tr>
<tr>
<td></td>
<td>10V</td>
<td>0.3A</td>
<td>100.0~33.0 mA</td>
</tr>
<tr>
<td></td>
<td>30V</td>
<td>3.0A</td>
<td>300.0~100.0 mA</td>
</tr>
<tr>
<td></td>
<td>100V</td>
<td>6.0A</td>
<td>600.0~200.0 mA</td>
</tr>
<tr>
<td></td>
<td>300V</td>
<td>10.0A</td>
<td>1,000~3.300 mA</td>
</tr>
<tr>
<td></td>
<td>600V</td>
<td>30.0A</td>
<td>3,000~11.000 mA</td>
</tr>
<tr>
<td></td>
<td>1,000V</td>
<td>60.0A</td>
<td>6,000~22.000 mA</td>
</tr>
<tr>
<td></td>
<td>3,000V</td>
<td>150.0A</td>
<td>11,000~40.000 mA</td>
</tr>
<tr>
<td></td>
<td>6,000V</td>
<td>300.0A</td>
<td>33,000~110.000 mA</td>
</tr>
<tr>
<td></td>
<td>10,000V</td>
<td>600.0A</td>
<td>66,000~220.000 mA</td>
</tr>
<tr>
<td></td>
<td>30,000V</td>
<td>1,500.0A</td>
<td>220,000~700.000 mA</td>
</tr>
<tr>
<td></td>
<td>60,000V</td>
<td>3,000.0A</td>
<td>700,000~2,200.000 mA</td>
</tr>
<tr>
<td></td>
<td>1,000,000V</td>
<td>6,000.0A</td>
<td>2,200,000~7,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>3,000,000V</td>
<td>15,000.0A</td>
<td>7,000,000~22,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>6,000,000V</td>
<td>30,000.0A</td>
<td>22,000,000~70,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>10,000,000V</td>
<td>60,000.0A</td>
<td>70,000,000~220,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>30,000,000V</td>
<td>150,000.0A</td>
<td>220,000,000~700,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>60,000,000V</td>
<td>300,000.0A</td>
<td>700,000,000~2,200,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>1,000,000,000V</td>
<td>1,500,000.0A</td>
<td>2,200,000,000~7,000,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>3,000,000,000V</td>
<td>3,000,000.0A</td>
<td>7,000,000,000~22,000,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>6,000,000,000V</td>
<td>6,000,000.0A</td>
<td>22,000,000,000~70,000,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>10,000,000,000V</td>
<td>10,000,000.0A</td>
<td>70,000,000,000~220,000,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>30,000,000,000V</td>
<td>30,000,000.0A</td>
<td>220,000,000,000~700,000,000.000 mA</td>
</tr>
<tr>
<td></td>
<td>60,000,000,000V</td>
<td>60,000,000.0A</td>
<td>700,000,000,000~2,200,000,000.000mA</td>
</tr>
<tr>
<td></td>
<td>1,000,000,000,000V</td>
<td>1,000,000,000.0A</td>
<td>2,200,000,000,000~7,000,000,000.000mA</td>
</tr>
</tbody>
</table>

### Resolution

<table>
<thead>
<tr>
<th>Frequency range</th>
<th>DC and 25 Hz to 2 kHz</th>
<th>25Hz to 2 kHz</th>
<th>40 to 1.2 kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resolution</td>
<td>1 mV/digit</td>
<td>10µA/digit</td>
<td>0.1 mW/digit</td>
</tr>
</tbody>
</table>

### Frequency range

<table>
<thead>
<tr>
<th>Crest factor</th>
<th>Less than 3</th>
<th>Less than 3</th>
<th>Corresponds to those of the range of voltage and current measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Accuracy</td>
<td>±0.1% of reading - 0.02% of range = 1 digit</td>
<td>±0.1% of reading - 0.02% of range = 1 digit</td>
<td>at cos θ = 1</td>
</tr>
<tr>
<td>Temperature</td>
<td>Less than ±0.02%/°C</td>
<td>Less than ±0.02%/°C</td>
<td>Less than ±0.02%/°C</td>
</tr>
<tr>
<td>Coefficient</td>
<td>(on temperature range of 5 to 20°C, 26 to 40°C and frequency range of 50 Hz to 1 kHz)</td>
<td>(on temperature range of 5 to 20°C, 26 to 40°C and frequency range of 50 Hz to 1 kHz)</td>
<td>(on temperature range of 5 to 20°C, 26 to 40°C and frequency range of 50 Hz to 1 kHz)</td>
</tr>
</tbody>
</table>

### Input impedance and power consumption

<table>
<thead>
<tr>
<th>Allowable input voltage</th>
<th>Corresponds to those of the range of voltage and current measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Common mode voltage</td>
<td>1,000V</td>
</tr>
<tr>
<td>Allowable momentary input voltage</td>
<td>10 x range or 100A (peak value), whichever is lower</td>
</tr>
<tr>
<td>Effect of common mode voltage (at 50 and 60 Hz)</td>
<td>Corresponds to those of the range of voltage and current measurements</td>
</tr>
</tbody>
</table>

### Auxiliary input terminals

<table>
<thead>
<tr>
<th>Input voltage</th>
<th>1V (input resistance: 10kΩ)</th>
<th>1V (input resistance: 10kΩ)</th>
</tr>
</thead>
</table>

---

Figure C.3: Specifications of the three phase AC meter
FEATURES
Small Size: 4 Channels/inch
Low Power: 25mW (AD204)
High Accuracy: ±0.025% max Nonlinearity (K Grade)
High CMR: 130dB (Gain = 100 V/V)
Wide Bandwidth: 5kHz Full-Power (AD204)
High CMV Isolation: ±2000 V pk Continuous (K Grade)
(Signal and Power)
Isolated Power Outputs
Uncommitted Input Amplifier

APPLICATIONS
Multichannel Data Acquisition
Current Shunt Measurements
Motor Controls
Process Signal Isolation
High Voltage Instrumentation Amplifier

GENERAL DESCRIPTION
The AD202 and AD204 are members of a new generation of low cost, high performance isolation amplifiers. A new circuit design, novel transformer construction, and the use of surface-mounted components in an automated assembly process result in remarkably compact, economical isolators whose performance in many ways exceeds that previously available from very expensive devices. The primary distinction between the AD202 and AD204 is that the AD202 is powered directly from +15V dc while the AD204 is powered by an externally supplied clock (AD246).

The AD202 and AD204 employ transformer coupling and do not require the design compromises that must be made when optical isolators are used: each provides a complete isolation function, with both signal and power isolation internal to the module, and they exhibit no long-term parameter shifts under sustained common-mode stress. Power consumption, nonlinearity, and drift are each an order of magnitude lower than can be obtained from other isolation techniques, and these advantages are obtained without sacrifice of bandwidth or noise performance.

The design of the AD202 and AD204 emphasizes ease of use in a broad range of applications where signals must be measured or transmitted without a galvanic connection. In addition, the low cost and small size of these isolators makes component-level circuit applications of isolation practical for the first time.

PRODUCT HIGHLIGHTS
The AD202 and AD204 are full-featured isolators offering numerous benefits to the user:

Small Size: The AD202 and AD204 are available in SIP and DIP form packages. The SIP package is just 0.25" wide, giving the user a channel density of four channels per inch. The isolation barrier is positioned to maximize input to output spacing. For applications requiring a low profile, the DIP package provides a height of just 0.350".

High Accuracy: With a maximum nonlinearity of ±0.025% for the AD202K/AD204K (±0.05% for the AD202J/AD204J) and low drift over temperature, the AD202 and AD204 provide high isolation without loss of signal integrity.

Low Power: Power consumption of 35mW (AD204) and 75mW (AD202) over the full signal range makes these isolators ideal for use in applications with large channel counts or tight power budgets.

Wide Bandwidth: The AD204's full-power bandwidth of 5kHz makes it useful for wideband signals. It is also effective in applications like control loops, where limited bandwidth could result in instability.

Excellent Common-Mode Performance: The AD202K/AD204K provide ±2000V pk continuous common-mode isolation, while the AD202J/AD204J provide ±1000V pk continuous common-mode isolation. All models have a total common-mode input capacitance of less than 5pF inclusive of power isolation. This results in CMR ranging from 130dB at a gain of 100 to 104dB (minimum at unity gain) and very low leakage current (2nA maximum).

Flexible Input: An uncommitted op amp is provided at the input of all models. This provides buffering and gain as required, and facilitates many alternative input functions including filtering, summing, high-voltage ranges, and current (transimpedance) input.

Isolated Power: The AD204 can supply isolated power of ±7.5V at 2mA. This is sufficient to operate a low-drift input preamp, provide excitation to a semiconductor strain gage, or to power any of a wide range of user-supplied ancillary circuits. The AD202 can supply ±7.5V at 0.4mA which is sufficient to operate adjustment networks or low-power references and op amps, or to provide an open-input alarm.
## SPECIFICATIONS

(typical \( T = +25^\circ C \) and \( V_s = +15V \) unless otherwise noted)

<table>
<thead>
<tr>
<th>Model</th>
<th>AD204J</th>
<th>AD204K</th>
<th>AD202J</th>
<th>AD202K</th>
</tr>
</thead>
<tbody>
<tr>
<td>GAIN</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Range</td>
<td>1V/V-1000V/V</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Error</td>
<td>( \leq 0.5% \text{typ} \ (\leq 1% \text{max}) )</td>
<td>( \leq 0.5% \text{typ} \ (\leq 1% \text{max}) )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>vs. Temperature</td>
<td>( \pm 20ppm/\text{°C} \ (\pm 40ppm/\text{°C} \text{max}) )</td>
<td>( \pm 20ppm/\text{°C} \ (\pm 40ppm/\text{°C} \text{max}) )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>vs. Time</td>
<td>( \pm 50ppm/1000 \text{Hours} )</td>
<td>( \pm 50ppm/1000 \text{Hours} )</td>
<td></td>
<td></td>
</tr>
<tr>
<td>vs. Supply Voltage</td>
<td>( \pm 0.001%V )</td>
<td>( \pm 0.001%V )</td>
<td>( \pm 0.001%V )</td>
<td>( \pm 0.001%V )</td>
</tr>
<tr>
<td>Nonlinearity (G=1/V/V)</td>
<td>( \pm 0.01% \text{max} )</td>
<td>( \pm 0.025% \text{max} )</td>
<td>( \pm 0.025% \text{max} )</td>
<td>( \pm 0.025% \text{max} )</td>
</tr>
</tbody>
</table>

| INPUT VOLTAGE RATINGS |        |        |        |        |
| Linear Differential Range | \( \pm 5V \) | \( \pm 5V \) | \( \pm 5V \) | \( \pm 5V \) |
| Max CMV | 750V rms | 1500V rms | 750V rms | 1500V rms |
| Continuous (dc and ac) | \( \pm 1000V \text{peak} \) | \( \pm 2000V \text{peak} \) | \( \pm 1000V \text{peak} \) | \( \pm 2000V \text{peak} \) |
| Common-Mode Rejection (CMR), @ 60Hz | \( R_n = 100k \Omega \ (\text{HI & LO Inputs}) \ G = 1 \) | \( R_n = 100k \Omega \ (\text{HI & LO Inputs}) \ G = 100 \) | \( R_n = 100k \Omega \ (\text{HI & LO Inputs}) \ G = 1 \) | \( R_n = 100k \Omega \ (\text{HI & LO Inputs}) \ G = 100 \) |
| Leakage Current Input to Output | 104dB min | 104dB min | 104dB min | 104dB min |

| INPUT IMPEDANCE |        |        |        |        |
| Differential (G = 1/V/V) | \( 10^{12} \Omega \) | \( 10^{12} \Omega \) | \( 10^{12} \Omega \) | \( 10^{12} \Omega \) |
| Common Mode | \( 2G\mu F \pm 5% \) | \( 2G\mu F \pm 5% \) | \( 2G\mu F \pm 5% \) | \( 2G\mu F \pm 5% \) |

| INPUT BIAS CURRENT |        |        |        |        |
| Initial, \( @ +25^\circ C \) | \( \pm 100nA \) | \( \pm 100nA \) | \( \pm 100nA \) | \( \pm 100nA \) |
| vs. Temperature (0 to +70°C) | \( \pm 2nA \) | \( \pm 2nA \) | \( \pm 2nA \) | \( \pm 2nA \) |

| INPUT DIFFERENCE CURRENT |        |        |        |        |
| Initial, \( @ +25^\circ C \) | \( \pm 5pA \) | \( \pm 5pA \) | \( \pm 5pA \) | \( \pm 5pA \) |
| vs. Temperature (0 to +70°C) | \( \pm 2nA \) | \( \pm 2nA \) | \( \pm 2nA \) | \( \pm 2nA \) |

| INPUT NOISE |        |        |        |        |
| Voltage, 0.1 to 100Hz | \( 50\mu V/\sqrt{Hz} \) | \( 50\mu V/\sqrt{Hz} \) | \( 50\mu V/\sqrt{Hz} \) | \( 50\mu V/\sqrt{Hz} \) |
| Frequency Response |        |        |        |        |
| Bandwidth (\( V_o \leq 10V \text{p-p}, G = 1-50V/V \)) | \( 5kHz \) | \( 5kHz \) | \( 2kHz \) | \( 2kHz \) |
| Settling Time (to 10% of 10V Step) | 1ms | 1ms | 1ms | 1ms |
| Offset Voltage (RTD) | \( \pm 15 \pm 15G \text{mV max} \) | \( \pm 15 \pm 15G \text{mV max} \) | \( \pm 15 \pm 15G \text{mV max} \) | \( \pm 15 \pm 15G \text{mV max} \) |

| RATED OUTPUT |        |        |        |        |
| Voltage, No Load | \( \pm 5V \) | \( \pm 5V \) | \( \pm 5V \) | \( \pm 5V \) |
| Voltage at Out HI or Out LO (Ref. Pin 32) | \( 3k\Omega \) | \( 3k\Omega \) | \( 7k\Omega \) | \( 7k\Omega \) |
| Output Ripple, 1kHz Bandwidth | \( 0.5\mu V \text{rms} \) | \( 0.5\mu V \text{rms} \) | \( 0.5\mu V \text{rms} \) | \( 0.5\mu V \text{rms} \) |

| ISOLATED POWER OUTPUT |        |        |        |        |
| Voltage, No Load | \( \pm 7.5V \) | \( \pm 7.5V \) | \( \pm 7.5V \) | \( \pm 7.5V \) |
| Current, No Load to Full Load | 2mA (Either Output) | 2mA (Either Output) | 2mA (Either Output) | 2mA (Either Output) |

| OSCILLATOR DRIVE INPUT |        |        |        |        |
| Input Voltage | \( 15V \text{pk-pk nominal} \) | \( 15V \text{pk-pk nominal} \) | \( 15V \text{pk-pk nominal} \) | \( 15V \text{pk-pk nominal} \) |
| Input Frequency | \( 25kHz \text{nominal} \) | \( 25kHz \text{nominal} \) | \( 25kHz \text{nominal} \) | \( 25kHz \text{nominal} \) |

| POWER SUPPLY (AD202 Only) |        |        |        |        |
| Voltage, Rated Performance | 400\mu A Total | 400\mu A Total | N/A | N/A |
| Voltage, Operating | \( +15V \pm 5\% \) | \( +15V \pm 5\% \) | \( +15V \pm 5\% \) | \( +15V \pm 5\% \) |
| Current, No Load (\( V_s = +15V \)) | N/A | N/A | N/A | N/A |
| TEMPERATURE RANGE |        |        |        |        |
| Rated Performance | \( 0^\circ C \text{ to } + 85^\circ C \) | \( 0^\circ C \text{ to } + 85^\circ C \) | \( 0^\circ C \text{ to } + 85^\circ C \) | \( 0^\circ C \text{ to } + 85^\circ C \) |

| PACKAGE DIMENSIONS |        |        |        |        |
| 5-Pin SSOP (Y) | 2.08" x 0.250" x 0.625" | 2.08" x 0.250" x 0.625" | 2.08" x 0.250" x 0.625" | 2.08" x 0.250" x 0.625" |

### NOTES
- *Specifications apply to AD204J.*
- Nonlinearity is specified as % deviation from a best straight line.
- *1 0.0uf F min. demplugging required (see text).
- *2 Peak pulse (1us rise time).
- Specifications subject to change without notice.
INSIDE THE AD202 AND AD204

The AD202 and AD204 use an amplitude modulation technique to permit transformer coupling of signals down to dc (Figure 1a and 1b). Both models also contain an uncommitted input op amp and a power transformer which provides isolated power to the op amp, the modulator, and any external load. The power transformer primary is driven by a 25kHz, 15V p-p square wave which is generated internally in the case of the AD202, or supplied externally for the AD204.

Within the signal swing limits of approximately ±5V, the output voltage of the isolator is equal to the output voltage of the op amp; that is, the isolation barrier has unity gain. The output signal is not internally buffered, so the user is free to interchange the output leads to get signal inversion. Additionally, in multichannel applications, the unbuffered outputs can be multiplexed with one buffer following the mux. This technique minimizes offset errors while reducing power consumption and cost. The output resistance of the isolator is typically 3kΩ for the AD204 (7kΩ for AD202) and varies with signal level and temperature, so it should not be loaded (see Figure 2 for the effects of load upon nonlinearity and gain drift). In many cases a high-impedance load will be present or a following circuit such as an output filter can serve as a buffer, so that a separate buffer function will not often be needed.

(Circuit figures shown on this page are for SIP style packages. Refer to third page of this data sheet for proper DIP package pin-out.)
Figure 9 shows how zero adjustment is done at the output by taking advantage of the semi-floating output port. The range of this adjustment will have to be increased at higher gains; if that is done, be sure to use a suitably stable supply voltage for the pot circuit.

There is no easy way to adjust gain at the output side of the isolator itself. If gain adjustment must be done on the output side, it will have to be in a following circuit such as an output buffer or filter.

![Figure 9. Output-Side Zero Adjustment](image)

**Common-Mode Performance.** Figures 10a and 10b show how the common-mode rejection of the AD202 and AD204 varies with frequency, gain, and source resistance. For these isolators, the significant resistance will normally be that the path from the source of the common-mode signal to IN COM. The AD202 and AD204 also perform well in applications requiring rejection of fast common-mode steps, as described in the Applications section.

![Figure 10a. AD204](image)

**Dynamics and Noise.** Frequency response plots for the AD202 and AD204 are given in Figure 11. Since neither isolator is slew-rate limited, the plots apply for both large and small signals. Capacitive loads of up to 470pF will not materially affect frequency response. When large signals beyond a few hundred Hz will be present, it is advisable to bypass –VISO and +VISO to IN COM with 1μF tantalum capacitors even if the isolated supplies are not loaded.

At 50/60Hz, phase shift through the AD202/AD204 is typically 0.8°C (lagging). Typical unit – unit variation is ±0.2°C (lagging).

![Figure 11. Frequency Response at Several Gains](image)

(Circuit figures shown on this page are for SIP style packages. Refer to third page of this data sheet for proper DIP package pin-out.)
Current Transducers
0 to 50A, 0 to 100A

Miniature current transducers using Hall effect technology to accurately measure instantaneous value of ac or dc currents up to 50A or 100A, in total isolation from the circuit being monitored. The transducers can provide either current or voltage output, and are designed for PCB mounting. The output is linearly related to the primary current flowing through the centre core. The sensitivity of the transducers may be increased by increasing the number of times that the current carrying conductor passes through the centre core, e.g. 5 passes through a 50A transducer reduces the measuring range to 10A, and increases the resolution by a factor of 5. Housed in a black flame retardant Noryl case.

Also available is a version which provides a true RMS output option which for each nominal current rating (50A or 100A) gives an output of 1 volt full scale.

FOR SUITABLE POWER SUPPLIES, SEE PAGE 714

Mfrs. List No. LTA50P/SP1 = 107-803, LTA100P/SP1 = 107-804, LTA50PR = 280-008, LTA100PR = 280-010
DESCRIPTION

The CIO-DAS08 multifunction analog and digital I/O board is designed to be compatible with MetraByte's popular DAS08. Installed in any IBM PC/XT/AT/PS/30 or compatible computer the CIO-DAS08 turns your personal computer into a medium speed data acquisition and control station suitable for laboratory data collection, instrumentation, production testing, or industrial monitoring.

The CIO-DAS08 is supported by a broad range of software to allow programmed control in BASIC, C, FORTRAN and PASCAL. Many menu controlled data logging, analysis and control programs are available from a number of third party developers. In fact, any software designed for MetraByte's popular DAS-08 will work with the CIO-DAS08: we guarantee it!

In addition, the CIO-DAS08 comes with a complete PIO-12 compatible 8255 and 37 pin connector.

CIO-DAS08 CONNECTOR

- **Gain & Range Select Switch:** Accessible from outside PC
- **8 Analog Inputs:** 3, 16 Bit Counters
- **3 Digital In:** 4 Digital Out
- **37 Pin D Connector:** PIO-12 Compatible
- **100% DAS-08 Compatible**
The CIO-DAS 16 is supported by a broad range of software to allow programmed control in BASIC, C and PASCAL. Many menu controlled data logging, analysis and control programs are available from a number of third party developers. In fact, any software designed for MetraByte's popular DAS-16 will work with the CIO-DAS 16; we guarantee it! In addition, the CIO-DAS 16 comes with a complete CIO-DIO24 compatible 8255 and 37 pin connector!
CIO-SSH16
16 Channel Simultaneous Sample & Hold Accessory Board

16 DIFFERENTIAL INPUTS

INDIVIDUAL SWITCH SELECTABLE

GAINS OF:

1
10
100
300
500
600
700 & 800

16 SAMPLE & HOLD AMPLIFIERS.

LESS THAN 50 NANoseconds

APERTURE UNCERTAINTY

EACH CHANNEL IS

INDIVIDUALLY CALIBRATED

CALIBRATION SOFTWARE INCLUDED

FULLY SUPPORTED BY ComputerBoards AND THIRD PARTY SOFTWARE

DUE TO AUTOMATIC, HARDWARE TRIGGERING.

SCREW TERMINALS

FOR 12-22 AWG

37 PIN D TYPE CONNECTOR

MATES WITH ALL CIO-AD ANALOG INPUT BOARDS

DESCRIPTION

The CIO-SSH16 simultaneous sample and hold accessory acts as a front end signal amplification and capture for the CIO-DAS16 series of analog input boards.

There are two major functions on the board. Sixteen differential amplifiers have individual switch selectable gains of 1, 10, 100, 200, 300, 500, 600, 700 and 800 providing very flexible amplification of individual signals. After amplification, each channel has a sample and hold which is controlled by the CIO-DAS analog input board. The total aperture uncertainty for all 16 circuits is less than 50 nanoseconds.

The CIO-SSH16 eliminates the channel to channel skew associated with multiplexed A/D inputs. A fast A/D board sampling at 100,000 samples per second will exhibit a minimum channel to channel skew of 10 microseconds. Since the skew is additive from channel to channel, the 16 channel total scan skew is 160 microseconds. In applications where a number of signals must be analyzed and compared, such as high speed transient analysis and spectrum analysis, a channel to channel skew may be unacceptable.

Even low speed applications, such as oscillographic recording and display may require simultaneous sampling of all channels.

BLOCK DIAGRAM

There are 16 separate fully differential amplification and sample & hold circuit blocks on the SSH16. One block, channel 0, uses the sample & hold chip on the CIO-DAS16. The S&H Trigger line enters TRACK whenever the CIO-DAS16 enters TRACK on channel 0. When the CIO-DAS16 enters HOLD for channel 0 the entire SSH16 enters HOLD also. The SSH16 remains in HOLD mode while the CIO-DAS16 samples channels 1, 2, 3...N. All SSH16 acquisition runs begin with channel 0 and by taking advantage of the CIO-DAS16 S&H chip, the A/D conversion and transfer rate equals the maximum throughput of the CIO-DAS.
9 SPECIFICATIONS

9.1 POWER CONSUMPTION

+5V Supply: 107 mA typical / 180 mA max.
+12V Supply: 6 mA typical / 10 mA max.
-12V Supply: 10 mA typical / 16 mA max.

NOTE: Additional power will be drawn by user’s connections to the power pins accessible on CIO-DAS08 connectors.

9.2 ANALOG INPUTS

<table>
<thead>
<tr>
<th># Channels</th>
<th>Type</th>
<th>Resolution</th>
<th>Accuracy</th>
<th>Speed</th>
<th>Monotonicity</th>
<th>Linearity</th>
<th>Ranges</th>
<th>Overvoltage</th>
<th>Input Current</th>
<th>Input Impedance</th>
<th>Gain Temp. Coef.</th>
<th>Zero Drift</th>
<th>Gain Drift</th>
</tr>
</thead>
<tbody>
<tr>
<td>8, Single Ended</td>
<td>12 bits, 4095 divisions of full scale.</td>
<td>0.01% of reading +/- 1 bit.</td>
<td>Successive approximation</td>
<td>25 uSec - AD674</td>
<td>Guaranteed over operating temp.</td>
<td>+/- 1 bit</td>
<td>+/- 5 Volts</td>
<td>+/- 10 Volts</td>
<td>0 to 10 Volts</td>
<td>+/- 10 Volts/continuous</td>
<td>100 nA max @ 25 deg. C.</td>
<td>10 Meg Ohms</td>
<td>+/-FS +/- 25 ppm/deg C</td>
</tr>
</tbody>
</table>

9.3 SAMPLE & HOLD AMP.

<table>
<thead>
<tr>
<th># Channels</th>
<th>Type</th>
<th>Resolution</th>
<th>Accuracy</th>
<th>Speed</th>
<th>Monotonicity</th>
<th>Linearity</th>
<th>Ranges</th>
<th>Overvoltage</th>
<th>Input Current</th>
<th>Input Impedance</th>
<th>Gain Temp. Coef.</th>
<th>Zero Drift</th>
<th>Gain Drift</th>
</tr>
</thead>
<tbody>
<tr>
<td>8, Single Ended</td>
<td>12 bits, 4095 divisions of full scale.</td>
<td>0.01% of reading +/- 1 bit.</td>
<td>Successive approximation</td>
<td>25 uSec - AD674</td>
<td>Guaranteed over operating temp.</td>
<td>+/- 1 bit</td>
<td>+/- 5 Volts</td>
<td>+/- 10 Volts</td>
<td>0 to 10 Volts</td>
<td>+/- 10 Volts/continuous</td>
<td>100 nA max @ 25 deg. C.</td>
<td>10 Meg Ohms</td>
<td>+/-FS +/- 25 ppm/deg C</td>
</tr>
</tbody>
</table>

9.4 REFERENCE VOLTAGE OUTPUT

<table>
<thead>
<tr>
<th>Type</th>
<th>Reference Range</th>
<th>Temp. Coef.</th>
<th>Load Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>OP1 - OP4 low</td>
<td>+10 Volts +/- 0.1 V</td>
<td>50 ppm/deg C max.</td>
<td>2 mA max.</td>
</tr>
<tr>
<td>OP1 - OP4 high</td>
<td>2.7 V Min -0.4 mA current source</td>
<td>2.4 V min @ -200 uA</td>
<td></td>
</tr>
<tr>
<td>IP1 - IP3 low</td>
<td>2 V max @ 20 uA</td>
<td>-0.5 V min</td>
<td></td>
</tr>
<tr>
<td>IP1 - IP3 high</td>
<td>0.5 V max @ 2.5 mA</td>
<td>2.0 V min, 7 V max</td>
<td></td>
</tr>
<tr>
<td>8255 output low</td>
<td>0.5 V max @ 8 mA current sink</td>
<td>5 LSTTL loads</td>
<td></td>
</tr>
<tr>
<td>8255 output high</td>
<td>0.8 V Max</td>
<td>200 uA</td>
<td></td>
</tr>
<tr>
<td>8255 input low</td>
<td>2 V max @ 20 uA</td>
<td></td>
<td></td>
</tr>
<tr>
<td>8255 input high</td>
<td>0.5 V max @ 2.5 mA</td>
<td></td>
<td></td>
</tr>
<tr>
<td>8255 drive capability</td>
<td>2 V max @ 20 uA</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

9.5 DIGITAL I/O
### 7.9 SPECIFICATIONS

#### POWER CONSUMPTION

| +5  | 75 mA MAX, | 620 mA Typical |
| +12 | 21 mA MAX, | 8.4 mA Typical |
| -12 | 31 mA MAX, | 21 mA Typical |

#### ANALOG INPUT

| Channels | 8 Differential | 16 Single Ended |
| Resolution | 12 Bits | 1 part in 4095 |
| Accuracy | 0.01% of reading, ±1 bit |
| Input Range | ±10V MAX | ±0.25V MIN |
| Coding | Bipolar Offset binary |
| Overvoltage | ±25V continuous |
| Input Current | 250 mA MAX | 12 mA typ @ 25 deg C |
| Temp. Coeff. | ±25 ppm/deg C MAX |
| Conversion Time | AD16/50K | 15 uSec MAX |
| Model | AD16/100K | 8.5 uSec MAX |
| Linear | ±1 Bit |
| Zero drift | ±10 ppm/deg C MAX |
| Gain drift | ±30 ppm/deg C MAX |
| Vref Output | ±5V ±0.05V |

#### SAMPLE & HOLD AMP

| Acquisition time | 1 uSec to 1% of full scale step |
| Aperture u. | 0.3 uSec Typical |
| Model | HARRIS HA-2425 |
| Temp. Coeff. | ±30 ppm/deg C MAX |
| Load current | ±5mA MAX |

#### D/A CONVERTERS

| Channels | Two Independent |
| Type | 12 bit multiplying, double buffered |
| Linearity | ±1/2 Bit |
| Monotonicity | ±1/2 Bit |
| Output range | ±10V MAX |
| Output drive | ±15mA MIN |
| Output resistance | <0.1 Ohm |
| VREF Input range | ±10V |
| Full scale out | Gain = -1 \* VREF |
| Settling time | 30 uSec to 1% for full scale step |

#### PROGRAMMABLE TIMER

| Type | 82C54 |
| Counters | 3, 16 Bit down counters |
| XTAL | 1 or 10 MHz |
| Output Drive | 2.2mA @ 0.45V |
| Frequency | DC to 10MHz |
| Active count edge | Negative |
| Clock pulse width | 300Sec High, 50nSec Low |

#### DIGITAL I/O

- **Input/Output**: 24 bits, 255, three 8 bit ports
- **Input only**: 4 bits
- **Input low volts**: 74LS244
- **Input volts**: 74LS244
- **Input low volts**: 74LS244
- **Input high volts**: 74LS244
- **Output only**: 4 bits
- **Output volts**: 74LS244
- **Output low volts**: 74LS244
- **Output high volts**: 74LS244
- **Trigger**: On board
- **External**: User supplied TTL pulse

#### DMA A/D TRANSFER

- **Level**: 2, 7
- **Enable**: INTE Bit
- **Trigger**: CIO-DAS16 Control register
- **Termination**: Single Cycle
- **Pacing**: Internal
- **External**: User supplied TTL pulse

#### POWER OUTPUTS

Direct from PC bus

#### ENVIRONMENTAL

| Model | Normal Temp Model |
| Operating Temp | 0 to 50 Deg C |
| Storage Temp | 70 to 100 Deg C |
| Humidity | 0 to 90% non-condensing |
| Weight | 10.25 Oz |

| Model | Extended Temp Model |
| Operating Temp | -30 to 60 Deg C |
| Storage Temp | -165 to 150 Deg C |
| Humidity | 0 to 90% non-condensing |
| Weight | 10.25 Oz |
Appendix D: Specifications and data sheet of the test induction motor

### Nameplate Rating

<table>
<thead>
<tr>
<th>Nameplate Rating</th>
<th>AMIENT TEMP. 40 °C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type</td>
<td>Form</td>
</tr>
<tr>
<td>1K</td>
<td>F5XW</td>
</tr>
<tr>
<td>Volts</td>
<td>415</td>
</tr>
<tr>
<td>Amperes Full Load</td>
<td>14.4</td>
</tr>
<tr>
<td>Hs</td>
<td>50.1</td>
</tr>
<tr>
<td>Poles</td>
<td>d</td>
</tr>
<tr>
<td>Full Load Solution</td>
<td>1440</td>
</tr>
<tr>
<td>Secondary Volts</td>
<td>1060</td>
</tr>
<tr>
<td>Amperes</td>
<td>14.5</td>
</tr>
<tr>
<td>Volts</td>
<td>505</td>
</tr>
<tr>
<td>Amperes</td>
<td>800</td>
</tr>
</tbody>
</table>

### Test Characteristics

<table>
<thead>
<tr>
<th>No Load Test</th>
<th>50: Hz</th>
<th>Locked Rotor Test</th>
<th>50: Hz</th>
<th>Locked Rotor Test</th>
<th>25: Hz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Volts</td>
<td>Amperes</td>
<td>Watts</td>
<td>Amperes</td>
<td>Watts</td>
<td>Amperes</td>
</tr>
<tr>
<td>415</td>
<td>5.82</td>
<td>400</td>
<td>14.5</td>
<td>84</td>
<td>1060</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

### Load Characteristics Data

<table>
<thead>
<tr>
<th>Load (%)</th>
<th>Current (A)</th>
<th>Efficiency (%)</th>
<th>Power Factor (%)</th>
<th>Slip (A)</th>
<th>Breakdown Torque (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>5.82</td>
<td>6.48</td>
<td>8.47</td>
<td>10.82</td>
<td>13.85</td>
</tr>
<tr>
<td>25</td>
<td>6.48</td>
<td>8.47</td>
<td>10.82</td>
<td>13.85</td>
<td>17.14</td>
</tr>
<tr>
<td>50</td>
<td>8.47</td>
<td>10.82</td>
<td>13.85</td>
<td>17.14</td>
<td>Breakdown Torque: 0.52</td>
</tr>
<tr>
<td>75</td>
<td>10.82</td>
<td>13.85</td>
<td>17.14</td>
<td>25</td>
<td>Locked Rotor Current (A): 69.6</td>
</tr>
<tr>
<td>100</td>
<td>13.85</td>
<td>17.14</td>
<td>25</td>
<td>125</td>
<td>Locked Rotor Torque: 888</td>
</tr>
<tr>
<td>125</td>
<td>17.14</td>
<td>25</td>
<td>125</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

### Temperature Rise (Deg. C) (Room 22.8 °C)

- No Load: 3.75 Hz 415 Volt 100 Load % 54.3 LS 37 33 Stator: 0.9447 Ohms
- Locked Rotor: 54.3 LS 37 33 Stator: 0.9447 Ohms

### Insulation Resistance

<table>
<thead>
<tr>
<th>Stator</th>
<th>Megohms</th>
<th>Megger Voltage</th>
<th>AC 60Hz for 1 Min</th>
<th>Air gap (in.)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1000</td>
<td>500</td>
<td>1900</td>
<td>0.4</td>
</tr>
<tr>
<td>Rotor</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

### Vibration (P-P)

- Stator: 6.0 µm
- Rotor: 6.0 µm

### Noise

- Stator: 60.6 dB (A)
- Rotor: 60.6 dB (A)

### Remarks

- Over Speed 120%, 2 min. .... Satisfactory
- Momentary Excess Torque 160%, 15 sec. .... Satisfactory
THREE PHASE INDUCTION MOTOR

7.5 kW

Rating

<table>
<thead>
<tr>
<th>Type</th>
<th>Form</th>
<th>Volts</th>
<th>Amperes Full Load</th>
<th>Hr</th>
<th>Poles</th>
<th>Full Load Speed, r.p.m.</th>
<th>Secondary Volts</th>
<th>Ampere</th>
<th>Insulation Class</th>
<th>Time Rating</th>
<th>Code</th>
<th>Frame No.</th>
</tr>
</thead>
<tbody>
<tr>
<td>EX</td>
<td>FBGN</td>
<td>415</td>
<td>14.4</td>
<td>50</td>
<td>4</td>
<td>1440</td>
<td>F</td>
<td>PCR</td>
<td></td>
<td></td>
<td></td>
<td>D132M</td>
</tr>
</tbody>
</table>

![Speed vs. Torque Curve (A)]

![Speed vs. Current Curve (B)]

Current (Amps)

Slider
Appendices

TOSHIBA CORPORATION

7.5 KW THREE PHASE INDUCTION MOTOR

LOAD CHARACTERISTIC CURVE

415 V 50 Hz

Current (Amp.)

Eff. P.I. (%)

Load (%)

Eff.

P.F.

SLIP

CURRENT

SLIP (%)
Appendix E: Calculation of Induction Motor Variable Losses

It is known that the developed torque on the motor shaft, $T$, can be expressed as:

$$ T = \frac{I_2^2 R_2}{s} $$  \hspace{1cm} (E.1)

where $I_2$ is the rotor current referred to the stator and represents the load, $R_2$ is the rotor resistance referred to the stator and $s$ is the rotor slip. According to induction motor approximate equivalent circuit (Figure 2.2):

$$ I_2 = \frac{V_1}{R_1 + \frac{R_2}{s} + jX} $$  \hspace{1cm} (E.2)

where $V_1$ is the motor input voltage, $R_1$ is the stator winding resistance and $X$ is the total machine leakage reactance. Under normal operating conditions where $V_1 = 1$ pu and $s$ is very small (about 0.04 for the test motor), it can be written:

$$ I_2 \approx \frac{s}{R_2} $$  \hspace{1cm} (E.3)

Equation (E.1), then, can be re-written as:

$$ T \approx I_2 $$  \hspace{1cm} (E.4)

which describes a linear relationship between $T$ and $I_2$.

Using Equation (E.2), the rotor slip, $s$, can be calculated as:

$$ s = -\frac{R_2}{\sqrt{\frac{1}{I_2^2} - X^2 - R_1}} $$  \hspace{1cm} (E.5)

which varies almost linearly with $I_2$ from $s \approx 0$ at $I_2 = 0$ (no-load condition) to $s = s_0$ at $I_2 = 1$ pu (full load conditions) where $s_0$ can be calculated as:
\[ s_0 = \frac{R_2}{\sqrt{1 - X^2 - R_i}} \] (E.6)

and hence:

\[ s = s_0 I_2 \] (E.7)

The phase angle, \( \Phi \) between the stator input voltage, \( V_1 \), and load current, \( I_2 \), can be calculated as:

\[ \Phi = \tan^{-1} \left( \frac{X}{R_1 + \frac{R_2}{s}} \right) = \tan^{-1} \left( \frac{X}{R_1 + \frac{R_2}{s_0 I_2}} \right) \] (E.8)

which is a linear function of \( I_2 \) as

\[ \Phi \approx \Phi_0 I_2 \] (E.9)

with

\[ \Phi_0 = \tan^{-1} \left( \frac{X}{R_1 + \frac{R_2}{s}} \right) \] (E.10)

The following equation can be written regarding the current:

\[ I_i^2 = I_m^2 + I_2^2 + 2I_2 I_m \sin \Phi \] (E.11)

where \( I_1 \) is the stator current and \( I_m \) is the magnetising current. Under normal operating conditions \( \Phi \) is very small and hence \( \sin \Phi \approx \Phi \) (rad) \( \approx \Phi_0 I_2 \). Therefore, Equation (E.11) is modified as:

\[ I_i^2 = I_m^2 + I_2^2 + 2I_2 I_m (I_2 \Phi_0) \\
= I_m^2 + I_2^2 (1 + 2I_m \Phi_0) \] (E.12)

Fundamental copper losses in the motor, \( W_{cu} \), can be expressed as:
\[ W_{cu} = I_m^2 R_1 + I_2^2 (1 + 2I_m \Phi_0) R \]
\[ = W_{\text{const}} + W_{\text{load}} \]  
(E.13)

where \( R \) is the total machine resistance \( (R = R_1 + R_2 + R_{\|}) \), \( W_{\text{const}} \) is the constant part of the copper losses and \( W_{\text{load}} \) is the load dependent part of the copper losses. Since iron losses and windage and friction losses are almost constant at nominal voltage, they can be assumed as part of \( W_{\text{const}} \). Therefore, total machine losses, \( W_{\text{total}} \), can be approximated as:

\[ W_{\text{total}} = W_{\text{const}} + W_{\text{load}} \]  
(E.14)

This Equation is used to derive a derating factor for induction motors as described in Chapter 6.