Loss Minimization Control of
Interior Permanent Magnet Motor Drives

By: Sadegh Vaez

A thesis submitted to the
Department of Electrical and Computer Engineering
in the conformity with requirements for the
Degree of Doctor of Philosophy

Queen's University,
Kingston, Ontario, Canada

March 1997

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0-612-22499-6
Abstract

In this thesis a novel on-line adaptive loss minimization control strategy for electric motor drives is introduced. Based on this strategy an adaptive loss minimization controller (ALMC) is proposed for inverter-fed interior permanent magnet (IPM) motor drives after the minimum loss operation of one of these motors is analyzed. The ALMC provides a new pattern of change in d-axis stator current to achieve a minimum drive input power at any operating condition. Gaining insight from the analysis of the motor speed variations in response to the changes in d-axis current, a concept of forced compensation is introduced. Using this concept a non-linear speed controller (NLSC) is proposed to achieve desirable motor dynamics in transient state and maintain the output power constant in steady state while the input power is being reduced by the ALMC. The harmonized operation of the ALMC and the NLSC results in a smooth and fast system performance thus overcoming the major drawbacks of other on-line loss minimization control approaches like the torque pulsations and a long search time to reach a minimum input power. The proposed loss minimization strategy provides more energy saving in comparison with other methods and extends the application of loss minimization control to a new class of motor drives requiring an efficient and smooth operation in the face of frequent changes in the operating point like in electric vehicles. The analysis, design, simulation, DSP implementation and extensive test results of a current vector control system including the proposed ALMC and NLSC, when applied to an experimental 1 hp IPM motor drive are presented in detail. The simulation confirmed by the experimental results proves the validity of this new loss minimization control strategy.
Acknowledgement

I would like to express my sincere thanks and appreciation to Professor V. I. John for his continuous and valuable advice, suggestions and encouragement throughout the course of my PhD program. I am also much indebted to Professor M. A. Rahman for kindly providing me the excellent opportunity to carry out the experimental part of this work at his lab at the Memorial University of Newfoundland, thus enjoying his superb expertise, advice and encouragement. I am grateful to Professor G. E. Dawson for giving me the chance to use the extensive computing facilities during different stages of my research and to Professor P. C. Sen for introducing the author to drives and for his generosity in providing me timely access to many important references in the field.

Thanks are also due to Dr. A. Hoque for his sincere help and co-operation during the experimental phase of this work and to Messers R. Newman, D. Guy and D. Johnson of the Memorial University for offering an excellent technical support. I wish to extend my thanks to my colleagues.

Lastly I would like to record my gratitude to my wife for her encouragement, understanding and patience during many long days required for the preparation of this thesis.
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List of Symbols

ALMC  adaptive loss minimization controller
B     viscous coefficient
f     inverter output frequency
f_c   carrier wave frequency
i_a, i_b, i_c  3-phase currents
i_d   direct axis stator current in synchronously rotating reference frame
i_q   quadrature axis stator current in synchronously rotating reference frame
i_d'  direct axis stator current command in rotating reference frame
i_q'  quadrature axis stator current command in rotating reference frame
i_{od}  torque producing component of i_d
i_{oq}  torque producing component of i_q
i_{ed}  direct axis component of iron loss current
i_{eq}  quadrature axis component of iron loss current
J     rotor inertia constant
K     nonlinearity gain of NLSC
K_{T_d}  direct axis torque constant
K_{T_q}  quadrature axis torque constant
L  load inductance
L_d  direct axis inductance
L_q  quadrature axis inductance
NLSC  nonlinear speed controller
p  derivative operator
P  number of pole pairs
P_{DC}  inverter input power
P_{motor}  motor input power
R  load resistance
R_s  or R_\alpha  stator phase resistance
R_c  iron loss resistance
T_c  carrier wave period
T_e  motor developed torque
T_L  load torque
T_s  sampling time
v_a, v_b, v_c  3-phase voltages
v_a^*, v_b^*, v_c^*  3-phase voltage commands
v_d  direct axis voltage in synchronously rotating reference frame
v_q  quadrature axis voltage in synchronously rotating reference frame
v_d^*  direct axis voltage command in rotating reference frame
v_q^*  quadrature axis voltage command in rotating reference frame
W_{E}  electrical loss
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>$W_{Fe}$</td>
<td>motor iron loss</td>
</tr>
<tr>
<td>$W_{Cu}$</td>
<td>copper loss</td>
</tr>
<tr>
<td>$\eta$</td>
<td>efficiency</td>
</tr>
<tr>
<td>$\theta_e$</td>
<td>electrical angle, rotor position</td>
</tr>
<tr>
<td>$\theta_m$</td>
<td>mechanical angle, rotor position</td>
</tr>
<tr>
<td>$\psi_a, \psi_b, \psi_c$</td>
<td>3-phase flux linkages</td>
</tr>
<tr>
<td>$\psi_d$</td>
<td>direct axis flux linkage</td>
</tr>
<tr>
<td>$\psi_q$</td>
<td>quadrature axis flux linkage</td>
</tr>
<tr>
<td>$\psi_m$</td>
<td>magnet flux</td>
</tr>
<tr>
<td>$\omega_e$</td>
<td>electrical angular frequency</td>
</tr>
<tr>
<td>$\omega_m$</td>
<td>mechanical angular frequency</td>
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CHAPTER 1

Introduction

Loss minimization control of motor drives has been around for more than two decades now [1]. The methods used for this purpose have evolved considerably during this period. They range from a simple voltage control procedure of Nola [2] to more sophisticated intelligent on-line approaches. These methods, traditionally applied to DC and induction motors, have been effective in enhancing the motor efficiency and in saving electric energy. With increasing energy cost and the growing demand for stricter control of environment, there is a consensus among the power utilities, the manufacturers of electric devices and the consumers to improve the energy efficiency of electric devices [3]. In this perspective the approaches to loss minimization control of electric motors, as the major consumer of electric energy, seem more needed than ever before.

While the state of the art loss minimization controllers are capable of handling the requirements of many motor drive applications, there are many more situations in which the limitations and drawbacks of such controllers prevent or reduce their usefulness. In many applications the operating conditions, i.e. motor speed and load, vary over a wide range. Also system parameters, the motor inductances and resistances for instance, change considerably depending on the operating condition, ambient temperature, etc.
These variations and changes make it difficult or useless to apply the existing loss minimization control methods. Some methods are not fast enough to cope with the rapidly changing motor speed or torque, others may suffer from parameter dependency and need for very complicated mathematical models of the system. This work is an attempt to overcome some of these difficulties by making the practice of loss minimization control more effective. At the same time it is the objective of this work to broaden the application of loss minimization control to include new areas of motor drives.

The electric vehicle (EV) as an application of motor drives is a potential area for loss minimization control. On the other hand the EV is a perfect example of systems with difficulties in applying the existing loss minimization control methods. The EV speed and load torque vary frequently depending on the road condition and driver’s desire. Motor drive parameters are affected considerably by the changing operating point and the environmental conditions. A fast, smooth and parameter insensitive loss minimization controller is needed to meet the requirement of such an application. In this work the EV is chosen as the application focus to achieve the objectives mentioned above. The EV is briefly reviewed in the first section of this chapter. Different motors used in the EV are considered next on a comparison basis, while highlighting the features of interior permanent magnet motors as a suitable choice for application in EVs. Finally the efficiency of electric vehicle motor drives is discussed and the need for a new loss minimization control strategy suitable for the applications like EVs is emphasised.
1.1 Electric Vehicles

The electric vehicle (EV) has its origin back to about a century ago when EVs had the upper hand in the new born automotive industry. This situation did not last long. The internal combustion engine (ICE) won the race shortly afterwards and has been holding its position as the sole mainstream of the car making industry since then [4]-[7]. The EV has not been able to penetrate the market in spite of many attempts in the 60s and the early 80s [8]. The motivation for these attempts were the air quality concerns and the sudden increase in the oil prices. The technological barriers and economical shortcoming were mainly responsible for the EV failures. The high maturity of ICE technology also had been effective in limiting any competition from EVs.

A renewed interest in the electric vehicles and hybrid vehicles has been revived in recent years which seems to be more serious and more fruitful than ever before [9]. The motivation this time is mainly the environmental concern in the light of public awareness and governmental and industrial response. EVs when commercialized can reduce air pollution in urban areas considerably. The only emission they produce is the one at the power plants. From the economical point of view EVs are the most flexible fuel vehicles. The diversity of the fuels depends on the kinds of power plants. While almost all ICE cars use oil, many power plants use other energy resources which are either less scarce and cheaper or cleaner. EVs are more efficient when driven in cities with heavy traffic since the electric motors do not use energy during stops [6].

The response of the auto industry to this new wave of EV revival has been overwhelming. Almost all major car makers all over the world [9], in Europe [10], North
America [11] and the far east [12]-[13] have programs on the EV in hand. Hundreds of millions of dollars are being spent on EV R&D globally each year [5]. A number of major improvements like fast battery charging techniques [14], high power-high efficiency-light weight motors and controllers [10] have been reported. During the nineties many prototype EVs have been built and tested with remarkable improvements in design and performance [6], [10], [14]-[16]. A major auto manufacturer has initiated the mass production of EVs just a few months ago [17]. Others are expected to follow soon.

Tough laws concerning air quality and protection of the environment from pollution and government incentives to encourage the production of zero or minimum emission cars add a new dimension to the issue [11]. One argument is that if the EV protects the society’s health and reduces harm to the human’s life why should not the public wealth be spent for it just as health care expenses. These laws and incentives compensate partially the higher cost of EVs and greatly affect the final conclusion about the EV future. This is the new aspect of the problem which might assure economic viability, although the long term perspective mainly depends on the EVs technological improvements.

However, there are major technological obstacles yet to be overcome before EVs enjoy wide public acceptance and gain a noticeable market share [18]. The EV range is not sufficient and the performance can not compete with the one easily obtained by the ICE cars [4], [9]. It is still expensive and non competitive to manufacture an electric car. Much work is needed to overcome these and other problems [19]-[20].
1.2 The EV Motor Drive

The EV technology is multidisciplinary since the EV is a harmonized system consisting of battery, charger, propulsion system, body, control system, etc. The electric propulsion system is the heart of an EV. It includes the motor drive and its control. The motor selection for the EV is a challenging task due to the wide and sometimes contradicting specifications which are required for high energy efficiency and desirable performance. The basic requirements of motor configuration include [4]:

- High torque capability over the range 0-25% of maximum rated speed and adequate torque at high speed.
- Constant rated power output capability over the range of 25-100% of maximum rated speed.
- High propulsion energy efficiency while operating in a representative vehicle driving cycle.
- Capability of efficient regenerative braking.
- Good transient and steady state performance. This means a rapid, smooth and accurate control of motoring and braking at all operating speeds, independent of separate control of power source and ensuring a wide range of speed and torque control including adjustment by the monitoring of acceleration, deceleration and braking.
- High ratio of rated output power to total system weight and high torque/current ratio.
- Good reliability including:
  - ruggedness and simplicity
- capability of working in a harsh environment (temperature, vibration, dust, oil, moisture, etc.)

- capability of working in the case of partial failures.

- Lower voltage as the battery desires.

- Lower EMI and harmonics

- Low cost

It is extremely difficult to incorporate all these requirements in a single motor-drive system. While many types of motors (DC, IM, PM, etc.) have been applied in EVs, or specially designed for this purpose, there is no comprehensive comparison among them regarding the above requirements [21]. In a limited perspective Table 1 can give a sense of comparison for motors and their drivers [22]. Based on Table 1 IM’s have moderate priority in most features and PM motors have the maximum number of first priorities. Table 2 presents a quantitative comparison among motors and drivers for different kinds of systems [16].

Table 1

<table>
<thead>
<tr>
<th>Type of systems</th>
<th>DC</th>
<th>AC</th>
<th>PM</th>
</tr>
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<tbody>
<tr>
<td>Cost, mass production</td>
<td>2</td>
<td>1</td>
<td>3</td>
</tr>
<tr>
<td>Cost, limited production</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>Weight</td>
<td>3</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Volume</td>
<td>3</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Efficiency</td>
<td>3</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Robustness</td>
<td>2</td>
<td>1</td>
<td>3</td>
</tr>
<tr>
<td>Adaption to transmission</td>
<td>3</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Complexity of control</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
</tbody>
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Priority order: 1, 2, 3
Table 2

<table>
<thead>
<tr>
<th>DRIVES</th>
<th>DC (sep.)</th>
<th>IM</th>
<th>PM</th>
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<tr>
<td>Motor cost</td>
<td>high</td>
<td>low</td>
<td>medium</td>
</tr>
<tr>
<td>Motor efficiency (drive efficiency)</td>
<td>low 86%</td>
<td>low-med 90%</td>
<td>high 93%</td>
</tr>
<tr>
<td>Reg. mainten.</td>
<td>yes</td>
<td>no</td>
<td>no</td>
</tr>
<tr>
<td>Reliability</td>
<td>low</td>
<td>high</td>
<td>high</td>
</tr>
<tr>
<td>Durability</td>
<td>low</td>
<td>high</td>
<td>high</td>
</tr>
<tr>
<td>Motor size (litre/KW)</td>
<td>large 0.55</td>
<td>medium 0.35</td>
<td>small 0.14</td>
</tr>
<tr>
<td>Power density (KW/Kg)</td>
<td>low 0.42</td>
<td>medium 0.70</td>
<td>high 1.00</td>
</tr>
<tr>
<td>Controller</td>
<td>chopper</td>
<td>inverter</td>
<td>inverter</td>
</tr>
<tr>
<td>No. of devices</td>
<td>1-2</td>
<td>6</td>
<td>6</td>
</tr>
<tr>
<td>Contr. cost</td>
<td>low</td>
<td>high</td>
<td>high</td>
</tr>
<tr>
<td>Sensors</td>
<td>current</td>
<td>speed current</td>
<td>position current</td>
</tr>
</tbody>
</table>

There is no agreement among the specialists regarding the suitable motor for EV application. However, a clear trend is the shift from DC machines to AC ones as is the case in many variable speed drive systems [8], [21]. Induction motors are being used extensively due to low cost, robustness and maturity of construction. They also have advantages of high overload capability and low torque ripple. However, induction motors suffer from low power density, large size and low/medium efficiency [23].

Permanent magnet (PM) motors have recently emerged as the main competitor to induction motors for the EV applications due to a number of advantages i.e. potential high efficiency, high power density and small size [10], [24]-[28]. Permanent magnet motors traditionally are built in small sizes and low power ratings. They are extensively
used in control systems and servo mechanisms. With the introduction of high energy rear earth permanent magnet materials like Samarium-Cobalt (Sm-Co) and Neodymium-Iron-Born (Nd-Fe-B), improvement in power electronic devices like high power transistors and the advent of new control theories and techniques using powerful microcomputer hardware, the permanent magnet motors are penetrating the area of drive systems aggressively. These machines are undertaking a major market expansion as can be seen from the chart in Fig. 1.1 [29].

Fig. 1.1 Trends in the permanent magnet motor markets [29]
Two classes of PM motors seem to be more appealing to the EV propulsion system. The first one is the brushless DC motor (BLDCM) with the magnets mounted on the rotor surface. The motor is fed by a rectangular current wave form. The surface mounted magnets in BLDCM reduce its reliability and limit the maximum speed due to the limited strength of the adhesive material being used in mounting the magnets on the rotor surface. BLDCM in the original configuration is not capable of operating over a wide speed range over the constant power region. This is because of the nearly uniform effective air gap and the lack of sufficient rotor saliency. A version of BLDCM has been developed for the EV applications which uses double stator windings [30]. These windings can be connected in series or in parallel to produce high torque or high speed respectively [30]. The motor is used in a prototype EV [10].

More recently the buried or interior permanent magnet (IPM) motor with high energy magnets is becoming increasingly acceptable as the favourite choice for EV's [24]-[28]. The high efficiency of permanent magnet motors is combined with less torque ripples and a higher reliability due to non projecting magnets. But the most appealing feature of this machine is the special rotor configuration which results in a very strong saliency. This provides an excellent flux weakening capability and a wide range of high speed operation suitable for variable speed drives [31]. Fortunately the machine is capable of handling this high speed since the magnets are buried inside the rotor. It is usually supplied by a sinusoidal source and causes less torque ripple than BLDCM. Both the reluctance torque and the magnet alignment torque contribute to the total torque. The machine has also high temperature capability. Because of these advantages the IPM motor better fulfils the major requirements of EVs mentioned above. Therefore it is
chosen as the preferred motor for EV use in this research.

Fig. 1.2 shows the cross section of an IPM motor [26]. The stator is similar to that of induction motors. The rotor has a particular configuration. The flat rectangular shaped pieces of permanent magnets are buried inside the rotor body. They are alternately poled and radially magnetized. The permeance to the q-axis is much higher than that of the d-axis. As a result the machine has considerable rotor saliency and $L_q$ is much higher than $L_d$. The reluctance torque depends on the saliency coefficient $L_q/L_d$ or on the air gap reactance ratio $X_q/X_d$. Flux weakening is achieved by applying a negative d-axis stator current. It produces a d-axis armature flux which opposes the magnet flux. This provides the high speed constant power operation regime.

Fig. 1.2 Cross section of an IPM motor [26]

Parameter determination methods, modelling and performance characteristics of the motor can be found in the literature [32]-[41]. The flux weakening operation of the machine is discussed in several papers [42]-[47]. The investigation of the machine core loss by using the finite element calculation and measurement is reported by Schiferl and
Lipo [48]. Using this result an equivalent circuit analysis of the machine, taking into account the saturation effect, is presented by Consoli et al. [49]-[50]. Control of permanent magnet motors with rotor saliency as variable speed drives is studied by a number of researchers. This work is reported in [51]-[63]. Finally the loss minimization control of a permanent magnet motor with rotor saliency is discussed and a DSP control implementation is presented in [64]. A number of papers mentioned above will be reviewed later in more details.

1.3 Efficiency of EV Motors

The EV power supply capacity is limited to the energy stored in the battery. The battery supplies the propulsion system as well as all other energy consuming devices like water pump, oil pump, power brake and steering, air conditioner, etc. High efficiency operation of all these devices is essential for achieving a long travel range without sacrificing too much of the EV performance. Since most of the energy is delivered to the electric motor, high efficiency of the motor is a crucial task in EV design. Higher efficiency contributes to a longer range for a certain battery or to a lighter battery for a specific range. It also reduces the number of charge cycles and extends the battery life. This in turn results in a saving in the battery replacement cost which is the major life cycle cost of the EV.

Motor losses are a major factor in limiting the motor torque production capability. This factor is more important in permanent magnet machines where the high temperature may cause demagnetization of the permanent magnet material. This is especially true in
the case of Neodymium-Iron-Boron which has a relatively low transformation temperature. This magnet demagnetization may cause further reduction in the efficiency [65]. Less motor losses means less heat production and thus higher torque capability for the machine or a smaller and lighter machine for the same torque. Higher losses also cause more heat in the power wiring. It may also effect the inverter rating, size and weight.

While modern motors used in EVs usually have high efficiency over a range of operating conditions (usually around the rated torque and the rated speed), the efficiency drops at other conditions. It is a real challenge to obtain a flat efficiency curve over the required torque speed characteristic of variable speed drives as in the EV application.

Table 3 shows the constant speed range and the driving cycle range for a number of EVs built in 90's [9]-[10], [16], [66].

<table>
<thead>
<tr>
<th>VEHICLE</th>
<th>DEVELOPER</th>
<th>RANGE at constant speed</th>
<th>RANGE over driving cycle</th>
</tr>
</thead>
<tbody>
<tr>
<td>FEV</td>
<td>Nissan</td>
<td>250 Km (40 Km/h)</td>
<td>100 Km (city driving)</td>
</tr>
<tr>
<td>BMW-E1</td>
<td>BMW</td>
<td>215 Km (at 50 Km/h)</td>
<td>143 Km (FTP75-cycle)</td>
</tr>
<tr>
<td>PSA Citela</td>
<td>Peugeot</td>
<td>210 Km (at 40 Km/h)</td>
<td>110 Km (city driving)</td>
</tr>
<tr>
<td>Ecoster</td>
<td>Ford</td>
<td>320 Km (at 40 Km/h)</td>
<td>160 Km (FUD)</td>
</tr>
</tbody>
</table>

It is evident from the above table that the EV range decreases substantially over the driving cycle operation for all the EVs. Several factors contribute to this range reduction...
including the low efficiency of electric motors at low and high speeds. A loss minimization control strategy is capable of increasing the EV motor drive efficiency at different operating conditions, provided it meets the challenging requirements of the EV. The control strategy must be parameter insensitive meaning that a minimum loss operation is achieved regardless of a wide variation in the system parameters as is the case in IPM motor drives. The strategy should also be able to work under frequent changes in the operating conditions as in an EV.
CHAPTER 2

Loss Minimization Control of Electric Motor Drives

In this chapter the currently available loss minimization control methods applied to electric motor drives are reviewed and the objectives of the present work are outlined. Different methods of loss reduction in the electric motor drives are classified in the first section. The literature review of loss minimization control of motor drives, the main concern of this thesis, will be carried out next. An extensive review of loss minimization control applied to DC, Induction and PM motors are presented. Finally the objectives of the thesis are clearly defined after concluding the drawbacks and limitations of current methods.

2.1 Methods of Loss Reduction

The reduction of losses in electric motor drives can be achieved by different methods including:

1. Motor selection by matching closely the application requirements and the motor specifications. For an electric vehicle for instance it is important to consider the single or multiple motor drive system, fixed gear or shifting gears, maximum torque needed,
maximum voltage available from the battery and so on.

2. Design the motor for low loss (efficiency motor). It is achieved by design optimization methods and by using more and higher quality materials. For instance the use of more copper to reduce copper loss and the use of low hysteresis laminated steel cores to reduce iron losses [67]-[69].

3. Improvement of voltage and current waveforms of the motor power supply to reduce harmonic losses. Waveform shaping techniques are used to produce desirable voltage and current supplied to ac motors by inverters [70]-[72].

4. Loss minimization control i.e. the operation of motors at the minimum loss condition. It is based on the fact that each operating point (speed and torque) can be obtained by many combinations of independent motor variables like voltage and current giving different levels of total loss. The one which results in the minimum loss is chosen. This method of loss minimization has received a lot of attention from researchers for DC motors and induction motors. F. Nola in particular received overwhelming response for the loss minimization of induction motors by power factor control. He adjusted the applied voltage as a function of load [2]. However work in this area for permanent magnet motors has been rare. One of the objectives of this work is to study the potential of this method for permanent magnet motors.

Loss minimization control so far has been researched in two ways. One is the optimal control that minimizes the energy losses for a closed cycle or over a pre specified speed profile (transient loss minimization). The other is loss minimization at every operating point of the torque-speed characteristic (steady state loss minimization). Fig. 2.1 summarizes the loss minimization methods as described above.
2.2 Literature Review of Loss minimization Control

DC and Induction Motors

Several investigations have been reported in recent years on the loss minimization of motors working on fixed cycles. A. M. Trzynadlowski applied calculus of variation to minimize copper losses of a separately excited DC motor with constant load over a limited rotation of the rotor in a specified time interval [73]. The best speed profile was found first. The optimal field current was then calculated over the speed profile in such a way to minimize the copper losses. These currents were stored in a look up table and used as the current command. The iron losses were neglected in this study. R. D. Lorenz
et al. presented a loss minimization control for a field orientation induction motor during a closed-cycle operation [74]-[75]. Dynamic programming approach was used to minimize the copper and iron losses. The total loss was formulated as a cost function. The limits on the motor flux, speed, voltage and current were considered as the constraints of the dynamic programming problem. The optimal flux and the flux producing d-axis current trajectories over the cycle were obtained by solving the problem. These trajectories were approximated in closed forms as functions of time for real time implementation. The approximate expression depends on the total travelling time as a parameter. This parameter can be modified in real time to accommodate varying positioning requirements. A microcomputer implementation of the control system was also presented. In short travel cycles the method was not effective since the flux could not vary as quickly as desired. P. Famouri et al. proposed the control of DC motors for high efficiency under accelerating condition [76]. They worked on a DC series motor used for traction purposes. The machine model was linearized first by a nonlinear feedback control. The copper loss was then formulated as a quadratic performance index. The optimal motor current was found in closed form as a function of time by applying the Pontryagin’s Maximum Principle. Simulation results presented to show the reduction of losses in comparison with conventional control. The iron loss was not considered.

Steady state loss minimization control has mainly been reported in the literature by two methods [1]:

a) Feedforward controller- In this method, also referred to as the off-line loss minimization control, the control signal(s) corresponding to minimum loss condition are
found for each pair of speed and torque from a set of equations and drive parameters programmed into the controller. It is also possible to precalculate the optimal control signal(s) and store them in a memory as a look up table for a range of operating conditions. The controller then reads the right control signal corresponding to the present speed and torque and applies it to the drive. A. Kusko et al. described this kind of controller for both DC motors and induction motors. In the case of DC motors the total loss is formulated as a function of field current $I_f$, armature current $I_a$, speed $n$, torque $T$ and the motor parameters. By solving the derivative of the loss expression with respect to $I_f$ the optimal expression for $I_f$ is obtained as a function of $n$, $T$, and the motor parameters. The controller calculates torque by using the measured values of $I_f$, $I_a$ and the speed. Then it calculates the optimal $I_f$. This procedure is repeated continuously to ensure the operation of the motor at the minimum loss condition at all times. In the case of induction motors the total loss is formulated as a function of $T$, $n$, rotor frequency $f$ and the motor parameters. The $f$ corresponding to the minimum loss is obtained by solving the derivative of the loss with respect to $f$. If a linear motor model is used, the $f$ does not depend on $T$. Based on this assumption a simplified control scheme was proposed in [1]. The controller continuously measures the speed and calculates $f$ to obtain minimum loss operation. The speed regulator uses voltage to adjust the torque and maintain the speed independent of the rotor frequency control. An optimal efficiency control algorithm for an induction motor supplied by a current source inverter was suggested by H. G. Kim et al. [77]. It was shown that any torque and speed pair can be obtained by many combinations of rotor frequency (or slip) and flux, each corresponding to a different level of loss. At light loads, the motor flux was reduced to decrease the
iron loss and to improve the power factor. This resulted in an increase in the stator current and the copper loss to maintain the torque constant. However the total losses decreased. The required relationship between the stator current and the rotor frequency for the minimum loss was obtained numerically. A control loop was suggested by minor modification to the constant flux control scheme. Using this method more than ten percent improvement in the efficiency was reported at a quarter of full load. The dynamic response of the control system was also considered by small signal study. The non-ideal effects like saturation and harmonic losses were ignored. S. Funabiki et al. presented the high efficiency control of DC shunt motors [78]. They considered the magnetic saturation and armature reaction in modelling the motor loss by approximation of the no load characteristic. An off-line numerical search method was applied to find the set of optimal $I_r$ and the armature current $I_a$, minimizing the loss for each speed and torque. Maximum values of $I_r$ and $I_a$ were considered as the constraints for the numerical search for the optimal current. The current commands for discrete values of torque and speed were stored in a RAM. The current commands corresponding to the commanded torque and the measured speed were read from the RAM, then the armature current and the field current are regulated by the corresponding digital PI controllers. Experimental results showed that the proposed loss control was effective at low loads. The reduction of loss by 5.7% was reported. In this study the motor and converter losses were presented by a complicated expression containing many parameters. These parameters are subject to variations and their values are not readily available even for the rated operating condition. S. Chen et al. considered the optimal efficiency operation of induction motors taking into account core saturation, source harmonics, and skin effect
in an equivalent circuit analysis [79]. The harmonic losses were considered by using Fourier series analysis and superposition theorem. Skin effect was accounted for by introducing modification coefficients for the rotor resistance and the rotor reactance. The core saturation was considered by coefficient of $R_m$ and $X_m$ found from no load test. The optimal voltage and rotor frequency corresponding to the minimum loss were calculated for a range of torque and speed by computer simulation. The results showed the effect of non ideal factors on the optimal efficiency and on the corresponding values of the input voltage and the rotor frequency. The measurement results on an open loop system showed that the efficiency obtained by this method increased by up to 15% with respect to constant V/f control. However the closed loop speed control of the system was not presented. A. K. Adnanes et al. considered the efficiency improvement of EVs with both DC drives and IM drives [80]. They found the optimum control strategy, using a similar method reported in [1], for efficiency over a typical driving cycle. They reported 5-6% improvement in efficiency. The results obtained for stationary speed-torque points only and the experiment was not carried out. A method of loss minimization control for both DC and induction motors is reported in [81]-[82]. It found the optimal flux corresponding to the minimum loss based on a highly approximated expression. The expression depended on several parameters. These parameters were to be determined by measurement at different values of speed and torque. A low cost control strategy was presented in [83] which adjusts the displacement power factor based on a reference displacement angle. The reference angle must be known in advance.

b) Testing controller- This controller, also referred to as on-line or adaptive loss minimization controller calculates the power input to the drive from measured values of
input voltage and current. The controller then varies one or more variables of the motor and monitors the changes in the input power while the drive regulator keeps the output power constant. The controller finally adjusts the variable(s) for the minimum input power. A major difference with the feedforward method is the on line search on the actual system for the optimum control variable(s) instead of calculating the control variable(s) or using precalculated values. A. Kusko et al. explained the method and suggested its block diagram for a DC drive [1]. D. S. Kirschen et al. designed and implemented a testing controller for induction motors [84]-[85]. An adaptive controller adjusted the direct axis stator current for the minimum input power condition. It resulted in a reduction in the rotor flux in most cases. No special means were provided for maintaining the output power. This task is done by the speed control loop which adjusts the quadrature-axis current to compensate the effect of changes in the rotor flux. It was shown that about 50% reduction in the loss was possible at loads below 0.1 pu for the motor used in the experiment. When the motor operated at the rated flux condition, the controller took about 30 seconds to reduce the flux and brought the motor to the minimal loss operating condition. Therefore the controller suffered from a long search time. The long search time was needed since the flux was reduced in several small steps in order not to cause undesirable speed disturbances produced by large changes in the rotor flux. However the time interval between the steps must be long enough to suppress the transient period caused by each step due to the sluggish response of the speed controller. A rather similar approach was proposed in [86] where the third harmonic component of air gap flux was sensed and used to find out the motor speed. It was also used to decouple the torque and flux producing components of the stator current. The control
system is not applicable to high performance drives. P. Famouri et al. presented the implementation of a testing controller for an induction motor similar to the one reported in [85]. However they did not use field oriented control. Instead a modified constant V/f inverter was employed [87]. The rotor frequency was changed in steps in search for minimum input power. Results of simulation showed efficiency improvement over constant V/f control at low speed for square-law torque applications like fans and pumps. An 8-bit microprocessor was used for the implementation. The controller improves the efficiency of a 10 hp motor by 7.76% at 50% of the rated speed and 25% of the rated torque. The search time in this case was also long. Another modified V/f controller was recently proposed for efficiency optimization of induction motors by Celand et al. [88]. It was based on an open-loop approach. The stator voltage was perturbed stepwise in the direction that minimizes the input power. The motor speed was adjusted after each step by using a highly simplified expression as a function of voltage programmed as a part of software controller. The motor was capable of saving energy in motors working at essentially steady state with rarely intermittently altered speed command. This is because of the open-loop scalar-control used. The search time is also long. G.S. Kim et al. developed a controller aiming at both high efficiency and high dynamic performance [89]. They employed a nonlinear feedback controller which decoupled the flux dynamics from the speed (torque) dynamics. By this method the flux could be changed in large steps without causing undesirable disturbances on the speed or torque. The nonlinear controller needed the accurate information of rotor flux. A rotor flux estimator was proposed which needed the on line value of rotor resistance. This was obtained by an identification algorithm. The input power was measured at the DC link between the
rectifier and the inverter. The rotor flux was changed according to an on-line search algorithm until the minimum power consumption was found. The control was implemented on a single-chip microcomputer. The adaptive search algorithm was programmed to be executed when some change in the motor speed was detected or when a specified time period (100 sec.) elapsed. However the decoupling routine was executed much more frequently (0.5 msec.). The experimental results showed that a short search time with larger flux steps could be obtained with no undesirable disturbances in the motor speed. However the identification and decoupling processes involved a considerable amount of calculation which should have been done on-line. Moreover the method was highly parameter dependent. Recently fuzzy logic was used in the loss minimization control of induction motors [90]. It was employed to adjust the step size of flux change in search for an optimal flux condition. As the motor flux got closer to the optimum value, a Fuzzy controller reduced the step size of change in the d-axis stator current. This improved the accuracy and dynamics of the controller. A feedforward compensator was proposed to reduce the low-frequency pulsation torque due to stepwise change in the d-axis current. The compensator used a simplified torque equation not taking into account iron losses. Therefore the mechanical torque was not maintained constant. The compensator also depended on the motor parameters.

**Permanent Magnet Motors**

In contrast with DC motors and induction motors, the loss minimization control of permanent magnet motors has only been rarely reported in the literature. While many aspects of PM motors like modelling, design and performance have been the focus of intensive research in recent years, there is very little work on loss minimization control.
R. C. Colby and D. W. Novotny discussed the operation of PM motors at maximum efficiency with no provision for loss control [91]. They determined the optimum operating point, for the constant power operation, based on a simplified per phase equivalent circuit model. They derived the motor loss expression as a function of direct axis stator current, motor speed and motor flux. Assuming a constant output power and no variation in the motor parameters, the optimum d-axis current and speed were found by differentiating the loss expression with respect to the current and speed. Loss contour plots were introduced to show the nature of the loss variation with changes in the operating point and to show the existence of the optimum point. The same authors presented a testing efficiency optimizing controller for non-salient, scalar-controlled PM motors working in the synchronous mode (i.e. with no rotor position feedback) [92]. In this approach the output voltage of the inverter was adjusted to minimize the DC link current. The motor speed was maintained constant by independent open-loop control of the inverter frequency. In this work the DC link current and not the input power to the motor drive was reduced. The dynamic performance of the system was not satisfactory for high performance applications due to the open loop nature of the drive. As a result the search time was long (more than 10 seconds). The only work concentrated on the loss minimization control of PM motors in brush-less mode, as far as we know, has been reported by S. Morimoto et al. [64]. Here we review this work in some detail. The controller is a feedforward one. Morimoto et al. used a d-q equivalent circuit with synchronously rotating reference frame in order to formulate the motor electrical loss consisting of the stator copper loss and the iron loss. The latter was represented by a constant resistance in both d- and q- axis circuits in parallel with total induced voltage.
Using the voltage and current equations in connection with the torque expression, the motor loss was formulated as a function of d-axis stator current, torque and speed. The derivative of loss expression with respect to the d-axis current gave the optimal d-axis current. This expression could only be solved analytically if the PM motor had no saliency i.e. $L_q = L_d$. Morimoto et al. used a numerical solution to find the optimal value of the d-axis current, $i_d$, for a general case with rotor saliency. The optimal quadrature axis current, $i_q$, was found from the torque expression by inserting the optimal $i_d$ and the desired torque value. Having obtained the optimal values of $i_d$ and $i_q$, the motor loss was minimized by a feedforward current vector control method similar to the case of induction machines. The minimum loss was achieved by finding the smallest combination of copper loss and iron loss. The d-axis current controlled the flux level and subsequently the iron loss. The iron loss reduced in most operating conditions by applying a negative $i_d$ to produce a component of armature flux opposite to the permanent magnet flux. As a result the copper loss was increased. However the optimal values of $i_d$ and $i_q$ gave the minimum total loss. The simulation results showed about 5% or higher efficiency improvement over the conventional $i_d=0$ control for a wide range of speed at the rated torque. Also it was shown that the efficiency increased substantially at the rated speed and high loads. The relation between $i_d$ and $i_q$ was approximated by a second order function for computer implementation. A DSP based control system was implemented based on the proposed method. A TMS320C25 digital signal processor was employed to execute the whole computation. The test results were presented to show the effectiveness of the controller. It was shown also that a good transient response existed by using a digital PI controller implemented also by the same DSP. The method suffered from
parameter dependency. S. Vaez and V. I. John showed that by this method a minimum loss condition could not be achieved for many operating points [93]. They introduced variations into the motor parameters by presenting them as function of system variables. Then compared the minimum loss conditions at different operating points with constant and variable parameters. The simulation results proved that for operating points other than the rated speed and torque the optimal value of d-axis current changes substantially with the variations in the motor parameters. Recently a high efficiency PM motor control was proposed in [94]. The motor voltage was controlled to maintain unity power factor at all operating conditions. This provided the minimum input current and high efficiency. This is because the input current verses the input voltage has a V shape curve, reaching a minimum at unity power factor. However it resulted in a minimum copper loss not the total electrical loss since the iron loss was ignored.

2.3 Objectives of the Thesis

From the previous section the limitations and drawbacks of the currently available loss minimization control approaches can be summarized as follows:

The off-line methods:

i) require detailed information about the motor drive. These include an extensive mathematical model of the system and the accurate values of motor drive parameters.

ii) are sensitive to system parameter variations. These variations can be incorporated into the system model to some extent. However, the dependency of system parameters to many variables make it very difficult to come up with an accurate and realistic model.
This may lead to the operation of the motor drive under non-minimum loss conditions.

iii) need a large memory if the information regarding the minimum loss conditions are to be precalculated and stored in a memory. Otherwise they need a significant amount of computation at every sampling to find the optimal control variable from a loss model programmed into the control software.

The on-line methods:
i) take long time to transfer the system to a minimum loss condition after the steady state is detected. Therefore not much energy saving is achieved in the systems with frequent transient periods e.g. electric vehicles.

ii) suffer from stepwise changes in the control variable. As a result the system performance may not be smooth enough for sensitive applications.

iii) rely on the ability of the control system to maintain the output power constant. The available methods either fail to meet this requirement or suffer from parameter dependency which undermines the basic advantage of an on-line approach.

These limitations and drawbacks reduce the amount of energy saving in the current applications and restrict the loss minimization control to be successfully applied to many motor drives with frequent changes in the speed and/or load, and variation in the parameters. The motor drives used in EVs are examples of these kinds of applications. One objective of this work is to overcome some of these limitations and drawbacks and expand the practice of loss minimization control to new applications by proposing a novel loss minimization control strategy combining the advantages of both off-line and on-line methods.

The literature review also reveals that much effort has been directed toward the
loss minimization control of DC, induction and recently synchronous reluctance motors. However the PM motors in this regard are almost ignored. The reason for this could be twofold; firstly the traditional applications of PM motors are in the low power range, and secondly the inherently high efficiency of PM motors. However with the advent of high energy PM materials and reducing cost of power electronics PM motors are aggressively penetrating the area of drive applications with high power ratings, one example being the EV. Also by recognizing the great potential of interior permanent magnet (IPM) motors for application in variable speed drives [31], it has been realized that the potential high efficiency of these motors can only be fully realized if the motor works under a proper loss minimization control [64], [92]-[93]. Therefore it is also an objective of this work to apply the proposed loss minimization control strategy to IPM motor drives. This would be the first time that an on-line loss minimization controller is applied to a PM motor. The IPM motor is subject to considerable parameter variations mainly due to saturation. Therefore an on-line approach is the most suitable approach for this motor.

It is also an objective of this thesis to develop a new set of modelling and simulation for IPM motors to analyze the performance of these motors with and without the loss minimization control. The models should contain the system losses and give insight into the dynamics of IPM motor drives in the situation of a varying total flux as is the case in loss minimization control.

Finally the design and experimental implementation of a complete inverter-fed IPM motor drive system including the new loss minimization control strategy is an intention of the present work. The validity of the proposed control strategy should be examined by extensive tests carried out on the laboratory setup.
CHAPTER 3

Minimum Loss Operation of IPM Motor Drives

In this chapter the formulation of IPM motor losses is presented based on an equivalent circuit analysis. The condition for minimum loss operation of IPM motor drives is also presented as a basis for any loss minimization control strategy. By considering many possible variations in the equivalent circuit parameters, it is shown that the minimum loss condition is significantly affected by these parameter variations. As a practical case the minimum loss operation of an IPM motor in an electric vehicle under a standard driving cycle is considered. The potential for energy saving when the motor runs under the minimum loss condition with and without consideration of the parameter variations is discussed. The benefits of on-line loss minimization control strategies which inherently take these variations into account are highlighted.

3.1 Minimum Loss Condition

A motor can operate at a specific speed and torque with different value of stator current and magnetic flux (or supply voltage). However there is only one pair of current
and flux at which the motor electrical loss is a minimum. The electrical loss consists of copper loss and iron loss. The motor copper loss is proportional to the stator current squared, and the iron loss is roughly proportional to the magnetic flux squared. Therefore a compromise should be made between the current and the flux to achieve the minimum electrical loss. In an IPM motor flux consists of two parts; magnet flux and armature flux. D-axis armature flux produced by the negative d-axis stator current can demagnetize the magnets and reduce the total flux and subsequently the iron loss. In machines with saturation in the bridge areas between magnets on the rotor however, the iron loss may be reduced by lowering the demagnetization level [48]. In any case a current vector control strategy can control the electrical loss [64], [95]. Minimum loss condition for each speed and torque is met by finding the current vector which minimizes the electrical loss.

Assuming sinusoidal voltage, current and flux, a conventional d-q axis equivalent circuit [64], shown in Fig. 3.1, is used to find the electrical loss $W_E$ as a combination of copper loss $W_{Cu}$ and iron loss $W_{Fe}$:

$$W_E = \frac{3}{2} R_s (i_d^2 + i_q^2) + \frac{3}{2} R_c (i_{cd}^2 + i_{cq}^2)$$

(3.1)

At a constant speed and torque, the equivalent circuit and the torque equation:

$$T = \frac{3}{2} i_{q} p_n \left[ \psi_m + (L_d - L_q) i_{od} \right]$$

(3.2)

are used to find all the current components in (3.1) in terms of $i_{od}$. By substituting for these currents in (3.1), $W_E$ is reduced to a function of one variable only i.e. $i_{od}$. $W_E$, $W_{Cu}$
Fig. 3.1 IPM motor equivalent circuit at steady state, (a) d-axis, (b) q-axis.

Fig. 3.2 Motor loss (Watt) versus $i_{od}$ with constant parameters for motor #1 at 0.5 Nm and 3000 rpm; $W_E$ solid, $W_{Fe}$ dash, $W_{Cu}$ dot-dash.
and $W_{fe}$ versus $i_{od}$ are plotted in Fig. 3.2 for motor #1 (see Appendix A for the motor data). $W_E$ versus $i_{od}$ is a convex with a minimum at the optimal $i_{od}$. The optimal value of $i_{od}$ can be found by solving the derivative of $W_E$ with respect to $i_{od}$. The shape of $W_E$ versus $i_{od}$ stems from the fact that the stator current versus $i_{od}$ is a V-shaped curve. This causes the same shape for $W_{Cu}$ as in Fig. 3.2. It can be seen that the convex $W_E$ is shifted to the left of $W_{Cu}$ curve under the influence of increasing flux weakening at the higher values of negative $i_{od}$. The shifting is affected by the changes in the motor operating point (i.e. motor speed and torque) and parameters as discussed in the next section. The minimum loss operation of an IPM motor at each operating point is achieved by a current vector control in which the motor runs at the optimal $i_{od}$.

3.2 Effects of Parameter Variations on the Minimum Loss

The work reported on minimum loss operation of PM motors so far, assumed constant motor parameters. However in practice the motor parameters vary over wide ranges depending on the operating point and the ambient temperature. Morimoto et al. varied quadratic inductance, iron loss resistance and stator resistance, one at a time, by a variable coefficient [64]. They showed, by simulation, that the loss minimization control with the varying parameters resulted in a negligible difference from the one obtained with the constant parameter control. However their study was limited only to the rated operating point. It is shown here that the motor can be controlled to operate with much lower loss than the one obtained with the constant parameter control at other
operating points. This is important since the loss minimization control is essentially desired for variable speed drives working at different loads in which the minimum loss condition can not be achieved at all operating points by only motor design optimization.

Saturation

Magnetic saturation is a major source causing changes in machine parameters. It depends on the motor design. For machines with high power permanent magnets and small air gap, as modern IPM motors, the effect of saturation on the parameter variations is significant [35]-[41], [96]-[97]. Saturation in general tends to reduce machine inductances and the induced voltage in the stator winding. A machine model with extra parameters can be derived to include the saturation effects [98]-[99]. However the model presented in Fig. 3.1 can also be used with minor modifications. Saturation is caused by the flux produced by the stator current. Therefore it can be taken into the d-q axis equivalent circuit of Fig. 3.1 by substituting constant inductances, \( L_d \) and \( L_q \), and constant magnet flux, \( \psi_m \), by functions of current vector components \( i_{od} \), \( i_{oq} \) [35], [96].

For the machine #1 following equations represent saturation as obtained in[35]:

\[
\begin{align*}
L_d &= L_d^0(1 - 0.017i_{od}), \quad i_{od} \geq 0 \\
L_d &= L_d^0(1 - 0.089i_{od}), \quad i_{od} < 0 \\
L_q &= L_q^0(1 - 0.038i_{oq}) \\
\psi_m &= \psi_m^0, \quad i_{oq} \leq 0 \\
\psi_m &= \psi_m^0 - 0.0022(i_{oq} - 2), \quad i_{oq} > 2
\end{align*}
\]

Substituting \( i_{oq} \) in the above equations in terms of \( i_{od} \) and then substituting the results into the loss equation described before results in \( W_L \) with saturation as a function of \( i_{od} \).
Fig. 3.3 Electrical loss (Watt) with respect to $i_{od}$ for motor #1 at 0.5 Nm with saturation (solid) and without saturation (dash).

Fig. 3.4 Motor loss (Watt) versus $i_{od}$ with saturation for motor #1 at 0.5 Nm and 3000 rpm; $W_E$ solid, $W_F$ dash, $W_C$ dot-dash.
It is depicted in Fig. 3.3 at two values of motor speed (solid line). \( W_E \) with no saturation is also shown for the sake of comparison (dash line). \( W_E \) with and without saturation are different as expected. The important point is that the optimal values of \( i_{od} \) in these cases are quite different. The difference is more at high speed-low torque operation where the iron loss dominates the copper loss. Fig. 3.4, in comparison with Fig. 3.2, explains the effect of saturation for optimal \( i_{od} \). Saturation in Fig. 3.4 assists the reduction of \( W_{Fe} \) at the flux weakening region (high values of negative \( i_{od} \)). As a result the optimal \( i_{od} \) in Fig. 3.4 shifts to a more negative value than in Fig. 3.2.

The effect of saturation in the rotor bridge areas can be modelled by a variable iron loss resistance. Therefore it is represented in the following subsection.

**Variations in \( R_e \) and \( R_s \)**

Iron loss consists of eddy current loss and hysteresis loss. A constant iron loss resistance, \( R_e \) in Fig. 1, can represent the eddy current loss rather accurately but it is not a good representative of the hysteresis loss [77], [93]. The true value of \( R_e \) is difficult to obtain. However an approximation of \( R_e \) as a function of motor frequency can be found which models both components of iron loss as:

\[
R_e = R_{en} \frac{K + 1/f_n}{K + 1/f} \tag{3.4}
\]

where \( R_{en} \) is the value of \( R_e \) at rated speed, \( f \) is the motor electrical frequency, and \( K \) is a constant. See Appendix B for the deriving of (3.4). \( W_E \) versus \( i_{od} \) for machine #2 (see Appendix B for the motor data) is depicted in Fig. 3.5 with \( R_e \) as in (3.4) by solid line.
Fig. 3.5 Electrical loss (Watt) versus $i_{od}$ for machine #2; variable $R_c$ solid, constant parameters dash, $R_a = 2R_{an}$ dot, variable parameters dot-dash.

Fig. 3.6 One pole cross section of an IPM motor with saturation in the rotor bridges [48].
and with constant $R_c = R_a$ by dash line. It can be seen again that the values of optimal
$i_{od}$ are quite different. $W_E$ is also depicted in Fig. 3.5 for the same machine with an
elevated $R_s = 2R_a$ by a dotted line. The minimum $W_E$ shifts towards a less negative value
of $i_{od}$ when $R_s$ increases due to an increase in copper loss. The increase in $R_s$ is caused
by different factors including a rise in the ambient temperature.

Now the simultaneous variation of a combination of parameters is discussed with
reference to machine #2. This machine is less affected by saturation since saturation
mainly occurs on the q-axis [64]. Therefore only $L_q$ is modelled as a function of $i_{aq}$ by:

$$L_q = L_q^0 (1 - Ji_{aq})$$

(3.5)

where $L_q^0$ is the value of $L_q$ at $i_{aq} = 0$ and $J$ is a constant. Having $L_q$ at the rated operating
point, $L_{qr}$, from the motor data in Appendix A, and assuming that $L_q$ varies between $2L_{qr}$
and $1/2L_q$ as suggested in [64], $L_q^0$ and $J$ are found as 45.56 (mH) and 0.077
respectively. It is assumed that the minimum and maximum values of $L_q$ occur at $i_{aq} = 0$
A and $i_{aq} = 13$ A respectively.

$W_E$ versus $i_{od}$ for machine #2 including the simultaneous variation in $L_q$, $R_c$ and
$R_s$ is shown in Fig. 3.5 by dot-dash line. It is evident that the combination of changes
in different motor parameter adds to the difference between the values of optimal $i_{od}$
with and without parameter variations.

Another factor affecting the iron loss in some IPM motors is the saturation in the
iron bridges between rotor magnets [48]. A rotor construction of this type is shown in
Fig. 3.6. A negative value of d-axis current produces a flux component which opposes
the magnet flux in the stator path. However this component assists the leakage flux in
the rotor bridges. As a result the bridge saturates and the flux finds a lower reluctance path through the air gap and back to the rotor again. This causes the distortion of total air gap flux density distribution and extra iron loss associated with the flux density harmonics. The quantization of this extra loss is quite complex. However a simplified approach based on the equivalent circuit loss analysis may approximate this loss in a limited range. In this approach the iron loss resistance is modelled as function of $i_d$ as follows:

$$R_c = K_1 (K_2 + K_3 i_d)^2$$  \hspace{1cm} (3.6)$$

where $K_1$, $K_2$ and $K_3$ are decided by analyzing the experimental results of motor loss at different values of $i_d$. It is evident from (3.6) that a negative $i_d$ reduces $R_c$. This in turn increases $i_{ad}$ and $i_{eq}$ in Fig. 3.1. As a result a more negative $i_d$ contributes to both higher copper loss and iron loss. Several experimental plots of the drive input power for an IPM motor (motor #3 in Appendix A) with saturation in the rotor bridges are shown in Fig. 3.7. These plots are provided by measurement of the total input power to the system at 1/2 rated speed and different load conditions. It is evident in these plots that the optimal value of $i_d$ for a minimum input power at each load shifts towards zero.

3.3 Potential Efficiency Improvement and Energy Saving

The optimal $i_{ad}$ for motor #2 with parameter variations over a traction type torque-speed characteristic is shown in Fig. 3.8. Fig. 3.9 shows the motor efficiency in this case, $\eta_1$. The motor efficiency with the optimal $i_{ad}$ at every point calculated by the motor
Fig. 3.7 Variations in the drive input power with respect to d-axis current in the case of saturation in the rotor bridge areas. Measurement results.
Fig. 3.8 Torque (pu.) solid, and $i_{od}$ (pu.) dash versus speed.

Fig. 3.9 Motor efficiency over the whole speed range with ($\eta_1$) and without ($\eta_2$) parameter variations.
constant parameter model, \( \eta_2 \) is also shown in Fig. 3.9. It is seen that efficiency improvement is achieved over most of the operating range by finding optimal values of \( i_{otd} \) with parameter variations.

A computer software is developed by using the Mathcad\textsuperscript{TM} package to simulate the motor electrical loss over a standard EV driving cycle (see Appendix C). The software models the motor electrical loss with constant and variable parameters. In each case, the derivative of the loss with respect to \( i_{otd} \) is solved analytically to find out the optimal value of \( i_{otd} \) for each operation point. The symbolic calculation feature of the package is employed for this purpose. The optimal values of \( i_{otd} \) (found from the loss model with and without parameter variations) are then substituted back into the loss equation involving variable parameters to find out the electrical loss over the driving cycle for each case. Some assumptions are made. 1) The losses other than motor electrical loss are not studied. 2) The driving cycle used is J227a-D schedule with 28 sec. acceleration period from stop to maximum speed and by 50 sec. cruising at maximum speed, followed by coasting (10 sec.), braking (9 sec.) and idle (25 sec.) periods. 3) No gear shifting is considered. The tractive force needed for EV is scaled to be compatible with the motor rating. Motor torque during acceleration is shown in Fig. 3.9 followed by constant torque equal to 0.25 rated torque at maximum motor speed (cruising). Torque during coasting is zero and during braking is negative. \( W_e \) in both cases are shown in Fig. 3.10. Percentage of saving in electrical loss over a driving cycle is shown in Fig. 3.11. It can be seen that a substantial energy saving is possible when parameter variations are taken into account in minimum loss operation. Energy saving especially is excellent at cruising.
and coasting periods.

The practical significance of parameter variations in the minimum loss operation of IPM motor drives and the difficulties associated with the accurate modelling of these variations necessitate an on-line loss minimization control approach. This is the only approach which takes into account thoroughly the effects of parameter variations in a real situation. The main challenge however is to introduce a fast, smooth and practical on-line loss minimization strategy which can be used in high performance motor drives with rapidly changing operating conditions as in an EV.

![Graph 1](image1.png)

**Fig. 3.10** Electrical loss at minimum loss operation over J227a-D driving cycle with (solid), and without (dash) parameter variations.

![Graph 2](image2.png)

**Fig. 3.11** Percentage of saving in electrical loss over the driving cycle when parameter variations considered.
CHAPTER 4

IPM Motor Drive Control System
Design and Performance

Design and analysis of a complete IPM motor drive control system, are presented in this chapter. The overall drive system is described first. The mathematical model of IPM machines in a synchronously rotating reference frame is derived next as a basis for the controller design. The model is decoupled by nonlinear feed-forward approach and current controllers are designed. A linearized model of the mechanical subsystem is introduced which includes both components of electrical torque in IPM motors i.e. magnet torque and reluctance torque. By the help of this model the speed controller is designed and the effect of changing motor flux on the speed controller performance is evaluated. Extensive computer simulation is carried out to validate the designed control system and analyze its performance under different test conditions including changes in speed, load and flux commands. A new machine model is introduced as a basis for the simulation. This model extends the model used in the design by taking into account the motor iron loss. Using this model both the motor dynamics and the motor electrical loss can be simulated. The motor loss analysis will be presented in the next chapter.
4.1 Drive System

The simplified block diagram of adaptive loss minimization control of inverter-fed IPM motor drives is presented in Fig. 4.1. The error between the commanded and actual speed values are applied to the speed controller. The output of speed controller serves as q-axis current command, $i_q^*$. This current in conjunction with d-axis current command and their actual values, $i_d$ and $i_q$, are applied to the decoupling current controllers to produce the voltage commands $v_d^*$ and $v_q^*$. Current and voltage limiters are also included to prevent excessive current and voltage commands. The voltage commands $v_d^*$ and $v_q^*$ are transferred to the three phase stator voltage commands, $v_s^*$, $v_b^*$, and $v_c^*$ in the stationary reference frame. These voltages are applied to the PWM subsystem to provide the inverter with six gating signals. The inverter is a three phase voltage source type driven by a DC power supply. The power supply consists of a utility power supply plus a rectifier or a battery in the case of an EV. The inverter supplies the IPM motor with three phase variable voltage, at variable frequency. The motor is a three-phase, wye connected, four-pole IPM motor. Two motor phase currents are detected by current transducers and sampled by A/D converters. They are then transferred to d-q current components by Park's transformation. These current components are fed back to the current controller. An optical encoder is attached to the motor. The output signals from the encoder are used to find out accurate rotor positions which are needed for vector transformation and speed detection. The speed is detected based on the rotor position and fed back to speed controller after being filtered. The drive input power is calculated by
means of sampled DC voltage and current to the inverter. The DC current is filtered before it is used in the input power computation. The power computation block is also responsible for smoothing the DC power signal. This signal is used as an input to ALMC together with the speed error. The ALMC adjusts d-axis current command by providing it with $\delta i_d$ signal to minimize DC input power on line. The ALMC output is also applied to the non-linear speed controller to compensate the reluctance torque it produces and to keep the speed constant.

The ALMC and non-linear speed controller are the subject of next chapter. The design, analysis and simulation of other parts in Fig 4.1 which establish a complete IPM
motor drive control system, are presented in this chapter.

4.2 Machine Model

The mathematical model of IPM synchronous machines can be derived from conventional wound rotor synchronous machines with some modifications. In IPM machines, the excitation is provided by permanent magnets instead of a field winding. Therefore the field dynamics is removed. There are also no damper windings in inverter-fed IPM machines. By these modifications the voltage equations of a 3-phase IPM motor in machine variables (a-b-c stationary reference frame) is restricted to the stator equations only and represented by [101]-[102]:

$$
\begin{bmatrix}
 v_a \\
v_b \\
v_c
\end{bmatrix} =
\begin{bmatrix}
 R_x & 0 & 0 \\
 0 & R_x & 0 \\
 0 & 0 & R_x
\end{bmatrix}
\begin{bmatrix}
 i_a \\
i_b \\
i_c
\end{bmatrix} +
\begin{bmatrix}
 \psi_a \\
\psi_b \\
\psi_c
\end{bmatrix}
$$

(4.1)

where

$$
\begin{bmatrix}
 \psi_a \\
\psi_b \\
\psi_c
\end{bmatrix} =
\begin{bmatrix}
 L_{aa} & M_{ab} & M_{ac} \\
 M_{ba} & L_{bb} & M_{bc} \\
 M_{ca} & M_{cb} & L_{cc}
\end{bmatrix}
\begin{bmatrix}
 i_a \\
i_b \\
i_c
\end{bmatrix} +
\begin{bmatrix}
 \sin(\theta_e) \\
\sin(\theta_e - \frac{2\pi}{3}) \\
\sin(\theta_e + \frac{2\pi}{3})
\end{bmatrix}
$$

(4.2)

The self inductances $L_{aa}$, $L_{bb}$, $L_{cc}$; and the mutual inductances $M_{ab}$, $M_{bc}$ and $M_{ca}$ depend on the rotor position, $\theta_e$. Thus the voltage equations in machine variables are time varying when the rotor is not stalled. The well known Park’s Transformation is used to
transfer these equations to the synchronously rotating reference frame. In this especial reference frame, all the above inductances are transferred to constants. Also all the sinusoidally varying signals are referred to DC quantities along d-axis and q-axis in steady state. This simplifies the machine model enormously. However a number of assumptions are considered in the derivation. Firstly, the induced emf is assumed to be sinusoidal. Secondly, the eddy current and hysteresis losses are considered negligible. Finally magnetic saturation is ignored. The Park's Transformation [101] is carried on as follows:

\[
\begin{bmatrix}
    f_q \\
    f_d \\
    f_0
\end{bmatrix} = \frac{2}{3}
\begin{bmatrix}
    \cos(\theta_e) & \cos(\theta_e - \frac{2\pi}{3}) & \cos(\theta_e + \frac{2\pi}{3}) \\
    \sin(\theta_e) & \sin(\theta_e - \frac{2\pi}{3}) & \sin(\theta_e + \frac{2\pi}{3}) \\
    \frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\begin{bmatrix}
    f_a \\
    f_b \\
    f_c
\end{bmatrix}
\]  

(4.3)

where f can be voltage, current or flux. For a balance 3-phase condition the zero sequence component, f_0, does not exist. Fig. 4.2 shows the transformation from 3-phase a-b-c stationary reference frame to d-q reference frame.

![Fig. 4.2 Transformation from stationary reference frame to a synchronously rotating reference frame.](image)

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By using (4.3) the stator voltage equations in (4.1) is transformed to the synchronously rotating reference frame as follows [103]:

\[ v_q = R_i i_q + p \psi_q + \omega_e \psi_d \]  
\[ v_d = R_i i_d + p \psi_d - \omega_e \psi_q \]  

where

\[ \psi_q = L_q i_q \]  

and

\[ \psi_d = L_d i_d + \psi_m. \]  

Substituting (4.6) and (4.7) into (4.4) and (4.5) yields:

\[ v_q = R_i i_q + L_q p i_q + \omega_e i_d L_d + \omega_e \psi_m \]  
\[ v_d = R_i i_d + L_d p i_d - \omega_e L_q i_q. \]  

The stator frequency, \( \omega_e \), is related to the rotor position \( \theta_e \) by:

\[ \omega_e = p \theta_e. \]  

The equivalent circuit of IPM motors is obtained based on (4.8)-(4.9) as in Fig. 4.3.

---

**Fig. 4.3** Equivalent circuit of IPM motors; (a) d-axis circuit, (b) q-axis circuit.

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The torque-speed dynamics of the motor is governed by:

\[ T_e = T_L + B \omega_m + J_P \omega_m \]  \hspace{1cm} (4.11)

where \( \omega_m = \omega_c/P \) and the electrical torque is given by:

\[ T_e = \frac{3}{2} P (i_q \psi_q - i_d \psi_q). \]  \hspace{1cm} (4.12)

Substituting (4.6) and (4.7) into (4.10) yields:

\[ T_e = \frac{3}{2} P [\psi_m + (L_d - L_q) i_d] i_q. \]  \hspace{1cm} (4.13)

The input power to the motor is expressed as:

\[ P_{\text{motor}} = v_a i_a + v_b i_b + v_c i_c \]  \hspace{1cm} (4.14)

or in terms of d and q variables as:

\[ P_{\text{motor}} = \frac{3}{2} (v_d i_d + v_q i_q). \]  \hspace{1cm} (4.15)

4.3 Current Decoupling and Controllers

One advantage of d-q model of the motor is the design and operation of current controllers in a straightforward manner without worrying about the loss of bandwidth in high speed. As it is seen in Fig. 4.1, d and q axes current components are controlled by corresponding controllers to provide d and q voltage commands. However d-axis and q-axis circuits are coupled. Thus a decoupling circuit is required for the controllers design and performance. The decoupling circuit and controllers design are presented here.

It is evident from (4.8) and (4.9) that the machine model, as the machine itself,
is highly nonlinear due to the existence of terms involving products of $i_d$ and $i_q$ with $\omega_e$. This makes the design of current controllers too involved. Fortunately, the current dynamics is much faster than the speed dynamics by an order of about 100 to 1 depending on the machine ratings and applications. So it is reasonable to ignore the speed dynamics in designing the current controllers and vice versa. This means that $\omega_e$ in (4.8) and (4.9) is regarded as a constant for the purpose of current controller design. However the system is still nonlinear due to the cross coupling effects of term $\omega_e L_d i_d$ in (4.8) and term $\omega_e L_q i_q$ in (4.9). These effects are dominant, especially at high speed, since $L_d$ and $L_q$ are relatively large in IPM motors. Thus $i_d$ and $i_q$ can not be controlled independently by simple linear controllers. Linear control theory can only be applied to the machine model to design, for example, PI controllers for both $i_d$ and $i_q$ if (4.8) and (4.9) are linearized first. This means d-axis voltage depends on $i_d$ and its dynamics only, and q-axis voltage depends on $i_q$ and its dynamics only. This can be achieved by the feed-forward compensation signals as shown in Fig. 4.4 [54].

![Block diagram of current decoupling and controllers](image.png)

---

Fig. 4.4 Block diagram of current decoupling and controllers.
The current errors $\delta i_d = i_d^* - i_d$ and $\delta i_q = i_q^* - i_q$ are applied to the PI compensators $C_d(s)$ and $C_q(s)$ respectively. $C_d(s)$ and $C_q(s)$ compensate the dynamics of d-axis and q-axis voltage commands respectively, due to their own currents errors, $\delta i_d$ and $\delta i_q$. However the nonlinear terms $v_{dq}$ and $v_{q0}$ are added to the outputs of PI compensators, $v_{d1}$ and $v_{q1}$, to linearize the motor dynamics by decoupling $v_{d1}$ from $i_q$ and $v_{q1}$ from $i_d$. In an ideal situation the linearized current control systems in d and q axes are shown in Fig. 4.5 where the system transfer functions $G_d(s)$ and $G_q(s)$ are given as:

$$G_d(s) = \frac{1}{L_d s^2 + R_s}$$  \hspace{1cm} (4.16)

$$G_q(s) = \frac{1}{L_q s^2 + R_s}$$  \hspace{1cm} (4.17)

![Block diagram of linearized current control system](image)

Fig. 4.5 Block diagram of linearized current control system; (a) d-axis, (b) q-axis.

The linearized system is used to design PI compensators $C_d(s)$ and $C_q(s)$. The design is
essentially an iterative approach. The root locus method is used in combination with Bode plots as a part of the iterative approach. These classical controller design methods make it possible to come up with good approximation of controller parameters as a starting point. The design is iterated based on the results of computer simulation. Simulation is done by using both simplified system model and extensive, more complete model. The simplified model is used to determine the shape and specifications of designed controllers quickly and to modify the design. The complete model is used later to fine tune the parameters to achieve the desired system response specifications in the time domain. The final controller tuning is done at the experimental phase. A computer software is written in Matlab™ [104] to facilitate the iterative design approach (see Appendix D). The software takes the open loop transfer function of the machine in d or q axis and plots its root locus. The system closed loop poles is decided on the locus according to the desired specifications and the controller gain is obtained. Bode magnitude and phase plots are drawn next and the gain and phase margins are calculated and depicted. The closed loop transfer function response to a unit step command is then obtained. Finally the program triggers the simulation based on the simplified model which includes controller, machine and voltage saturation blocks. Several system signals including input current command and actual current are plotted upon the termination of the simulation. During the design procedure if the Bode plots, system response to unit step or simulation results are not satisfactory, new pole locations are chosen on the locus and the design procedure is repeated until acceptable results are obtained. The procedure is explained for d-axis current controller design. The design specifications are defined as follows:
a) A fast current response corresponding to a short rise time well below 10 μsec.

b) A well damped response with low overshoot.

c) Zero steady state error.

A properly designed PI controller can meet these specifications. If this PI controller is defined by the transfer function:

\[ C_d(s) = K_{pd} + \frac{K_{id}}{s}, \quad (4.18) \]

the open loop system transfer function in d-axis is obtained as:

\[ C_d(s) G_d(s) = K_{pd} \frac{s + K_{id}/K_{pd}}{s(L_d s + R_t)}. \quad (4.19) \]

The value of \( K_{id}/K_{pd} \) is chosen rather high in order to provide strong integrating action. After a number of iteration this is chosen as \( K_{id}/K_{pd} = 60.0 \). The root locus of (4.19) is plotted as in Fig. 4.6. The roots of the closed loop transfer function are decided on the locus such that a well damped response with high damping ratio is resulted. One root is shown in Fig. 4.6 close to the centre of the plot. The other root is far on the left. This corresponds to a gain of \( K_{pd} = 31 \). The Bode magnitude and phase plots are depicted in Fig. 4.7. The gain and phase margins are calculated and shown on the same plots. These margins are usually quite satisfactory over a wide range of controller parameters as is the case for the parameters chosen above. However low gain or phase margin is an indication of the system being close to instability. If this happens it is overcome in the next iteration. The step response of the transfer function is plotted in Fig. 4.8. The rise time is less than 5 milliseconds and the overshoot and the steady state error are
practically negligible. Therefore all the design specifications are achieved.

In order to evaluate the designed parameters further, the simplified system model

![Fig. 4.6 Root locus plot of the linearized system in d-axis.](image)

**Fig. 4.6 Root locus plot of the linearized system in d-axis.**

![Gain dB and Phase deg plots](image)

**Fig. 4.7 Bode plots of the open-loop d-axis current; (a) gain plot, (b) phase plot.**
in the d-axis is formed as in Fig. 4.9. The model includes $C_d(s)$, a voltage limiter and $G_d(s)$. The d-axis current is initially at some demagnetising (negative) value. A step command is applied to increase $i_d$ to zero, corresponding to rated magnet flux. The simulation results are plotted in Fig. 4.10 and Fig. 4.11. Again very good results are obtained. Further parameter tuning is done during the extensive system simulation in section 5 of this chapter.

![Fig. 4.8 Step response of d-axis current controller.](image)

![Fig. 4.9 Simplified motor model in d-axis.](image)
The q-axis current controller design is basically the same as d-axis controller design. Therefore it is not elaborated here.

4.4 Speed controller

The speed controller, like the current controller is designed based on an iterative procedure. The classical controller design theory, simulation and experience are combined to achieve a practically well performing controller. Therefore both the classical design and simulation are repeated many times during the design. It is therefore important to come up with an approach that gives good results and save time at the same time. A linearized machine model is introduced here and used at the design stage which makes the whole design approach easier and faster.
It is a common practice to drive the mechanical machine model at the rated motor flux which is, in the case of IPM motors, the magnet flux. At this special condition d-axis current component is assumed zero. This assumption simplifies the design approach considerably since the electrical torque in (4.13) is reduced to a linear equation depending on one variable, q-axis current, only as [105]:

\[ T_e = \frac{3}{2} P \psi_m i_q \]  \hspace{1cm} (4.20)

This approach eliminates the reluctance torque component from the model and does not allow to evaluate performance of the designed speed controller at other flux levels. However in IPM motors the reluctance torque plays an important role due to the rotor saliency i.e. \( L_q > L_d \). Also in some applications, e.g. EV, the speed command may often change at the flux weakening condition. Therefore it is important to consider the speed transient at different flux levels. This may be done by a non-linear simulation in which the electrical dynamics is considered together with the mechanical dynamics. However due to the time consuming nature of such a simulation it is difficult to incorporate it as a speed controller design stage. A faster design approach is obtained if a linearized model is developed which includes the non-zero \( i_q \) and therefore the reluctance torque.

Simulation of this model can be included as a speed controller design stage. It is used to evaluate the speed controller at different flux levels quickly and make wise decisions on how to modify the design in the next iteration, without running the extensive simulation. It is still necessary to run the extensive simulation at the end in order to fine tune all system controllers. However since the speed controller is designed with the help of the
proposed model, less effort and time are required for fine tuning.

Linearized Model

Equation (4.11) governs the mechanical dynamics of the drive. By assuming a constant load, the small signal version of this equation is given as:

\[ \delta T_e = J_p (\delta \omega_m) + B \delta \omega_m. \]  

(4.20)

By applying Euler formula to (4.13) \( \delta T_e \) is linearized as:

\[ \delta T_e = K_{r_q} \delta i_d + K_{r_d} \delta i_d \]  

(4.21)

in which the first term of right hand side represents the magnet torque component and the second term represents the reluctance torque component. The q-axis torque constant, \( K_{r_q} \), and the d-axis torque constants, \( K_{r_d} \), are given as:

\[ K_{r_q} = \left( \frac{\partial T_e}{\partial i_d} \right)_{i_q - i_{q0}} = \frac{3}{2} P \left[ \psi_m + (L_d - L_q) i_{q0} \right] \]  

(4.22)

\[ K_{r_d} = \left( \frac{\partial T_e}{\partial i_d} \right)_{i_q - i_{q0}} = \frac{3}{2} P (L_d - L_q) i_{q0}, \]  

(4.23)

and \( i_{q0} \) and \( i_{d0} \) are the operating point around which the linearization takes place. By equating the right hand sides of (4.21) and (4.20) the small signal model of the system in time-domain is obtained as:

\[ K_{r_q} \delta i_d + K_{r_d} \delta i_d = J_p \delta \omega_m + B \delta \omega_m. \]  

(4.24)

By taking Laplace transform of (4.24), the linearized system dynamic equation in s-
domain is given as:

\[ K_{Tq} \Delta i_q(s) + K_{Td} \Delta i_d(s) = (Js + B) \Delta \omega_m(s). \]  \hspace{1cm} (4.25)

Based on (4.25) the closed loop mechanical system, including both the magnet and the reluctance torque components, can be shown by the block diagram in Fig. 4.12 where \( C(s) \) represents the speed controller transfer function. The signals \( \omega_m, i_d \) and \( i_q \) replace their corresponding small signals for the sake of simplicity. This model is used for the speed controller design and its evaluation at different flux levels. The model takes into account the flux weakening and the contribution of the reluctance torque like an extensive nonlinear model. However it ignores the current dynamics. Therefore it can be simulated very quickly and incorporated as a speed controller design stage.

![Block diagram of the linearized system.](image)

The transfer function relating speed to \( i_q^* \) is obtained from Fig. 4.12 as follows:

\[ \frac{\omega_m(s)}{i_q^*(s)} = \frac{K_{Tq}}{Js + B} \]  \hspace{1cm} (4.26)

**Controller Design**

The speed controller is designed for the system described by (4.26) at \( i_d = 0 \).
However its performance is evaluated at zero as well as non-zero $i_d$ values with the help of the block diagram in Fig. 4.12. At $i_d=0$, the $q$-axis torque constant is reduced to:

$$K_{Tq} = \frac{3}{2} P \psi_m. \quad (4.27)$$

The design specifications are defined as follows:

a) A well damped response with limited overshoot.

b) A rise time less than 1 sec.

c) Zero steady state error.

A properly designed PI controller can meet the specifications. As is the case in the current controller design, the root locus method in conjunction with Bode plots is used for the speed controller design. A program is written to accomplish the controller design (see Appendix D). The PI controller is defined by the transfer function:

$$C(s) = K_p + \frac{K_i}{s}. \quad (4.28)$$

The open loop system transfer function is obtained as:

$$K_pK_{Tq} \frac{s + K_i/K_p}{s(Js + B)}. \quad (4.29)$$

The value of $K_i/K_p$ affects the overshoot and the response swiftness. After a number of iterations this is chosen as $K_i/P = 3$. The root locus of (4.29) is plotted in Fig. 4.13. The roots of closed loop transfer function are decided on the locus such that the design specifications are met. One root is shown in Fig. 4.13 by a plus sign. The other root is
Fig. 4.13 Root locus plot of the linearized mechanical system.

Fig. 4.14 Bode plots of the open-loop mechanical system; (a) gain plot, (b) phase plot.
far on the left. The root locus gives the value of $K_p K_{Tq}$ for the selected poles. Therefore the controller parameters are calculated as $K_p = 0.1$ and $K_i = 0.3$. The Bode magnitude and phase plots are plotted as in Fig. 4.14. The gain and phase margins are usually quite satisfactory over a wide range of controller parameters as is the case for the parameters chosen above. However, the low gain or phase margin is an indication of the system being close to instability. If this happens it is overcome in the next iteration. In order to evaluate the designed parameters, the simplified system model in Fig. 4.13 is used for a quick simulation using Simulink™ [107]. The speed response at start is plotted in Fig. 4.15 for the rated motor flux i.e. $i_d = 0$. The rise time is less than 0.2 second. The steady state error is zero The overshoot is low. Therefore all the design specifications are achieved. A lower overshoot can be achieved if a controller is designed with a smaller value of $K_i$. However this reduces the swiftness of the speed response. This is discussed in section 4.5. The response to a change in speed command is also shown in the same figure. The response is very good. The controller output, q-axis current command, is plotted in Fig. 4.16. The high starting current (dashed line) justifies the use of a current limiter. The ability of the controller to reject a load disturbances is examined next. A load of half the rated torque is applied at $t = 1.5$ second and removed at $t = 3$ seconds. The simulation results are shown in Fig. 4.17 and 4.18. The results for a rated load disturbance is shown in Fig. 4.19 and Fig. 4.20. In both cases the original speed is restored quickly after speed changes. However, in the rated torque disturbance case higher speed changes are obtained as expected. During the iterations it is found that a lower speed change can be achieved at the price of a longer speed recovery time.
Fig. 4.15 Speed response at rated flux; dash speed command, solid speed.

Fig. 4.16 Quadrature current command at rated flux; dash before limiter, solid after limiter.

Fig. 4.17 Speed response to 1/2 rated torque disturbance at rated flux; dash speed command, solid speed.

Fig. 4.18 Quadrature current command at 1/2 rated torque disturbance; dash before limiter, solid after limiter.
Fig. 4.19 Speed response to rated torque disturbance at rated flux; dash speed command, solid speed.

Fig. 4.20 Quadrature current command at 1/2 rated torque disturbance and rated flux; dash before limiter, solid after limiter.

Fig. 4.21 Speed response at $i_d=-3$ A; dash speed command, solid speed.

Fig. 4.22 Quadrature current command at $i_d=-3$ A; dash before limiter, solid after limiter.
Fig. 4.23 Speed response to 1/2 rated torque disturbance at $i_d=-3$ A; dash (speed command), solid (speed).

Fig. 4.24 Quadrature current command at 1/2 rated torque disturbance and $i_d=-3$ A; dash before limiter, solid after limiter.

Fig. 4.25 Speed response to rated torque disturbance at $i_d=-3$ A; dash speed command, solid speed.

Fig. 4.26 Quadrature current command at rated torque disturbance and $i_d=-3$ A; dash before limiter, solid after limiter.
The controller performance is evaluated at other flux levels, i.e. \( i_d \neq 0 \), by simulating the developed linearized model in Fig. 4.12. The results are shown in Fig. 4.21 through Fig 4.26 for a value of \( i_d = -3 \) A. These plots show that the controller works quite well at this condition. In fact the speed overshoot is even less in this case in comparison with \( i_d = 0 \) case. This is because a lower magnet torque is needed due to existence of the reluctance torque. Further parameter tuning is done during the complete system simulation in section 4.5 of this chapter.

4.5 System Performance and Analysis

In this section the performance of the whole drive system, including current and speed controllers, is analyzed by extensive simulation. The main purpose here is to evaluate the system behaviour under both electrical dynamics and mechanical dynamics. A new machine model is introduced for the simulation. This is achieved based on a considerable extension in the previous model to accomplish motor iron loss. The importance of this extension, apart from its effects on system dynamics, is in the loss minimization algorithm which is discussed in the next chapter.

Motor Model with Iron Loss

The iron loss in the motor is considered in this model by the iron loss or core loss resistance, \( R_e \), in parallel with total induced voltages in the d-q axis equivalent circuit of Fig. 4.3. The new system is depicted in Fig. 4.27.
Fig. 4.27 IPM motor equivalent circuit with iron loss; (a) d-axis, (b) q-axis.

The voltage equations are given in terms of $i_q$ and $i_d$, the torque producing components of $i_q$ and $i_d$ respectively as follows:

\[
\begin{align*}
v_d &= R_s i_{od} + \left( \frac{R_s}{R_c} + 1 \right) L_d \frac{d i_{od}}{dt} + \left( \frac{R_s}{R_c} - 1 \right) \omega_e L_q i_{oq}, \\
v_q &= R_s i_{oq} + \left( \frac{R_s}{R_c} + 1 \right) L_q \frac{d i_{od}}{dt} + \left( \frac{R_s}{R_c} + 1 \right) \omega_e L_d i_{od} + \left( \frac{R_s}{R_c} + 1 \right) \omega_e \psi_m
\end{align*}
\tag{4.30} \tag{4.31}
\]

In (4.30) and (4.31) the dynamic effects of iron loss current components, $i_{oq}$ and $i_{od}$, on the $L_q$ and $L_d$ are ignored since these currents are usually much less than $i_{oq}$ and $i_{od}$. The electrical torque is given by:

\[
T_e = \frac{3}{2} P \left[ \psi_m + (L_d - L_q) i_{od} \right] i_{oq},
\tag{4.32}
\]

Substituting (4.32) into the machine mechanical dynamics, (4.11), yields:

\[
\frac{J}{P^2} \omega_e = \frac{3}{2} P \left[ \psi_m + (L_d - L_q) i_{od} \right] i_{oq} - B \frac{\omega_e}{P} - T_L,
\tag{4.33}
\]
where \( \omega_e \) is given by:

\[
\omega_e = p \theta_e. \tag{4.34}
\]

Equations (4.30)-(4.31) and (4.33)-(4.34) describe the system dynamics. The state representation of these equations is used for simulation. These are obtained as:

\[
p i_{oq} = \frac{R_c}{(R_s + R_c)L_q} \left[ -R_s i_{oq} + \frac{(R_s + R_c)L_d}{R_c} \omega_e i_{od} + v_q - \frac{(R_s + R_c)\psi_m}{R_c} \omega_e \right] \tag{4.35}
\]

\[
p i_{od} = \frac{R_s}{(R_s + R_c) L_d} \left[ \frac{(R_s + R_c) L_q}{R_c} \omega_e i_{oq} - R_c i_{od} + v_d \right] \tag{4.36}
\]

\[
p \omega_e = \frac{P}{2} \left( 3 \psi_m + (L_d - L_q) i_{od} \right) - \frac{B \omega_e}{P} - T_L \tag{4.37}
\]

\[
p \theta_e = \omega_e. \tag{4.38}
\]

The inputs to the state space model of the motor are stator voltage components \( v_q \) and \( v_d \), which in turn are outputs of decoupling current controller. Then \( i_d \) and \( i_q \) are calculated from the equivalent circuit of Fig. 4.27 as follows:

\[
i_d = \frac{R_c}{(R_s + R_c)} i_{od} + \frac{1}{(R_s + R_c)} v_d \tag{4.39}
\]

\[
i_q = \frac{R_c}{(R_s + R_c)} i_{oq} + \frac{1}{(R_s + R_c)} v_q \tag{4.40}
\]

The inverse Park's transformation is applied to stator current components in d-q axis reference frame to obtain phase currents in the stationary a-b-c reference frame as.
follows [101]:

\[
\begin{bmatrix}
  i_a \\
  i_b \\
  i_c
\end{bmatrix}
= \begin{bmatrix}
  \cos(\theta_e) & \sin(\theta_e) & 1 \\
  \cos(\theta_e - \frac{2\pi}{3}) & \sin(\theta_e - \frac{2\pi}{3}) & 1 \\
  \cos(\theta_e + \frac{2\pi}{3}) & \sin(\theta_e + \frac{2\pi}{3}) & 1
\end{bmatrix}
\begin{bmatrix}
  i_q \\
  i_d \\
  i_0
\end{bmatrix}
\]

(4.41)

where \(i_0\) is zero since the system is assumed to be a balanced three phase system.

Equations (4.35)-(4.41) together with (4.32) form the complete machine model.

**System Simulation**

The drive system performance is evaluated by extensive non-linear simulation. This is done in Matlab™ [104] environment by the simulation tool Simulink™[105]. The block diagram of the system is shown in Fig. 4.28. The motor model developed in the previous section forms the main subsystem. It is represented by the block IPM motor having four inputs i.e. \(v_a, v_b, v_c\) and \(\theta_e\); and ten outputs: \(i_{ad}, i_{aq}, \omega, i_d, i_q, T, i_a, i_b, i_c\) and \(\theta_e\). The inputs to this subsystem come from vector transformation block which transforms d-q voltage components to a-b-c phase voltages. The structure of subsystem IPM motor is shown in Fig. 4.29. It consists of another vector transformation block and machine dynamics block. The former contains eqn (4.41) and the latter includes eqns(4.35)-(4.40), (4.3) and (4.32). Two current components \(i_d\) and \(i_q\) are feedback to their corresponding current controllers to provide voltage commands. The current decoupling subsystems are also shown in Fig. 4.28. The speed signal \(\omega\) is feedback to the speed controller after division by the number of pole pairs. Output of speed controller passes through a limiter block and is used as the q-axis current command. The d-axis current command is a
Fig. 4.28 IPM motor drive system block diagram for simulation.
Fig. 4.29 IPM motor subsystem.
constant or it is decided by ALMC as proposed in the subsequent chapters.

The system is simulated extensively by a fast and accurate version of Runge-Kutta method on work station computers. The simulation is carried out for different IPM machines at different operating conditions. The results for the machine used in experiment are presented here. Many signals are traced and plotted for system analysis. The results for the controlled parameters designed in this chapter are shown in Fig. 4.30 through Fig. 4.41. The drive response, at no-load, to a rated speed command at \( t=0 \) is shown in Fig. 4.30. The motor speed rises quickly and settles to the commanded value in about 1 second. The steady state error is zero. A limited overshoot occurs at about \( t=0.2 \) second. To examine the speed controller robustness against changes in the speed command, a step drop reduces the speed command to 120 rad/sec. at \( t=1.5 \) second. The actual speed follows the commanded value closely with a negligible undershoot and a zero steady state error. The q-axis current command is shown in Fig. 4.31 where the maximum value is limited to twice the rated motor current. The d-axis current command remains almost zero during rapid speed transients reflecting the effectiveness of current decoupling circuit described in section 4.4. Comparison of speed and \( i_q \) with Fig. 4.15 and 4.16 respectively shows close agreement between the simulation results of simplified model and extensive model due to decoupling circuit. Somehow lower speed overshoot and \( i_q \) in Fig. 4.30 and 4.31 can be explained by the inclusion of electrical loss in the extensive model. Phase currents are shown in Fig. 4.32 and the torque in Fig. 4.33. The torque response is fast enough to reach the peak value in a few milliseconds. The voltage commands in the rotor reference frame and in the stationary reference frame are depicted
in Fig. 4.34 and 4.35 respectively. It is assumed that the DC power supply is able to provide the required inverter voltage. Therefore no voltage limiter is necessary. If it is not the case a provision should be made in order to prevent saturation of current controllers. This is discussed later.

The simulation results of load disturbance test is presented in Fig. 4.36 through

Fig. 4.30 Speed response at no load; dash speed command, solid speed.

Fig. 4.31 Current components; d-axis dash, q-axis solid.
Fig. 4.32 Motor phase currents at no load.

Fig. 4.33 Electrical torque at no load.

Fig. 4.34 Voltage command components; d-axis solid, q-axis dash.

Fig. 4.35 Phase voltage commands.
Fig. 4.41. Here a rated load torque is applied to the motor at \( t = 1.5 \) second and removed at \( t = 3 \) seconds both when the motor is running at steady state condition as can be seen in Fig. 4.36. The ability of the control system to restore the original speed, after a deep or a tip, is satisfactory. The q-axis current increases to about maximum value at the application of the load and remains high until the load is removed as seen in Fig. 4.37. This assures that the maximum motor current capacity is used to provide a quick response. The same conclusion can be achieved from the phase current response in Fig. 4.38. The motor torque is plotted in Fig. 4.39. Again it is seen that the torque increases to a high value quickly as the load is applied. However this value is less than the value at start. This is due to a higher motor loss at the loading condition with respect to the no load condition. Voltage commands in d-q axes and in the a-b-c- axes are shown in Fig. 4.40 and 4.41 respectively. It can be seen that phase voltages may reach a high value at

Fig. 4.36 Speed response to a rated load disturbance; motor speed solid, speed command dash.

Fig. 4.37 Current components at load disturbance test; d-axis solid, q-axis dash.
Fig. 4.38 Motor phase currents at disturbance load test.

Fig. 4.39 Electrical torque at disturbance load test.

Fig. 4.40 Voltage command components at disturbance load test; d-axis solid, q-axis dash.

Fig. 4.41 Phase voltage commands at disturbance load test.
rated speed rated load condition, unless a flux weakening approach is provided. This may limit the loading capability of the drive in practice if not enough output voltage is available from the DC power supply.

A very low speed overshoot may be desirable for some applications including EVs. This can be achieved by a PI speed controller if the integrating action is weakened. The speed controller is redesigned to accomplish this specification and the simulation is repeated for the new design. The results are presented in Fig. 4.42 through 4.47. It can be seen, in Fig. 4.42, that a lower speed overshoot at start is achieved at the price of longer settling time. Also the speed recovery time after the load disturbance is longer and changes in speed are higher when the load disturbance test is carried on.

Fig. 4.42 Speed response to a rated load disturbance (low overshoot); motor speed solid, speed command dash.

Fig. 4.43 Current components at load disturbance test (low overshoot); d-axis solid, q-axis dash.
Fig. 4.44 Motor phase currents at disturbance load test (low overshoot).

Fig. 4.45 Electrical torque at disturbance load test (low overshoot).

Fig. 4.46 Voltage command components at load disturbance test (low overshoot); d-axis solid, q-axis dash.

Fig. 4.47 Phase voltage commands at load disturbance test (low overshoot).
CHAPTER 5

Adaptive Loss Minimization Control

This Chapter is devoted to the introduction of new ideas in loss minimization control of motor drives in general and their applications to IPM motor drives. The basic concepts and principles of a new adaptive loss minimization controller (ALMC) are presented. The advantages of the proposed ALMC are emphasised in comparison with the existing methods. A nonlinear speed controller (NLSC) is also introduced which combines the function of a usual speed controller with a compensation process needed in many on-line loss minimization approaches. These concepts and principles are then used to analyze and design the ALMC and NLSC. A new insight into the behaviour of the motor is given by the help of the linearized model developed in the previous Chapter. Finally the whole motor drive system including the ALMC and the NLSC are evaluated by extensive simulation.
5.1 Basic Concepts and Principles

In this section new concepts and principles are presented for both the motor drive loss minimization process and compensation for changes in the operating point during the loss minimization process. These concepts and principles aim at the improvement of loss minimization control theory and practice to overcome some of the major difficulties associated with the existing approaches. The objective is to extend the loss minimization practice to a number of new applications in which the fast dynamics and high performance is as important as the efficiency gain.

Loss Minimization

Common to all adaptive loss minimization controllers is the on-line adjustment of a control variable, which affects the input power, in search for a minimum input power. A common practice in loss minimization of induction machines is to apply a step change to a control variable e.g. flux or d-axis component of stator current. Then wait for some time, long enough for the motor to pass the subsequent transient and come to a fairly steady state situation. Then compare the input power values before and after the change made in the control variable. If the power reduces, another step change is applied to the control variable until the minimum input power is achieved. A great deal of efforts have been directed towards the adjustment of the step size of the control variable [88]-[91]. However the stepwise change in the control variable is the source of disturbance in the developed torque [84]. It also causes speed fluctuations and may cause instability in the system[88]. A more important deficiency of the stepwise change in the control
variable is the long time the controller takes to reach the minimum input power [85]-[88], [90]-[91]. However, in many applications with frequent changes in the operating point (speed and torque), e.g. in EVs, the steady state period is short. Therefore the controller does not have enough time to reach the minimum input power condition. Thus the online loss minimization is impractical. Even if the steady state period permits the finding of the minimum input power, some energy is lost during a long search time. If the transient state is repeated frequently, the total energy saving may not be significant.

In the proposed ALMC the stepwise change of control variable is eliminated. Instead it is replaced by a continuous adjustment of the control variable. The stepwise change in the control variable is probably better match for the low speed of early microprocessors. However, with the increasing computational speed and falling price of fast processing units like DSPs it is possible to chose alternative patterns for the control variable based on incremental instead of stepwise change. One simple pattern of change in the control variable is a linear pattern with respect to time. In this case the control variable follows a ramp trajectory. Other more complicated patterns can also be chosen. Advantages and requirements of a continuous trajectory for the control variable are discussed here.

Avoiding the stepwise change in control variable eliminates the repeated stresses applied to the motor. As a result the system works smoothly. The possibility of increase or decrease in the motor speed caused by the stepwise changes is also eliminated. Therefore it provides a more reliable loss minimization controller. A faster loss minimization is achieved by the continuous changes in the control variable since the
relatively long transient period after each step change is avoided. The accuracy of the achievable minimum loss, in the stepwise methods, depends on the step size. The step size is reduced as the input power reduces. However, there is always a practical limit for the step size since the changes in the input power can not be determined if a small step size is chosen. In the proposed controller a new method is used for monitoring the changes in the input power. The input power is monitored continuously inside a moving window. Therefore as the control variable changes continuously, the change in the input power is determined at each instance by the difference between the two values at the beginning and at the end of the moving window. In this case there is a limit for the size of window. However since changes in both the control signal and input power occur continuously a closer value to a true minimum loss can be identified.

The proposed method demands more computation and memory. However the extra computation is in the matter of a few summations or subtractions during a sampling interval. This can quite conservatively be ignored even with a low speed processing unit. The continuous determination of change in the input power also requires more computation. This can be handled easily by many existing processors like DSPs. There is also a need for extra memory to store the values of the input power during a time interval equal to the period of the moving window.

**Speed Compensation**

A change in the control variable changes the motor flux. Therefore the developed torque and ultimately the motor speed changes unless some sort of compensation is provided to maintain the motor operating point (speed and torque). Kirschen et al. [84]-
[85] suggested fast speed dynamics for the speed controller to reduce the transient period after a change is made in the control variable. This method works if the subsequent transient change in the motor speed is permissible and a slow loss minimization controller is acceptable for the application. Kim et al. [89] presented a decoupling system for high dynamic performance motors. This is done by on-line calculation of a signal which decouples the motor speed from the flux (control variable). However the decoupling procedure is based on an approximation in the system model and depends on the motor parameters. A torque compensation method is proposed in [99]. It uses the torque equation to calculate the change in the q-axis current command required for the compensation of the change in the control variable (d-axis current command). This method also depends on the motor parameters. Also it ignores the motor iron loss in the torque equation. In some cases, e.g. IPM motors, the iron saturation plays an important role in the motor performance and efficiency as discussed in Chapter 3. Therefore using a parameter dependent compensator diminishes the important advantages of an on-line loss minimization approach.

The compensating nature of the speed controller is employed here to introduce a novel compensator which is both fast and independent of the system parameters. In a motor drive system the speed signal, rather than the torque signal, is more convenient to be considered for the compensation purpose in the loss minimization. This is because the speed signal can be measured readily or calculated from the rotor position without involvement of motor parameters. Moreover in a system with an outer speed loop, the speed signal is already available. The primary function of a speed controller is to respond
to a speed command. However in the case the speed command is unchanged, the controller resists against any change made to the motor speed caused by a change in the developed torque. This natural tendency is used here to compensate the effect of a change in the loss minimization control variable. As discussed in the previous chapters the d-axis component of motor stator current, $i_d$, is chosen as the control variable for the ALMC. As $i_d$ is changed by the ALMC to reduce the drive input power, the developed torque tends to vary. Since the load is assumed constant, the change in torque results in a change in the motor speed. However, the speed controller tries to maintain the speed at its commanded value. Therefore it modifies the q-axis current command in response to a change in $i_d$. This natural compensation process depends on the motor torque speed dynamics which is slow in nature. As a result once $i_d$ changes by the ALMC, the speed undergoes a transient period before the natural compensation process restores the commanded speed. The main idea which will be described in the proposed nonlinear speed controller is to speed up the natural compensation process when $i_d$ is changed by the ALMC. This is done by manipulating the input to the speed controller, i.e. the speed error signal, when the ALMC is active. However the speed error remains intact when the system leaves the steady state condition and the ALMC is deactivated. This provides a usual speed controller in the transient state.

5.2 Analysis and Design

Based on the concepts and principles described above the design of the ALMC
and the NLSC are presented in this section.

**Adaptive Loss Minimization Controller (ALMC)**

The ALMC functionally consists of five distinct components:

1. input power processing
2. steady state speed detection
3. direction test
4. loss minimization
5. fast dynamics.

In the actual control system implementation these functions are carried out by a number of software routines as described in detail in the next chapter. In simulation however, the simulation tool and language determine the way in which the functions are realized. Here the general description of the ALMC functions, common to both simulation and implementation, is presented.

The ALMC reduces the drive input power continuously. The power processing is carried out to find the change in the input power as explained above. The adaptive loss minimization controller becomes active when the motor speed reaches the steady state condition and remains in this state for a short period of time. For this purpose several samples of the motor speed, during a time span, are compared with the commanded value. If the error in all the comparisons is less than a prescribed tolerance value then the power reducing direction of $i_a$ is detected by a direction test. This is done by changing $i_a^*$ in a ramp manner for a specific period of time. If at the end of this period a reduction in the input power is determined the ramp continues in the same direction;
otherwise the direction of change in $i_d^*$ is reversed and the direction test is repeated. As a result the loss minimization mode continues until $i_d^*$ reaches its optimal value corresponding to a minimum input power, or a minimum loss. The minimum input power is determined where the reduction in the input power, in a time period equal to the one used for the direction test, falls below a certain limit. This is the end of the loss minimization mode. A new mode of operation, the triangular mode, then is started in which the direction test is repeated with an alternating direction. Therefore $i_d^*$ follows a triangular trajectory. This way if the optimal $i_d^*$ changes as a result of a gradual change in the operating condition or parameter variations, the ALMC follows the minimum loss condition by transferring from the triangular mode to the loss minimization mode. Such a change in the operating point can occur by a limited change in the mechanical load or commanded speed while the parameter variations may be the result of a temperature rise for instance.

If the speed error exceeds a predefined band for any reason, e.g. a load disturbance or a speed command change, the operation of ALMC is deactivated regardless of the current mode of operation. Subsequently the d-axis current command returns to its original value to ensure a fast system dynamics during the transient condition. The original $i_d^*$ can be a constant value or depends on the operating point. In any case it can be determined based on a criterion resulting in a desirable transient response. The ALMC becomes active once again when a steady state speed is detected. More details on the design of the ALMC are provided in the next chapter.
Nonlinear Speed Controller (NLSC)

The linearized model developed in Chapter 4, together with the natural compensation concept described above are employed to design a nonlinear speed controller. This controller functions both as a usual PI speed controller and a compensator. Fig. 4.12 is resketched here in Fig. 5.1 (a). This explains the drive performance under the transient as well as the steady state conditions. The system dynamics at the steady state speed with slow variations in $i_d^*$ can be extracted from this model as is shown in Fig. 5.1 (b). The slow variations in the d-axis current command is shown by $\delta i_d^*$. Since the speed command and the motor load are constant here, the system is in the steady state condition. Therefore the variations in the actual speed are the results of $\delta i_d^*$ through a change in the reluctance torque, $\delta T_d$. The PI speed controller in this condition behaves as a feedback to compensate the speed variations. The relation between $\delta i_d^*$ and $\delta \omega_m$ is obtain from Fig. 5.1 (b) as:

$$\frac{\delta \omega_m(s)}{\delta i_d^*(s)} = \frac{K_{T_d}}{s^2 + \frac{B + K_p K_{T_d}}{J}s + \frac{K_i K_{T_d}}{J}}$$

The compensation depends on the poles of (5.1). The poles in turn are functions of the proportional and the integral gains $K_p$ and $K_i$ respectively. For normal values of $K_p$ and $K_i$ which are chosen based on a desired speed response as obtained in Chapter 4, a natural compensation occurs as seen by the simulation results in Fig. 5.2. The motor speed is in the steady state, Fig. 5.2 (a), when $i_d^*$ is reduced in a ramp form similar to a loss minimisation situation as in Fig. 5.2 (b). The speed undergoes an increase as a
Fig. 5.1 System block diagram; (a) linearized system, (b) compensator.
result of an increasing $\delta T_d$. However the speed settles to the original value by the natural compensating action of the PI controller. It is seen in Fig. 5.2 (b) that the controller modifies $i_d^*$ corresponding to $i_d^*$ to restore the original speed. The compensating action can be speeded up by a forced compensation in which the PI controller transfer function is multiplied by a factor greater than unity. This moves the poles of (5.1) further to the left in the complex plane and improves the speed compensation. Fig. 5.3 shows the simulation results for this situation. The d-axis current command in this case, Fig. 5.3 (b), follows the same trajectory as in Fig. 5.2 (b). However, the change in the motor speed is much less as shown in Fig. 5.3 (a). This indicates that the forced compensation is effective during the loss minimization process. However a PI speed controller with high gains can not be used in the system since it could result in an undesirable speed transient in response to a speed command. A nonlinear speed controller (NLSC) is introduced next to provide a forced compensation action without deteriorating the speed response.

The fact that ALMC is active only in the steady state provides a excellent opportunity to design a NLSC which provides a forced compensation when the output from the ALMC is not zero. In the transient state however, the NLSC works as a simple PI controller resulting in a desirable speed response. The block diagram of the proposed NLSC is shown in Fig. 5.4. A PI controller with the original gains, designed for a desirable speed response, is used. However the speed error signal is supplemented by a new signal $|\delta i_d^*| K \delta \omega_m$. The constant $K$ is the non-linearity gain since its value determines the effect of non-linear term $|\delta i_d^*| \delta \omega_m$. The effective speed error signal is then given by:
Fig. 5.2 Simulation results of natural compensation.

Fig. 5.3 Simulation results of forced compensation.
Fig. 5.4 Block diagram of nonlinear speed controller (NLSC).
Fig. 5.5 Simulation results of NLSC K=2.

Fig. 5.6 Simulation results of NLSC K=20.
\[ \delta \omega_{\text{eff}} = \delta \omega_m (1 + |\delta i_d^*| K \delta \omega_m). \]  

(5.2)

Notice that the absolute value of the ALMC output is used in (5.2) since the sign of the supplementary signal is determined by the speed error signal. The NLSC ensures a magnified error signal corresponding to the magnitude of \( \delta i_d^* \). It works similarly to a PI controller with high controller gains when the ALMC is working. Otherwise \( \delta \omega_{\text{eff}} = \delta \omega_m \) in the transient state when the ALMC is off and \( \delta i_d^* = 0 \). The simulation of the system with the NLSC is given in Fig. 5.5 or \( K = 2 \). The speed compensation is improved with respect to the natural compensation with \( K = 0 \). By increasing the non-linearity gain a very good compensation is provided as seen in Fig. 5.6.

5.3 Simulation of the System

The simulation results of the whole motor drive system including the ALMC and the NLSC are presented in this section for different operating conditions. The saturation of rotor bridges as discussed in Chapter 3 is also taken into account at the end of the section.

The simulation is done by Simulink™ software. A system block diagram for the simulation is shown in Appendix E. The loss minimization algorithm is developed as a Matlab™ file and inserted into the simulation system as a block. Fig. 5.7 (a)-(h) shows the system performance under a low load condition at the rated speed. Different modes of operation can be seen in this figure. These include the transient state, direction test, loss minimization, triangular mode and fast dynamics. It is evident that the ALMC
Fig. 5.7 Simulation results of motor drive system including ALMC.
Fig. 5.7 Continued.
together with the NLSC provides an elegant loss minimization process. After the steady state speed is detected by the ALMC, the direction test determines that the d-axis current must be decreased for minimizing the drive input power. Then the input power is continuously reduced by a ramp form reduction of $i_d$ as long as the change in the input power is less than a small negative band as shown in the Fig. 5.7 (e) and (f). When the change in the input power exceeds the band, the loss minimization mode is terminated and the triangular mode starts. The whole process is smooth, fast and accurate. The system perfectly withstands a large change in the speed command when it is under the control of the ALMC. This is examined by applying a step increase in the commanded speed when the system works in the vicinity of minimum loss at the triangular mode as in Fig. 5.7 (a). As a result the operation of ALMC is terminated, the original d-axis command is restored (Fig. 5.7 (b)) and the new speed is achieved after a smooth and fast transient period.

In the case the rotor shows a significant saturation in the bridge areas due to the rotor structure, as discussed in Chapter 3, the motor performance differs. This is due to the fact that the iron loss increases with more negative values of $i_d$. To take this fact into account the machine model is modified by using a variable iron loss resistance as in (3.6). The simulation results for this case are shown in Fig. 5.8 (a)-(h). In this case a negative $i_d$ increases the total iron losses. Therefore an increasing $i_d$ is determined by the direction test routine to reduce the drive input power. By increasing $i_d$ the input power reduces continuously similar to the previous case. The fast and stable system response to a speed change is also shown in Fig. 5.8.
Fig. 5.8 Simulation results of motor drive system including ALMC at 1/2 rated speed with saturation at the rotor bridges.
Fig. 5.8 Continued.
CHAPTER 6

Control System Implementation

One of the objectives of this thesis was to build an experimental control system including all the new features introduced in this work like the adaptive loss minimization controller (ALMC) and nonlinear speed controller (NLSC). This was done and in this chapter the detailed description of both the hardware and the software of the system is presented. After discussing the implementation issues of each system component, the sample test results of that component are presented. The overall experimental evaluation of the whole system however remains to be addressed in the next chapter.

Starting with a general description of the experimental set up, the PWM (Pulse Width Modulated) system is presented next. PWM is an important part of the overall system and involves many theoretical and practical considerations. The focus here is more on the practical aspects. A successful system implementation depends to a great extent on the existence of system signals in desirable forms. This is specially important in the proposed on-line loss minimization control strategy where smooth signals are essential. Therefore the third section belongs to the signal measurement and manipulation. Then the implementation of the decoupling current controller is discussed, followed by the presentation of the major contributions of the work i.e. NLSC and ALMC in the last two sections of this chapter. These are based on the concepts, principles, analysis and design of NLSC and ALMC presented in the previous chapter.
6.1 Experimental Set up

The laboratory set up of the IPM control system is depicted in Fig. 6.1. All the software routines are developed in "C" code and downloaded into the DSP board installed on a PC as host computer. The DSP board receives signals corresponding to five system variables i.e. two phase currents, encoder output, DC bus voltage and DC link current. The outputs of the board consist of three modulated signals. However these signals need further processing to produce the inverter gating signals as desired by the inverter. This is done by a breadboard circuit named the hardware circuit in Fig. 6.1. This circuit also includes lock-out circuit and primary amplification of gating signals. A six-channel pulse amplifier further amplifies the gating signals as desired by the control circuits of the inverter transistors. The inverter is supplied by a DC power supply connected to a three phase main. The inverter provides three phase variable voltage variable frequency supply for a one hp, 4 pole synchronous IPM motor with Samarium Cobalt magnets (see Appendix A for the motor specifications). A dynamometer is connected to the motor through a belt and is used as a mechanical load. The current signals and the DC voltage are sensed by five transducers with good accuracy and negligible dc offset. The rotor position is detected with the help of a high resolution encoder installed on the motor shaft. Several digital multimeters monitor the voltage and current values during the experiments. Protecting fuses and circuit breakers are placed at the output and the DC input of the inverter. A picture of the set up is shown in Fig. 6.2.
Fig. 6.1 Experimental system set up.

Fig. 6.2 A view of the actual set up.
The DSP board [108] is a DS1102 designed for the development of high speed multivariable digital controllers and real-time simulations in various fields including drive control and vehicle control. The board is based on a floating point, 40 MHz TMS320C31 DSP with 50 ns single cycle instruction execution time. It performs parallel multiply and ALU operations on integers or floating-point numbers in a single cycle. The TMS320C31 supports a large address space with various addressing modes allowing the use of high-level languages for application development. Among the many features the DSP enjoys 32-bit instruction and data words, 24-bit addresses, eight 40-bit accumulators, 2- and 3- operand instructions and two 32-bit timers. The DSP is supplemented by a number of on-board peripherals to form the DS1102 controller board. A block diagram of the board is shown in Fig. 6.3. It contains two 16-bit and two 12-bit analog to digital converters and four digital to analog converters. The digital I/O subsystem is based on a second DSP, TMS320P14. This DSP can be programmed as a slave DSP. In this work it is used in the PWM generation. Two incremental sensor interfaces are able to accept the motor encoder signals. There is also available on the board 128K words memory fast enough to allow zero wait state. The board data sheet [108] and the block diagram of the main DSP [109] are included in Appendix F.

The main DSP communicates with a host computer through a host interface. All off-chip memory and I/O can be accessed by the host while the DSP is running. This allows easy system setup and monitoring. A special software [110] is used for getting on-line information about the control system variables under process in the DSP. This information can be plotted on the host monitor or stored on a disk for later analysis.
Fig. 6.3 Block diagram of the DSP board.
6.2 PWM System

A three phase voltage source inverter is used to drive the IPM motor. It converts a DC input voltage, $V_{DC}$, to a three-phase AC voltage, with variable magnitude and variable frequency. A schematic diagram of the inverter is shown in Fig. 6.4. It consists of six Darlington transistor modules, $T_1$-$T_6$, which are arranged in three pairs corresponding to three phases. The inverter control circuit is omitted in Fig. 6.4 for the sake of simplicity. However this controls the switching of transistor modules. Different switching strategies are applied to voltage source inverters. A popular technique in industrial applications is sinusoidal pulse width modulation (SPWM). This technique is used throughout the experiments here and its principles are reviewed here briefly.

SPWM

Fig. 6.5 shows the basic idea of a SPWM [111]. A triangular carrier wave of frequency $f_c$ is compared with three sinusoidal voltage command signals (or modulating signals). The points of intersection between the triangle wave and each modulating signal determine the switching instances of transistor pairs for the corresponding phase. In Fig. 6.5 only a small fraction of a cycle for modulating waves are shown. Since the frequency of modulating signals is much more than $f_c$, these sinusoidal signals look like constant signals in a short interval. In this figure the lower modulating signal, for instance, is the commanded voltage of phase c. Intersections of this signal with carrier wave determines the switching points of transistors $T_2$ and $T_5$ as in the last output voltage in Fig. 6.5. The pulse and notch widths of this voltage correspond to the on periods of $T_2$ and $T_5$.
Fig. 6.4 Schematic diagram of a 3-phase voltage source inverter

Fig. 6.5 Principle of sinusoidal pulse wave modulation (only a small fraction of a cycle for sine modulating signals are shown).
respectively.

The slave DSP on the DS1102 board in connection with a simple hardware circuit is used to provide switching signals for transistor modules. This DSP is programmed in PWM mode by specifying the number of channels (modulating signals) and choosing a carrier wave frequency. A maximum number of 6 modulating signals can be specified. By applying a normalized sinusoidal modulating wave, with the magnitude between -1 and 1, a modulated signal is produced by the DSP. Such a signal is shown by PWM₀ in Fig. 6.6. This is produced by first normalizing the modulating wave of phase a with respect to its maximum possible magnitude. Then the normalized signal is applied to the DSP. The pulse width of PWM₀ signal corresponds to the normalized value of modulating signal. If the carrier wave period is shown by T_c, a normalized modulating wave of magnitude zero provides a pulse width of T_c/2. While normalized values of -1 and 1 provide the switching signals with the pulse widths of zero and T_c respectively. Therefore PWM₀ in Fig. 6.6 is produced by a modulating wave of normalized magnitude less than zero.

The outputs from the slave DSP in PWM mode need some modifications before they are used as transistor switching signals [112]. If PWM₀ in Fig. 6.6 is compared with the output voltages in Fig. 6.5, it is evident that the pulse periods in Fig. 6.5 are symmetric with respect to the centre of a carrier wave period. However pulses produced by the slave DSP have the same starting point but they are not symmetric with respect to the centre of a T_c period. Fortunately this discrepancy can be removed if two DSP channels are used to produce one switching signal. Fig. 6.6 shows the
Fig. 6.6 Generating transistor gating signals by modifying outputs of built in PWM subsystem.
procedure for producing the switching signal of phase a. $PWM_0$ and $PWM_1$ signals are produced by two of the DSP Channels. The modulating signals for these channels are the normalized values of the phase voltage command and its negative respectively. By applying $PWM_0$ and $PWM_1$ to a X-OR logic the switching signal for phase "a" i.e. $PWM_a$ is produced. This signal is in the desired form as compared with the output voltages in Fig. 6.5. By this way all six channels of the slave DSP are used to provide three switching signals.

**Lock Out Circuit**

Each of the three switching signals controls a pair of transistor modules on an inverter leg. In the case of phase "a" for instance, the signal $PWM_a$ turns on $T_1$ during its pulse and turns on $T_2$ during its notch. It is important that at each moment only one transistor on a leg is on and the other transistor on the same leg is off. Otherwise a shoot through will damage the transistor pair. Due to the transistor turn-on and turn-off delays, a lock out circuit is provided in order to prevent any shoot through in the case a transistor on a leg is turning off while the other transistor on the same leg is turning on [111]. This is done by the provision for a lock-out time at the switching instances over which both transistor gating signals are off. In practice a mono-stable multivibrator is used to provide two lock-out periods at the leading and trailing edges of signal $PWM_a$. These are shown in Fig. 6.6 by signals $Q^-_1$ and $Q_2$. The lock-out time, $t_l$, must be at least equal to the total switching time of a transistor. This period is usually designed conservatively to provide enough protection against an inverter failure. Applying $Q^-_1$ and $PWM_a$ signals to an AND logic gives the switching signal for $T_1$ i.e. the signal $g_1$. While
Fig. 6.7 Diagram of PWM hardware subsystem.

Fig. 6.8 Picture of PWM hardware subsystem built on breadbord.
applying $Q_2$ and $PWM_4$ to a NOR logic gives the switching signal for $T_4$ i.e. the signal $g_4$. Both $g_1$ and $g_4$ are shown in Fig. 6.6. The circuit diagram of the PWM hardware subsystem is shown in Fig. 6.7. Two Op-Ams or drivers in this circuit are used to improve the gating signals before they are applied to a 6 channel pulse amplifier. A picture of the PWM hardware subsystem built on a breadboard is shown in Fig. 6.8.

**Test Results of PWM Inverter**

The PWM system is examined extensively before it is used in the motor drive control system. Several static tests are carried out to ensure a proper operation of software and hardware subsystems of the SPWM. Three-phase sinusoidal modulating signals are produced by the main DSP. A DSP routine is developed for this task. The modulating signals are normalised to provide inputs to the slave DSP. Finally the circuit in Fig. 6.8 is used in connection with a 6-channel pulse amplifier to control the inverter. Three-phase resistive-inductive (RL) loads are used in the static test. The test is repeated for different values of DC supply voltage, maximum modulating wave magnitude and load. The experimental results proved the ability of SPWM inverter to provide a three phase supply for static loads. Fig. 6.9 shows the experimental results for a load with $R=5 \ \Omega$ and $L=8 \ \text{mH}$. The DC input voltage to the inverter in this case is $V_{dc}=20.5 \ \text{V}$. Fig. 6.9 (a) and (b) show the actual and normalized waveforms of the modulating (commanded) voltages. Fig. 6.9 (c) shows the modulated output voltage at the load terminals (line voltage). The phase current is shown in Fig. 6.9 (d). The high frequency noise on the phase current is caused by transistor switching. The relative height of spikes are reduced as the current increases. This is shown in Fig. 9.10 where the applied $V_{dc}$
Fig. 6.9 Experimental results of SPWM applied to a three phase RL load.
Fig. 6.10 Experimental results of SPWM applied to a three phase RL load with increased inverter input voltage.
is increased to 80V as shown in Fig. 9.10 (c). A rather smooth sinusoidal current is produced as in Fig. 6.10. (d).

6.3 Signal Measurement and Manipulation

The motor drive control system manipulates the signals in order to provide a desirable drive performance. Therefore a major step in the implementation of the control system is to access all the required signals. The primary signals like currents and voltages are measured directly through the transducers and A/D converters. The secondary signals are computed based on the primary signals. The input power to the motor drive, as a secondary signal for instance, is computed by using the inverter DC voltage and DC current. Both the primary and the secondary signals often need some kind of modifications before they are used in the control system. These modifications, e.g. filtering and averaging, improve the quality of the signals and subsequently the reliability and the performance of the motor drive. In this section measurement and manipulation of several system signals are presented.

Rotor Position and Speed Detection

The absolute value of the rotor position is needed for the motor speed detection and the axis transformation. The optical encoder is mounted on the motor shaft originally provides $2^{10}$ pulses per revolution. An interpolation scheme which is incorporated in the encoder increases the number of pulses per revolution to $4 \times 2^{10} = 4096$. The encoder interface on the DS1102 board consists of a quadrature decoder which converts the
encoder information to count up/down increments of $4 \times 4096 = 16384$ i.e. $10^{14}$ increments per rotor revolution. These are stored in a 24-bit signed counter. The counter output is scaled to a floating-point value in the range $-1.0 .. 1.0$ by the board. Therefore the counter in each direction (positive or negative) can store a total number of increments belonging to $(2^{24}/2 - 1)/2^{14} = 511.9999$ rotor revolutions. Thus a full counter stores $511.9999 \times 2\pi = 3216.9904 \ldots$ rad. If the scaled counter output is multiplied by this number the actual rotor position in radians is obtained. This is doubled to give the electrical rotor position since the motor has two pole pairs. The encoder index signal is used to clear the counter at each rotor revolution. Therefore an absolute rotor position in the range $0 .. 2\pi$ is obtained. This is depicted in Fig. 6.11 for the case with the motor starting from a stand still condition.

The motor mechanical speed is calculated by the last and present values of the rotor position as:

$$\omega_m(k) = \frac{\theta_m(k) - \theta_m(k-1)}{T_s}$$  \hspace{1cm} (6.1)$$

where $\theta_m(k)$ and $\theta_m(k-1)$ represent current and previous samples of rotor position in mechanical radians and $T_s$ is the sampling time. The speed signal computed by this method may contain high frequency noise as well as low frequency pulsations. The speed pulsation, if present, is a mechanical problem. It is usually caused by vibration as a result of loose mounting of the motor on the base, or by imperfect connection between motor and encoder. However the high frequency noise is the jitter effect or the quantization error. This noise provides a problem as accurate speed control is an
Fig. 6.11 Rotor position.

Fig. 6.12 Filtered motor speed.

Fig. 6.13 Filtered phase current.

Fig. 6.14 Inverter DC input voltage.
essential part of the loss minimization process. The speed signal is passed through a filter to eliminate the noise. The filter provides the average speed signal over the last several (e.g. five) samples. This is done by developing a software routine. The routine allocates five memory locations as a window to store values of the computed speed in the last five samples. At each sampling time the current value of speed is pushed into the window while the most previous value is pulled out of it. Then the content of the window is averaged and the result substitutes the last entered value into the window. By this way the value of filtered speed at each instance is an average of the current unfiltered speed and four previous filtered speed values. This results in a smooth speed signal as seen in Fig. 6.12. This signal is fed back to the speed controller. It is also used in the decoupling circuits after being multiplied by the number of pole pairs to provide electrical speed.

**Phase Currents and Their Components**

The motor phase currents are needed for current controllers. The currents in phase "a" and phase "b" are converted to the proper voltage signals in the range -10 .. 10 volts by the current transducers. These voltages are used as the inputs to the on board A/D converters. The DS1102 is capable of reading the A/D channels by first initializing A/D converters. A software function is then used to read each A/D channel. No hardware filters are required for current signals. However two similar software filters are designed to eliminate the current noise. The first order filters prove to be adequate for this purpose. They provides smooth signals with negligible time delays due to the high speed of the TMS320C31 DSP. A plot of the phase current is shown in Fig. 6.13.

The current components in the rotor reference frame, $i_d$ and $i_q$, are computed by
using Park's transformation. A fast transformation is achieved if (4.3) is modified for most efficient calculation. By substituting \(i_s\) in (4-3) in terms of \(i_a\) and \(i_b\) as \(i_s = -i_a - i_b\) only two phase currents are sufficient to calculate \(i_d\) and \(i_q\). Furthermore by expanding the sin and cos functions of angles \((\theta_e - 2\pi/3)\) and \((\theta_e + 2\pi/3)\) in terms of cos and sin functions of \(\theta_e\) and \(2\pi/3\), only two trigonometric functions instead of four are needed for the transformation. Since the cos and sin functions are the most time consuming operations in a vector transformation, this reduces the time required for the computation of \(i_d\) and \(i_q\) substantially. By making these two modifications the current components \(i_d\) and \(i_q\) are calculated by:

\[
\begin{align*}
    i_d &= i_a \sin(\theta_e) - \frac{1}{\sqrt{3}} (i_a + 2i_b) \cos(\theta_e) \\
    i_q &= i_a \cos(\theta_e) + \frac{1}{\sqrt{3}} (i_a + 2i_b) \sin(\theta_e)
\end{align*}
\] (6-2) (6-3)

where \(\theta_e\) is the electrical rotor angle. These current components are fed back to the decoupling current controller.

**Drive Input Power**

An accurate measurement of DC input power to the inverter is a key factor in the successful implementation of the proposed adaptive loss minimization controller. It is also extremely important that the input power as an input to ALCM is very smooth and almost free of noise. Therefore a particular attention is focused on the measurement, filtering and averaging of this signal to meet the requirements of ALCM. In this way the minimum input power can be controlled at a true minimum value.

The DC input power to the motor drive, \(P_{dc}\), is calculated as a product of the
inverter DC link current $I_{DC}$ and DC bus voltage $V_{DC}$. The bus voltage is reduced to a low voltage value in the range $-10$ to $10$ volts by a voltage transducer. This voltage is then applied to an A/D converter channel on the DSP board. Such a signal is shown in Fig. 6.14. Since $V_{DC}$ is high enough at almost all operating conditions the relative magnitude of noise on this signal is low as it is shown in Fig. 6.14. Therefore the sampled value of $V_{DC}$ is used in the input power calculation with no modification except for proper scaling to compensate the transducer gain and the A/D conversion. However the situation is different with $I_{DC}$. The DC link current is measured through a current transducer and an A/D channel. This signal is shown in Fig. 6.15 (a). The plot shows the current in the transient state as well as the steady state. In both states the signal is quite noisy. If this signal is used in the calculation of the input power it provides an input power signal with very high noise. This noise will interfere with the operation of the ALMC and will not allow the changes in the input power to be monitored. In fact the noise margin is often greater than the changes made by ALMC in the period over which the changes are measured. As a result the noise may cause a wrong detection of changes in the input power. This makes the ALMC entirely ineffective. It may even cause an increase in the input power instead of a decrease.

The noise content is reduced by averaging the resultant input power in a way similar to the averaging of the speed signal described above. The buffer size over which the averaging takes place should be chosen with some consideration. A bigger buffer
Fig. 6.15 Inverter DC current: (a) unfiltered, (b) filtered.

Fig. 6.16 Averaged drive DC input power: (a) computed by unfiltered DC current, (b) computed by filtered DC current.
contributes to the smoothness of the input power. However it causes more delays and reduces the sensitivity of the input power to the change in \( i_d \) which is provided by ALMC. A bigger buffer also occupies more memory space, although this is not a problem since a relatively large amount of on-board memory is provided with DS1102. A reasonable choice can be made for the buffer size fairly easily after a few trials. In any case the input power buffer is a few times larger than the speed buffer mentioned before. Fig. 6.16 (a) shows the input power after being averaged both over the transient and steady states. The plot does not contain spikes because of averaging. However a very high frequency noise provides a margin of error which is a few watts in the steady state. This is still a major problem for proper and accurate operation of ALMC.

The problem of high frequency noise is effectively removed if the DC current is filtered before it is used in the input power calculation. A first order low pass filter is designed for this purpose. The filtered \( I_{dc} \) is shown in Fig. 6.15 (b). If this current instead of unfiltered current of Fig. 6.15 (a) is used in the computation of the input power, a very desirable \( P_{dc} \) is obtained. This is shown in Fig. 6.16 (b). Comparison of Fig. 6.16 (a) and Fig. 6.16 (b) shows the effectiveness of the current filtering. The input power signal obtained by applying the current filtering and the input power averaging meets the requirement of ALMC. It is further elaborated later in this chapter.
6.4 Decoupling Current Controller

The d- and q-axis current controllers and the corresponding decoupling circuits are implemented according to Fig. 4.4. The controller equations are programmed in discrete form as:

\[
\begin{align*}
\nu_{di}(k) &= \nu_{di}(k-1) + (K_{rd}T_s - K_{pd}) \delta i_d(k-1) + K_{pd}\delta i_d(k) \\
\nu_{qi}(k) &= \nu_{qi}(k-1) + (K_{rq}T_s - K_{pq}) \delta i_q(k-1) + K_{pq}\delta i_q(k)
\end{align*}
\]  

(6.3)  

(6.4)

where \(\delta i_d\) and \(\delta i_q\) are current errors, and \(k\) and \(k-1\) denote the current and the previous sampled values respectively. The d- and q-axis voltage commands are then computed by adding the controller outputs, \(\nu_{di}\) and \(\nu_{qi}\), to the steady state values of d- and q-axis voltages as:

\[
\begin{align*}
\nu^*_d(k) &= \nu_{di}(k) + \nu_{dq}(k) \\
\nu^*_q(k) &= \nu_{qi}(k) + \nu_{qq}(k)
\end{align*}
\]  

(6.5)  

(6.6)

where

\[
\begin{align*}
\nu_{dq}(k) &= -\omega_s(k)L_qi_q(k) \\
\nu_{qq}(k) &= \omega_s(k)(\psi_m + L_di_d(k))
\end{align*}
\]  

(6.7)  

(6.8)

and \(\omega_s\) is the electrical speed.

In order to prevent excessive phase currents and voltages, current and voltage limiters are also designed and programmed as parts of current control loops [113].
Fig. 6.17 Experimental results of decoupling current controller performance, $i_d=0$;
(a) commanded values, (b) actual values.

Fig. 6.18 Experimental results of decoupling current controller performance, $i_d=-1$;
(a) commanded values, (b) actual values.
The performance of the decoupling current controller is evaluated experimentally. The results are shown in Fig. 6.17 and Fig. 6.18 for \( i_d = 0 \) and \( i_d = -1 \) respectively. It is seen that the actual current values follow the commanded values precisely. Also the d- and q-axis loops are decoupled since the variations in one current component, \( i_q \), does not interfere with the other current component, \( i_d \).

6.5 Nonlinear Speed Controller (NLSC)

The nonlinear speed controller as discussed before is responsible for two actions depending on the condition of operation. In the transient state, when the motor speed does not match the commanded speed, the controller works as a conventional PI controller trying to reduce the speed error. In this mode of operation the motor speed is changing. Therefore the ALMC is off and \( \delta l_q^* \) is zero i.e. the motor is running under the original flux level commanded by \( l_0^* \). When the speed error reduces to a small value and remains so for a certain period of time as explained in the previous chapter, the steady state condition is monitored by the ALMC. Thus the adjustment of \( i_d^* \) starts by providing the incremental or decrementally changing signal \( \delta i_d^* \). The change in \( i_d^* \) changes the developed reluctance torque and the system loss simultaneously. Therefore the motor speed changes if no adjustment is made to keep the total mechanical torque, applied to the load, constant. The speed controller in this mode of operation modifies the q-axis current command in response to the changing \( \delta i_d^* \) to compensate the speed variation.
Implementation of NLSC

The basic structure of the speed controller is a PI controller. However the input to the PI block is made up of the speed error signal and the output of the ALMC i.e. $\delta i_d^*$. The input to the PI controller is expressed as:

$$\delta \omega_{\text{eff}} = \delta \omega_m (1 + K |\delta i_d^*|)$$

(6.9)

where $\delta \omega_m$ is the speed error signal and $K$ is the non-linearity gain. It should be noticed that the absolute value of the ALMC output is used in (6.9). When the ALMC is off $\delta i_d^*$ is zero. Therefore $\delta \omega_{\text{eff}} = \delta \omega_m$ as in a usual PI speed controller. In the steady state however, where the ALMC produces a non zero output, the input signal to the PI controller is modified by an extra term i.e. $\delta \omega_m K |\delta i_d^*|$. The absolute value of $\delta i_d^*$ increases linearly with time as a result of ALMC action. However $\delta \omega_{\text{eff}}$ changes in such a way that the commanded speed is retained in spite of $\delta i_d^*$. The software implementation of PI controller is straightforward. It gives the current value of the q-axis current command as:

$$i_q^*(k) = i_q^*(k-1) + (K_i T_s - K_p) \delta \omega_{\text{eff}}(k-1) + K_p \delta \omega_{\text{eff}}(k)$$

(6.10)

where (k) and (k-1) stand for the current and the previous sampled values, and $K_i$ and $K_p$ represent the integral gain and the proportional gain of the PI controller respectively.

Performance of NLSC

The nonlinear speed controller was tested extensively before it is used in the loss minimization process. This was done by examining its performance by experiment under a varying d-axis command signal. The objective was to evaluate the ability of NLSC to maintain the motor speed. For this purpose the drive control system is set up with the
nonlinear speed controller but without the ALMC. First the non-linearity gain is set to K=0. This corresponds to a simple PI controller with no provision for the compensation of the changes in \( i_d^* \). The motor is then commanded to run at a certain speed with a constant d-axis command as in Fig. 6.19 (a). Shortly after the steady state speed is achieved, the d-axis command is changed linearly with time by a software routine which is specially developed for the test purpose. As a result \( i_d^* \) undergoes a ramp trajectory similar to the one in an actual loss minimization process. This is shown in Fig. 6.19 (b). The speed increases substantially as a result of a considerable change in the developed torque. The speed may run up unless the changes in \( i_d^* \) is terminated. By doing so the speed decreases. However a second ramp in \( i_d^* \) results in another speed increase. The plots of current components and phase current are also shown in Fig. 6.19 (c) and (d) respectively. Fig. 6.20 (a)-(d) show the result of a similar test with a K=2. The NLSC in this case modifies the q-axis current command in response to the change in \( i_d^* \) as it is seen in Fig. 6.20 (b) and (c). It is seen in Fig. 6.20 (a) that the NLSC reduces the change in the motor speed in spite of a longer period of change in \( i_d^* \). The test is repeated with K=15 and the results are shown in Fig. 6.21 (a)-(d). It is seen that the NLSC effectively compensates the speed variation in spite of the change in \( i_d^* \). A small variation however can be seen in the speed plot once the change in \( i_d^* \) is initiated. But this variation dies out quickly by the action of the NLSC. The speed variation is short in time and low in magnitude and does not effect the loss minimization process as will be shown in the next chapter.
Fig. 6.19 Experimental results of nonlinear speed controller performance, $K=0$. 

(a) Speed (15 rad/sec/div) 
(b) d- & q-axis current commands (2A/div) 
(c) d- & q-axis currents (2A/div) 
(d) Phase current (2A/div) 

dash: speed command 
solid: motor speed 
upper: q-axis 
lower: d-axis
Fig. 6.20 Experimental results of nonlinear speed controller performance, $K=2$. 

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dash: speed command
solid: motor speed

upper: q-axis
lower: d-axis
Fig. 6.21 Experimental results of nonlinear speed controller performance, K=15.
6.6 Adaptive Loss Minimization Controller (ALMC)

In this section the DSP implementation of the ALMC together with the sample test results of the system under the ALMC are presented. The block diagram of the ALMC is shown in Fig. 6.22. This includes a processing unit and the loss minimization algorithm (LMA). The processing unit receives one of the input signals to the ALMC, i.e. the drive input power, and finds out the change in the input power, \( \delta P_{dc} \) during a time window. The LMA, as the core of the ALMC is responsible for making proper decisions based on the signals it receives i.e. \( \delta \omega_m \) and \( \delta P_{dc} \). It produces the ALMC output signal, \( \delta i_d^* \).

**Power Processing Unit**

The input signals to the ALMC consist of the DC drive input power, \( P_{dc} \), and the speed error \( \delta \omega_m \). The motor speed signal and \( P_{dc} \) are subject to the averaging and smoothing processes before they are applied to the ALMC. However \( P_{dc} \) requires further processing before it is used in the loss minimization process. The power processing unit does this processing to find out a new signal \( \delta P_{dc} \). This signal represents the change of \( P_{dc} \) over a certain time interval \( \tau_d \) as it is needed by the adaptive loss minimization algorithm. The time interval \( \tau_d \) depends on the smoothness of \( P_{dc} \) and its sensitivity with respect to \( \delta i_d^* \). It is important to notice that \( \delta P_{dc} \) is calculated at every sampling step. Therefore it represents, at each instance, the difference between the current and the \( n^{th} \) previous values of the drive input power as follows:

\[
\delta P_{dc} (k) = P_{dc} (k) - P_{dc} (k-n)
\]  

(6.11)

129
Fig. 6.22 Block diagram of ALMC.

Fig. 6.23 Flowchart of loss minimization algorithm.
where \( n \) is found from \( \tau_a = nT_s \).

The signal \( \delta P_{dc} \) provides a natural way of monitoring the trend in the drive input power. A negative value of \( \delta P_{dc} \) indicates a falling input power while a positive value indicates a rising input power. Therefore it can be used in search for an optimum d-axis current corresponding to a minimum \( P_{dc} \).

### Loss Minimization Algorithm (LMA)

The loss minimization algorithm is the core of ALMC. It is basically a DSP routine incorporating an arrangement of logic functions. The algorithm monitors the two input signals \( \delta \omega_m \) and \( \delta P_{dc} \) all the time and generates the output signal \( \delta i_d^* \) correspondingly. A simplified flow chart of LMA is shown in Fig. 6.23. Three main jobs can be identified in the flow chart. The determination of the steady state speed, the direction test and the loss minimization. A steady state speed, as it will be defined, must be detected before the direction test is initiated. This test determines the direction of changes in the original d-axis current by which the drive input power decreases. The loss minimization process starts by changing the d-axis current in the proper direction.

The LMA monitors the motor speed error, i.e. the commanded speed minus the actual speed, all the time. When the error is less than a small value for three samples, with a fraction of second sampling interval \( \tau_a \), LMA detects a steady state speed. This is a more restricted definition of the steady state speed than the usual one. It is done by continuously keeping track of the current speed error sample \( \delta \omega_m(k) \) and two previous speed error samples \( \delta \omega_m(k-k_1) \) and \( \delta \omega_m(k-2k_1) \). The number \( k_1 \) is found from \( \tau_a = k_1T_s \) where \( T_s \) is the sampling time of the speed signal. The three speed error values are
compared. If the difference between every two consecutive values are less than a speed error band $\Delta \omega_m$ the steady state speed is assured and the direction test is initiated. A small value for $\Delta \omega_m$ is desired. That way a smooth speed signal needed by the ALMC is generated.

Once the steady state speed is detected the direction test starts. The signal $\delta i_q^*$ is originally equal to zero. The LMA incrementally or decrementally changes this signal in a ramp manner for a constant period of time equal to $\tau_d$. This is done by the accumulation of incremental or decremental changes in $\delta i_d^*$. An incremental change is presented by $\epsilon_d$ in the flow chart. The signal $\delta i_d^*$ is updated at each sampling time $T_s$. Therefore $\epsilon_d$ is a very small value. The direction test period $\tau_d$ is the same period over which $\delta P_{DC}$ is calculated as described before. Therefore at the end of this period $\delta P_{DC}$ gives the amount of change in the input power caused by the change in $\delta i_d^*$. The initial direction of change in $\delta i_d^*$ can be chosen by the programmer by choosing either a positive or a negative value for $\epsilon_d$. If the initial direction causes a value of $\delta P_{DC}$ less than a power band, $\Delta P_{DC}$, then the right direction has already been chosen. Otherwise the power reducing direction is the opposite direction. By this way a proper direction is found regardless of the original flux level. The value of power band determines the accuracy of ALMC. A small band requires a smooth $\delta P_{DC}$.

After the loss minimizing direction is found the LMA continues to make changes in $\delta i_d^*$ with the same slope as long as the condition of $\delta P_{DC} \leq \Delta P_{DC}$ is satisfied as in Fig. 6.24 (a). During this period $P_{DC}$ reduces smoothly and continuously as in Fig. 6.24 (b). The loss minimization period depends on the difference between the original d-axis
current command $i_{d0}^*$ and the optimum value of $i_d^*$. It also depends on the slope of the ramp which $\delta i_d^*$ follows during the loss minimization. This slope is controlled by the size of $\varepsilon_d$. A larger $\varepsilon_d$ results in a steeper ramp and shortens the loss minimization period. Therefore it saves more energy. However there is always a limit for the ramp slope. This limit depends on the ability of the nonlinear speed controller to compensate the speed variations at the onset of the direction test. In practice the performance of the NLSC was found to be excellent and a small value for the input power band can be chosen. This results in the continuation of the loss minimization period until a very accurate minimum loss condition is achieved.

The minimum loss condition is determined when a further change in the $\delta i_d^*$ results in a value of $\delta P_{dc}$ greater than $\Delta P_{dc}$. At this point the loss minimization period is terminated, the direction of change in $\delta i_d^*$ is reversed by substituting $\varepsilon_d$ by $-\varepsilon_d$ and the flow returns to the direction test as shown in Fig. 6.23. Since the motor drive is working on the verge of the minimum loss condition now, a change in the $d$-axis current command in the new direction does not satisfy the condition of $\delta P_{dc} \leq \Delta P_{dc}$. Therefore the direction test is repeated in the opposite direction. Thus the system enters a new mode of operation, the triangular mode, in which the direction test is repeated in an alternating direction and provides a triangular path for $i_d^*$ as shown in Fig. 6.24 (a). The coordinated behaviours of the ALMC and the NLSC provide a smooth motor drive operation. The operating point swings naturally in the vicinity of a true minimum loss condition as shown in Fig. 6.24 (b). The plots of phase current and phase voltage command are also shown in Fig. 6.24 (c) and (d) respectively. From these plots the
Fig. 6.24 Experimental results of ALMC performance.
function of the ALMC can be interpreted as the simultaneous adjustment of motor phase current and voltage to achieve a minimum input power.

**Fast Dynamics**

In order to maintain a fast and robust system dynamics a special software module is developed as part of the ALMC as explained in Chapter 5. The input to this module is the speed error signal and the output is a disabling signal. The signal $\delta\omega_m$ is monitored continuously during the operation of the ALMC. If the absolute value of this signal exceeds a certain limit, $\Delta\omega_m$, a disabling signal returns the original flux level by zeroing $\delta i_d^\ast$. This transfers the NLSC to a PI controller. The motor undergoes a transient period with a desirable dynamics which is provided by the original $i_{d0}^\ast$. The ALMC remains disabled until the steady state speed is detected. The performance of the drive system under this condition is experimentally evaluated. The system is quite stable and the transient response is desirable. The experimental results of this is presented in the next chapter along with many other test results.
CHAPTER 7

Experimental System Performance Evaluation

The system performances were examined by simulation in Chapters 4 and 5. The system components like speed and current controllers, NLSC and the ALMC were also validated separately by laboratory tests as presented in Chapter 6 to make sure that they perform well according to their design specifications. In this chapter however, a comprehensive experimental evaluation of the complete system is carried out. Many different operating conditions are considered and several control situations are adapted. An extensive set of test results are presented and their salient points discussed. These include the motor speed response under different loads and speed commands, load disturbance test, system performance under different modes of the ALMC at different speed and load values and the system dynamics during the transition from one steady state condition to another. The experimental results confirm the validity of the proposed adaptive loss minimization control.
7.1 Motor Speed Response

The speed responses were obtained for many speed and load values during the experimental course of the research. Three sets of sample results which cover a wide range of possibilities the motor may face in practice are presented here. Fig. 7.1 shows the experimental results of motor speed at no load without the ALMC. The commanded and actual speed plots are given in Fig. 7.1 (a). The PI controller parameters are adjusted such that no overshoot occurs. The d-axis current and the q-axis current commands are shown in Fig. 7.1 (b). It is evident that the decoupling circuit works well since the d-axis current remains at its commanded value \( i_d^* = 0 \) in spite of rapid changes in the q-axis current. A plot of the phase current is shown in Fig. 7.1 (c). The peak value of current does not exceed the current rated value of 3 A due to a moderate swiftness of the speed signal. The motor torque is plotted in Fig. 7.1 (d). The torque is calculated by the DSP software by using the conventional torque equation and the measured current components. This torque is of course approximate since all the motor parameters used in the torque equation i.e. \( \psi_m \), \( L_d \) and \( L_q \) are affected by operating condition. However at \( i_d = 0 \) the torque equation gives a better approximation of the actual developed torque due to the absence of saturation in the rotor bridge.

The speed responses are also obtained under a load applied by the dynamometer and the results are depicted in Fig. 7.2. The settling time of the speed signal in this case is a little longer and the steady state values of the q-axis current command, phase current and torque are higher as expected, Fig. 7.2 (a)-(d). A minor variation can be seen at the
Fig. 7.1 Experimental results of step response of speed at no load.
Fig. 7.2 Experimental results of step response of speed under load.
Fig. 7.3 Experimental results of change in speed command.
early moments of the speed signal transient. This is due to the saturation of the q-axis current controller due to insufficient DC bus voltage. The limitation is imposed by the existing DC power supply and is more serious when a higher speed command is applied. However as it is evident the saturation is removed quickly as a result of speed build up and the commanded speed value is reached precisely.

A change in the speed command is examined next. Fig. 7.3 shows the speed response of the motor, at no load, where the speed command is increased stepwise after the initial speed command is met. The speed controller withstands this situation reasonably well. As a result the new steady state speed is achieved after a low overshoot as in Fig. 7.3 (a). This test is done under the flux weakening condition, i.e. \( i_d < 0 \) as seen in Fig. 7.3 (b), to overcome the problem of saturation in the current controllers. This is achieved because of a higher phase current which corresponds to a lower motor voltage and subsequently a lower DC bus voltage. The comparison of Fig. 7.1 (c) and Fig. 7.3 (d) shows the elevated phase current in the flux weakening case. The excellent performance of the current controllers are validated by current plots in Fig. 7.3 (b) and (c) where \( i_d \) and \( i_q \) closely follow \( i_d^* \) and \( i_q^* \) respectively.

### 7.2 Load Disturbance Test

When the IPM motor is running at the steady speed with the ALMC deactivated intentionally a load is suddenly applied to the motor and removed shortly afterwards. The experimental results are shown in Fig. 7.4. The speed plot is shown in Fig. 7.4 (a)
Fig. 7.4 Experimental results of load disturbance test.
where a dip and a rise can be seen as results of the application and the removal of the load. The commanded and actual values of d- and q-axis currents are shown in Fig. 7.4 (b) and (c). Again the ability of current controllers to handle the situation properly is evident. A plot of the phase current is also shown in Fig. 7.4 (d).

7.3 Loss Minimization Control

The complete motor drive control system is examined experimentally in this section to evaluate the system performance under the ALMC. The motor operation is monitored over a relatively long time and different system variables are traced and plotted as in Fig. 7.5 for a no load condition. Fig. 7.5 (a) shows the commanded and the actual speed signals. The steady state condition as defined in Chapter 6 is detected by the ALMC at about t=2.5 seconds. The direction test is started next and finds out the direction of change in $i_d^*$ which reduces the input power. Since the motor exhibits significant saturation at the bridge between the magnets on the rotor, the input power reduces when the d-axis current is less negative. Therefore the ALMC continues to increase $i_d^*$ as seen in Fig. 7.5 (b). The NLSC modifies $i_q^*$ correspondingly in order to maintain the original operating point as seen in the same figure. The d- and q-axis current components follow their commanded values precisely as in Fig. 7.5 (c). As a result the drive DC input power reduces continuously and smoothly towards a minimum value experiencing about 30% reduction in about two seconds as seen in Fig. 7.5 (e). The DC link current is also shown in Fig. 7.5 (d). Notice that at the beginning of the
Fig. 7.5 Experimental results of system performance under ALMC.
Fig. 7.5 Continued.
loss minimization process the input power reduces rapidly. As the input power approaches its minimum value it becomes more flat and the slope of input power reduces. Once the input power reaches the vicinity of a minimum value the changes in the input power during a specific period of time as described before (0.5 second in this case) exceeds a small negative value (a few watts). At this time ALMC goes to the triangular mode. The d-axis current command follows a triangular trajectory as in Fig. 7.5 (b) with the corresponding modification in $i_d^*$. This ensures the system operation at a minimum input power even when the system parameters or operating point changes slowly.

In the case of a large change in the operating point, e.g. a major change in the speed command as in Fig. 7.5 (a), the ALMC is deactivated and the d-axis current command returns to its original value as in Fig. 7.5 (b). The speed controller acts as a simple PI controller and a new steady state speed is achieved after a fast transient state. The ALMC becomes active again and a new minimum loss condition corresponding to the new speed is obtained in a short time as seen in Fig. 7.5 (a)-(e).

The same test presented in Fig. 7.5 is repeated at a light load condition. The results are shown in Fig. 7.6 (a)-(e) where more energy saving is achieved. In general the amount of saving in energy depends on the original operating condition including the d-axis current.

The experimental results presented in this chapter confirm the simulation results presented in Chapter 5 and validate practically the proposed adaptive loss minimization control as a new and effective strategy. The experimental results prove this strategy as a fast, accurate and smooth on-line loss minimization control approach. This
Fig. 7.6 Experimental results of system performance under ALMC at loading condition.
combination of features improves the loss minimization processes in the existing applications. Furthermore, it extends the practice of loss minimization or efficiency optimization to a wide variety of new applications like high performance drives and electric vehicle beyond the limitations of existing methods.
CHAPTER 8

Conclusions

In this thesis a novel on-line adaptive loss minimization control strategy was introduced. Based on this strategy an adaptive loss minimization controller (ALMC) was proposed. The analysis, design, simulation, implementation and extensive test results of a current vector control system, including the ALMC, when applied to an inverter-fed IPM motor drive were presented. The loss control strategy was shown to overcome the limitations and drawbacks of the existing off-line and on-line loss minimization controllers by providing a smooth and accurate minimum drive input power in a short time at any desirable output power. These features allow the application of the ALMC to a new class of motor drives with both system parameter variations and frequent changes in the operating condition like in EVs.

The EV was reviewed at the opening chapter of this thesis as a perfect example of the above mentioned class of applications. The IPM motor was chosen as a suitable candidate for EV applications due to its inherent advantages. Before dealing with the ALMC, the IPM motor performances under an off-line loss minimization controller were presented after the potential energy saving at the minimum loss condition was studied. A detailed IPM machine model in the steady state was introduced for this study. The practical significance of the motor parameter variations and the difficulties associated
with the modelling of these variations were discussed. To avoid these difficulties it was concluded that an on-line loss minimization approach may be employed since this approach does not require a machine model. However an extensive literature review of loss minimization controls showed the drawbacks of the currently available on-line methods i.e. a long search time, torque disturbances, and a sluggish response or parameter dependency of the compensator responsible for maintaining the motor output power. These drawbacks were overcome in the proposed loss minimization control strategy to allow the application of loss minimization control to the situations requiring smooth operation in the face of parameter variations and frequent changes in the operating point.

8.1 Contributions

In this thesis the following improved and new analysis, models, methods and systems are successfully introduced.

i) The minimum loss operation of IPM motors are presented by taking into account the parameter variations under different operating points. The potential energy saving under minimum loss condition with and without parameter variations is analyzed over a typical EV driving cycle.

ii) Three mathematical machine models are developed for analysis of different aspects of IPM motors. A steady state model is presented which includes both copper loss and iron loss. The parameter variations are incorporated in this model in terms of motor
variables. This model is used to find out the motor electrical loss at different operating points. The model is expanded later to present a detailed dynamic model of IPM motors. Based on a state space version of the latter model a simulation program is developed to analyze the dynamic performance of IPM motor drives in many different situations including the case with the ALMC. A linearized model is also presented and used in the analysis and design of the IPM motor control system. In contrast to the conventional model of PM motors this model is derived at a non-zero value of the d-axis component of the stator current. Therefore both the reluctance and the magnet components of torque are presented in the model. Using this model it was possible to analyze quickly and conveniently the motor operation under any flux level. This model can be used in the iterative design of a speed controller without referring to the extensive model. It was also used in the analysis and design of the NLSC.

iii) An on-line adaptive loss minimization control strategy is introduced for electric motor drives. The fundamental idea in this strategy, in contrast to all other on-line loss minimisation methods, is a continuous pattern of change in the d-axis current (the motor flux) to find a minimum input power. This pattern of change provides significant improvements in the loss minimization process and in the motor drive performance. Firstly the torque disturbances caused by a stepwise change in the d-axis current is avoided thus providing a smooth system performance required in many applications. More importantly the transient period caused by each step change in the d-axis current is eliminated. This in conjunction with a fast compensation provided by the NLSC results in the removal of the wait period after each step needed for the passage of the transient
period. Therefore a fast loss minimization mode and a short search time to reach the minimum input power are achieved. This short time makes it possible to apply the ALMC to motor drives with frequent changes in the operating point like in EVs. The continuous change in the d-axis current also reduces the drive input power continuously. Thus a minimum input power is detected accurately in the case that the changes in the input power are monitored continuously.

iv) The motor speed variation in response to a change in the d-axis current is analyzed with the help of the linearized machine model mentioned above. Gaining new insights into the nature of the speed compensation process a concept of forced compensation is introduced. Based on this concept a nonlinear speed controller, NLSC, is proposed and built to maintain the drive output power constant while the input power is being reduced by the ALMC. The NLSC is a usual PI controller in transient state. In steady state however, the controller works in a forced compensation mode, speeding up the adjustment of the q-axis current (magnet torque) according to the change in the d-axis current (reluctance torque), thus maintaining the total torque and subsequently the motor speed. This is achieved by making use of a brilliant opportunity i.e. that the ALMC starts changing the d-axis current only after a steady state is detected.

v) A complete IPM motor drive control system including the ALMC and the NLSC is built using a TMS320C31 DSP based board. The extensive experiments on the laboratory set up at different motor speeds and loading conditions prove the validity of the proposed on-line adaptive loss minimization control strategy.

vi) The analytical, simulation and experimental results bring about a new insight into the
saturation of the iron bridges between rotor magnets. It is shown in the literature, for line start IPM motors, that the bridge saturation caused by the leakage flux at lower terminal voltages, results in a distortion in the air gap flux and hence an increase in the iron loss. It is shown in this thesis that a similar increase in the iron loss occurs in inverter-fed IPM motors as a result of a negative d-axis current. This is because a negative d-axis current in a vector control inverter-fed IPM motor corresponds to a reduced terminal voltage in a line start IPM motor. In both cases a reduced flux in the stator yoke is accompanied by an increased leakage flux in the rotor bridges and flux harmonics in the air gap. The overall result in any case is an increase in the iron loss.

8.2 Future Perspective

This work, as a step for both the improvement of loss minimization control of electric motor drives and the expansion of its application, provides new opportunities for further research and development. The application of the proposed loss minimization control strategy to other types of motors is a natural extension of this thesis. Induction and synchronous reluctance motor drives are the best candidates in this regard.

It was addressed briefly that the computing requirements for the digital implementation of the proposed strategy can easily be met by many processing hardware available in the market. However no special emphasis was placed on the conservative use of the computing hardware. The fast DSP and the extensive hardware features available on the specific board used in the implementation phase of the this research surpass by
far the requirements of the present control system in terms of speed of computation, memory, etc. In a commercial product however, the use of such a DSP board may not be justified due to economic concerns. It is worth trying therefore the design and implementation of the proposed control strategy with these concerns in mind. Apart from a possible development of an efficient version of the control software a number of other modifications can be made to reduce the computing requirements of the system while saving its fundamental advantages mentioned before. A modification in the input power processing unit of the ALMC for instance may reduce both the processing job and the memory needed for the control system implementation.

A simple pattern of continuous change in the d-axis current i.e. a ramp is examined in this thesis. Other continuous patterns may also be chosen. In contrast to the last suggestion the control system computation in this case becomes more demanding. However, in return it may be possible to achieve better performance and meet the demands of some other applications.

In the present work the adaptive loss minimization control strategy is applied to an experimental motor drive to prove the validity of new concepts and to establish the design procedure. The application of same strategy to a prototype EV motor drive will provide a better evaluation of the control strategy and show the actual energy saving possible in a practical situation.
References


APPENDIX A

Motors Specifications

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<tr>
<th>Motor #1</th>
<th>Motor #2</th>
<th>Motor #3 (experimental motor)</th>
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<td>Rated speed, rpm</td>
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<tr>
<td>B, Viscous coefficient, Nm/rad/sec.</td>
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<td>0.0008</td>
</tr>
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</table>
APPENDIX B

Calculation of Iron Loss Resistance

Assuming a sinusoidal flux, a well known representation of the iron loss is:

\[ W_{Fe} = K_e f^2 \Phi_{Max}^2 + K_h f \Phi_{Max}^2 \]  \hspace{1cm} (B1)

where the first term expresses the eddy current loss and the second term represents the hysteresis loss. \( K_e \) and \( K_h \) are constants depending on the core material and the core volume. \( K_e \) also depends on the lamination thickness. \( \Phi_{Max} \) and \( f \) represent the maximum total flux and the inverter frequency respectively. The iron loss can also be obtained from the d-q equivalent circuit in Fig. 3.1 as:

\[ W_{Fe} = \frac{\nu_{2d}^2}{R_c} + \frac{\nu_{2q}^2}{R_c} = \frac{\omega^2 \Phi_{Max}^2}{2R_c} \]  \hspace{1cm} (B2)

\( R_c \) is obtain from eqn’s (B1) and (B2) as:

\[ R_c = \frac{2\pi^2}{K_e + K_h/f} \]  \hspace{1cm} (B3)

At the rated frequency, \( f_n \) (corresponding to the rated rotor speed), the iron loss resistance becomes:

\[ R_{en} = \frac{2\pi^2}{K_e + K_h/f_n} \]  \hspace{1cm} (B4)
From eqn's (B3) and (B4)

\[ R_c = R_{cn} \frac{K + 1/f_n}{K + 1/f} \]  \hspace{1cm} (B5)

where \( K = K_c/K_b \). Having \( R_{cn} \) and \( f_n \) and assuming the value of \( R_c \) at maximum speed with \( f = 3f_n \) equal to \( 2R_{cm} \), yields \( K = 3f_n \). Substituting this value into eqn. (B5) the iron loss resistance is determined as a nonlinear function of \( f \) [77].
APPENDIX C

Minimum Loss Program

The minimum loss program is developed by using the package Mathcad™. The following equations are used:

\[ \text{iq} = \frac{T}{(P_s \cdot f_a - P_n \cdot L_d \cdot \text{id} - P_n \cdot L_d \cdot \text{id} \cdot \text{ro})} \]

\[ \text{iq} = \frac{1}{(P_n \cdot (\text{id} \cdot Lqn))} \left[ \frac{(-P_n \cdot f_a - P_n \cdot \text{iid} \cdot Ld) - P_n \cdot \text{iid} \cdot Lqn \ldots}{\left(\frac{P_n \cdot \text{iid} \cdot f_a^2 - 2 \cdot P_n \cdot \text{iid} \cdot f_a \cdot \text{iid} \cdot Ld}{\ldots} + \left(-2 \cdot P_n \cdot \text{iid} \cdot f_a \cdot \text{iid} \cdot Lqn\ldots\right) + \left(P_n \cdot \text{iid} \cdot Ld^2 - 2 \cdot P_n \cdot \text{iid} \cdot Ld \cdot Lqn \ldots\right) \right] \right] \]

\[ W_{fe} = \frac{1}{R_c} \left[ w^2 \cdot (r_o \cdot L_d \cdot \text{iq})^2 - w^2 \cdot (f_a \cdot L_d \cdot \text{iq})^2 \right] \]

\[ W_{cu} = R_a \left[ \left( \text{id} - \frac{w \cdot r_o \cdot L_d \cdot \text{iq}}{R_c} \right)^2 + \left[ \text{iq} - \frac{w \cdot (f_a - L_d \cdot \text{iq})}{R_c} \right]^2 \right] \]

\[ W_e = W_{fe} + W_{cu} : \]

\[ W_e = \frac{1}{R_c} \left[ w^2 \cdot (r_o \cdot L_d \cdot \text{iq})^2 + w^2 \cdot (f_a - L_d \cdot \text{iq})^2 \right] \ldots \]

\[ + R_a \left[ \left( \text{id} - \frac{w \cdot r_o \cdot L_d \cdot \text{iq}}{R_c} \right)^2 + \left[ \text{iq} - \frac{w \cdot (f_a + L_d \cdot \text{iq})}{R_c} \right]^2 \right] \]
Substitute for \( r_0 = \text{Ld/Lq} \):

\[
W_e = \frac{1}{Rc} \left[ \frac{w^2}{Ld} \left( \frac{Lq - 1}{2} \frac{Lq}{Ld \cdot \text{lam}} \right) \cdot \text{Ld}^2 \cdot \text{iod}^2 + \frac{w^2}{Ld} \cdot \left( \text{fa} - \text{Ld} \cdot \text{iod} \right)^2 \right] \cdot \frac{\text{Ld}^2}{\text{iod}^2} + \frac{\text{Ld}^2}{\text{iod}^2} \cdot \frac{\text{w}^2}{Rc} \cdot \left( \text{Ld} \cdot \text{iod} \right)^2 + \frac{\text{Ld}^2}{\text{iod}^2} \cdot \frac{\text{w}^2}{Rc} \cdot \left( \text{fa} \cdot \text{Ld} \cdot \text{iod} \right)^2
\]

Collect on \( \text{iod} \) (a symbolic operator of Mathcad arranging the equation in terms of descending orders of \( \text{iod} \)):

\[
W_e = \frac{1}{4Rc} \left[ \frac{w^2}{Ld} \cdot \left( \frac{Lq - 1}{2} \frac{Lq}{Ld \cdot \text{lam}} \right) \cdot \text{Ld}^2 \cdot \text{iod}^2 + \frac{w^2}{Ld} \cdot \left( \text{fa} - \text{Ld} \cdot \text{iod} \right)^2 \right] \cdot \frac{\text{Ld}^2}{\text{iod}^2} \cdot \frac{\text{w}^2}{Rc} \cdot \left( \text{Ld} \cdot \text{iod} \right)^2 + \frac{\text{Ld}^2}{\text{iod}^2} \cdot \frac{\text{w}^2}{Rc} \cdot \left( \text{fa} \cdot \text{Ld} \cdot \text{iod} \right)^2
\]

Substitute for \( \text{iod} \) to get \( W_e \) as a function of \( \text{iod} \):

\( \text{Results too big to be printed} \)

Differentiate w.r.t. \( \text{iod} \):

\( \text{Results too big to be printed} \)
Introduce motor parameters:

\[ f_a = 0.1077 \quad P_n = 2 \quad T_{mech} = 0.0588 \]
\[ n_n = 2000 \quad R_a = 0.572 \quad R_{cn} = 240 \quad L_d = 0.00872 \]
\[ n_m = 6000 \quad K_1 = 200 \quad K = 200 \quad w_n = n_n \cdot \frac{\pi}{60} \cdot P_n \]
\[ P_o = 1.4 \cdot 1.67 \cdot \frac{w_n}{P_n} \quad I_{am} = 6.5 \quad L_{qn} = 0.02278 \cdot 2 \]

Solve the derivative of \( W_e \) to find optimal \( i_{od} \):

\[ i_{od} = 5 \quad R_c = 523 \quad w = 700 \quad T = 1.6 \]

Given

\[ R_c = \frac{R_{cn}}{K \cdot 2 \cdot \pi \cdot \left( 1 + \frac{K \cdot 2 \cdot \pi \cdot i}{w_n} \right)} \quad w = \frac{n_m}{60} \cdot 2 \cdot \pi \cdot P_n \cdot i \]

\[ T = \text{if} \left( i < 10, 1.67 \cdot 1.4, \frac{P_o}{w} \right) \quad i_{od} = 0 \]

\[ F(i) = \text{Find}(i_{od}, R_c, w, T) \]

\[ i = 1 \ldots 30 \quad i_{od_i} = F(i) \quad T_i = F(i) \]

Calculate loss and efficiency with the optimal current vector

\[ w_i = \frac{n_m}{60} \cdot 2 \cdot \pi \cdot P_n \cdot i \]
\[ r_{qi} = \frac{L_{qn}}{L_d} \cdot \left( 1 - \frac{i_{od_i}}{2 \cdot I_{am}} \right) \]
\[
Rc_i = \frac{Rcn}{1 - \frac{K \cdot 2 \cdot \pi}{wn}}
\]

\[
Wfe_i = \frac{1}{Rc_i} \left[ \left( w_i \right)^2 \cdot \left( ro_i \cdot Ld \cdot ioq_i \right)^2 - \left( w_i \right)^2 \cdot \left( fa - Ld \cdot iod_i \right)^2 \right]
\]

\[
Wcu_i = Ra \left( \frac{iod_i - \frac{w_i \cdot ro_i \cdot Ld \cdot ioq_i}{Rc_i}}{Rc_i} \right)^2 \cdot \left( \frac{w_i \cdot (fa - Ld \cdot iod_i)^2}{Rc_i} \right)
\]

\[
We_i = Wfe_i - Wcu_i
\]

\[
wr_i = \frac{w_i}{Pn}
\]

\[
wrn = \frac{nn \cdot 2 \cdot \pi}{60}
\]

\[
Wm_i = Tmec \cdot wr_i
\]

\[
Wl_i = We_i + Wm_i
\]

\[
w_{i-1} = \frac{wr_i}{wrn}
\]

\[
P_i = T_i \cdot wr_i
\]

\[
Eff_i = \frac{P_i}{P_i + Wl_i} \cdot 100
\]

\[
id_i = iod_i - \frac{w_i \cdot ro_i \cdot Ld \cdot ioq_i}{Rc_i}
\]

\[
jq_i = ioq_i + \frac{w_i \cdot (fa - Ld \cdot iod_i)}{Rc_i}
\]

\[
Ia_i = \sqrt{\left( id_i \right)^2 + \left( jq_i \right)^2}
\]

\[
Va_i = \sqrt{\left( Ra \cdot id_i - w_i \cdot ro_i \cdot Ld \cdot ioq_i \right)^2 + \left[ Ra \cdot jq_i - w_i \cdot (fa - Ld \cdot iod_i) \right]^2}
\]

\[
Lq_i = Lqn \cdot \left( 1 - \frac{jq_i}{2 \cdot Iam} \right)
\]
Import results from the CPL control and calculate efficiency improvement:

\[ \text{DelEff}_i = \text{Eff}_i - \text{EEI}_i \]

Calculate percentage of loss reduction:

\[ \text{DelWe}_i = \frac{\text{WeI}_i - \text{We}_i}{\text{We}_i} \times 100 \]
APPENDIX D

Controller Design Programs

1. Current Controller Design

This is a matlab program for the design of the d-axis current controller. A simulation program may be initiated at the end of this program.

Rs = 1.93;
Ld = 0.04244;
Lq = 0.07957;
K1 = 60.0; % K1=KId*KPd

%M=logspace(2.4);
NUM=[1];
DEN=[Ld Rs];
figure(1);
%figure(2);
%bode(NUM,DEN,w);
%princsys(NUM,DEN);
figure(2);
%step(NUM,DEN);
%damp(DEN);

num=[1 K1];
den=[Ld Rs 0];
subplot(2,2,2);
axis('square')
rlocus(num,den);
sgrid;
axis([-100 0 -40 40]);
[KPd,poles]=rlocfind(num,den);
damp(poles);
[numc,denc]=feedback(KPd*num,den,1,1);
damp(denc);
[mag,phase,M]=bode(KPd*num,den);
subplot(2,2,4); step(numc,denc);
figure;
margin(mag,phase,M);
princsys(KPd*numc,denc);
KId=K1*KPd;
KPd KId

%[t,x,y]=rk45('pidB',0.02,[],[1.e-4,1.e-5,1.e-4]);
II. Speed Controller Design

% This is a matlab program for the design of the speed controller.

NUM=[0.942];
DEN=[0.003 0.0008];
K1=3;
%K1=KI/KP,
K2=KP*KTq
KTq=0.942;
num=[1 K1]; den=[0.003 0.0008 0];
subplot(2,2,2);
axis('square')
rlocus(num,den);
sgrid;
axis([-10 0 -5 5]);
[K2,poles]=rlocfind(num,den);
damp(poles);
[numc,denc]=feedback(K2*num,den,1,1);
damp(denc);
M=logspace(1,4);
[mag,phase,M]=bode(K2*num,den);
subplot(2,2,4);
step(numc,denc);
figure;
%M=logspace(1,4);
margin(mag,phase,M);
grid;
printsys(K2*numc,denc);
KP=K2/ KTq;
KI=K1*KP;
KP
KI

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APPENDIX E

System Block Diagram for Simulation
APPENDIX F

DSP Controller Board

I. DS1102 Board Data Sheet [108]

<table>
<thead>
<tr>
<th>Processor</th>
<th>Texas Instruments TMS320C31 floating-point DSP. running at 40 MHz clock rate and 50 ns cycle time. Two 32-bit on-chip timers/event counters. On-chip bidirectional 8 Mbaud serial link. On-chip DMA. 4 interrupt lines.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Memory</td>
<td>128K x 32-bit zero wait state memory. 2K x 32-bit on-chip memory.</td>
</tr>
<tr>
<td>16-bit ADCs</td>
<td>± 10 V input range. 10 μs conversion time. ± 5 mV offset voltage. ± 0.25 % gain error. 4 ppm/K offset drift. 25 ppm/K gain drift. &gt; 80 dB signal to noise ratio.</td>
</tr>
<tr>
<td>12-bit ADCs</td>
<td>± 10 V input range. 3 μs conversion time. ± 5 mV offset error. ± 0.5% gain error. 4 ppm/K offset drift. 25 ppm/K gain drift. &gt; 65 dB signal to noise ratio.</td>
</tr>
<tr>
<td>DACs</td>
<td>± 10 V output range. 4 μs settling time. ± 5 mV offset error. ± 0.5% gain error. 5 mA output current. 13 ppm/K offset drift. 25 ppm/K gain drift.</td>
</tr>
</tbody>
</table>
| **Slave-DSP** | Texas Instruments TMS320P14 DSP.  
25 MHz clock rate, 160 ns cycle time.  
32-bit arithmetic unit.  
4K x 16-bit on-chip PROM containing firmware.  
4K x 16-bit external program RAM.  
256 x 16-bit on-chip data RAM.  
Bit selectable 16-bit I/O port.  
6 high precision PWM outputs.  
  
  
  
  
  
event manager with capture inputs and compare outputs. |
| **Incremental encoder interface** | Fourfold pulse multiplication.  
8.3 MHz maximum count frequency.  
Three stage digital pulse filter.  
24-bit position counter.  
5 V / 200 mA sensor supply voltage. |
| **Host-Interface** | Four 16-bit and three 8-bit I/O ports in the 64K host I/O space.  
Memory and I/O are accessible by the host even while the DSP is running.  
DSP-host and host-DSP interrupts. |
| **JTAG-Interface** | On board test bus controller and emulator connector. |
| **Physical size** | 160 mm x 107 mm x 20 mm.  
Requires one half length PC-slot. |
| **Power supply** | +5 V ± 10 %, 1.5A  
±12 V ± 5 %, 100mA |
II. TMS320C31 DSP Functional Block Diagram [109]