A COMPARATIVE PERFORMANCE STUDY OF FOUR-POLE INDUCTION MOTORS AND SYNCHRONOUS RELUCTANCE MOTORS IN VARIABLE SPEED DRIVES
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ABSTRACT

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A comparative performance study of four-pole induction motors and synchronous reluctance motors in variable speed drives

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Design aspects of the Transversally Laminated Anisotropic (TLA) Synchronous Reluctance Motor (SynRM) are studied and the machine performance analysis compared to the Induction Motor (IM) is done. The SynRM rotor structure is designed and manufactured for a 30 kW, four-pole, three-phase squirrel cage induction motor stator. Both the IM and SynRM were supplied by a sensorless Direct Torque Controlled (DTC) variable speed drive. Attention is also paid to the estimation of the power range where the SynRM may compete successfully with a same size induction motor. A technical loss reduction comparison between the IM and SynRM in variable speed drives is done.

The Finite Element Method (FEM) is used to analyse the number, location and width of flux barriers used in a multiple segment rotor. It is sought for a high saliency ratio and a high torque of the motor. It is given a comparison between different FEM calculations to analyse SynRM performance. The possibility to take into account the effect of iron losses with FEM is studied. Comparison between the calculated and measured values shows that the design methods are reliable.

A new application of the IEEE 112 measurement method is developed and used especially for determination of stray load losses in laboratory measurements. The study shows that, with some special measures, the efficiency of the TLA SynRM is equivalent to that of a high efficiency IM. The power factor of the SynRM at rated load is smaller than that of the IM. However, at lower partial load this difference decreases and this, probably, brings that the SynRM gets a better power factor in comparison with the IM. The big rotor inductance ratio of the SynRM allows a good estimating of the rotor position. This appears to be very advantageous for the designing of the rotor position sensor-less motor drive.

In using the FEM designed multi-layer transversally laminated rotor with damper windings it is possible to design a directly network driven motor without degrading the motor efficiency or power factor compared to the performance of the IM.

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Keywords: Induction motor, synchronous reluctance motor, transversally laminated synchronous reluctance motor.
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Joutseno, June the 6th, 2003

Jorma Haataja
To my wife Marjo and
my children Katri, Karri and Petra
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ABBREVIATIONS AND SYMBOLS

Roman letters

$A$  Slope of the zero torque line
$a$  Ratio of insulation
$B, b$  Air-gap flux density
$C$  Intercept with the zero torque line
$D$  Stator bore diameter
$d$  Sheet thickness
$e_b$  Width of flux barrier
$e_{m}$  Electro motive voltage vector
$f$  Frequency
$I_0$  No-load current
$I_s$  Stator current’s RMS value
$i_s$  Stator current vector
$I_N$  Rated current
$K_c$  Carter factor
$k_e$  External loss factor
$k_i$  Stacking factor
$k_h$  Hysteresis loss factor
$L' l$  Stator stack effective length
$l_b$  Length of flux barrier
$l_d$  Per unit value
$I_b$  Base current
$I_m$  Magnetising current
$i_{md}$  Direct axis magnetising current
$i_{mq}$  Quadrature axis magnetising current
$i_s$  Stator current vector
$L$  Inductance
$l_b$  Base inductance
$L_d$  Direct axis inductance, $L_d=L_{sd}+L_{md}$
$l_d$  Per unit d-axis inductance
$L_m$  Magnetising inductance
$L_{md}$  Direct axis magnetising inductance
$L_{mq}$  Quadrature axis magnetising inductance
$L_{sar}$  Stator leakage inductance
$L_{sd}$  Direct axis stator leakage inductance
$L_{sq}$  Quadrature axis stator leakage inductance
$L_r$  Rotor leakage inductance
$l_q$  Per unit q-axis inductance
$m$  Number of phases
$N$  Number of turns in series per stator winding
$P_t$  Load independent losses
$P_L$  Load-dependent losses
$P_N$  Rated power
$p$  Number of pole pairs
$p_{out}$  Per unit output power
$Q_s$  Number of stator slots
$R_{cd}$  $R_{cq}$  d-, q-axis iron loss resistances
$R_s$  Stator phase winding resistance
$R_r$  Rotor resistance
$R$  Radius of rotor
$r$  Correlation factor
$r_{rr}$  Rotor radial rib width
$r_{tg}$  Rotor tangential rib width
$T_b$  Base torque
$T_e$  Torque
$t$  Time
$t_e$  Per unit torque
$U_b$  Base voltage
$u_s$  Stator voltage vector
$V$  Magnetic voltage
$W_{ins}$  Sum of the widths of the flux barrier layers
$W_{iron}$  Sum of the widths of the iron layers (flux guides)
$W_{LL}$  Stray-load loss
$W_{tot}$  Insulation ratio

Greek letters
$\delta$  Load angle
$\delta_e$  Equivalent air-gap length
$\delta_m$  Mechanical air-gap length
$\delta_i$  Internal power angle
$\delta_s$  Effective air-gap length
$\eta$  Efficiency
$\phi$  Angle between the voltage and current vectors (as in $\cos \phi$)
$\kappa$  External current angle
$\kappa_m$  Internal current angle
$\mu_0$  Permeability of vacuum
$\theta_r$  Rotor angle
$\sigma$  Conductivity
$\tau$  Pole span
$\tau_p$  Pole pitch
$\omega$  Angular velocity
$\xi_1$  Fundamental wave winding factor
$\psi$  Flux linkage vector
$\psi_D$  Direct axis damper winding flux linkage
$\psi_m$  Air-gap flux linkage
$\psi_{ind}$  Direct axis component of the magnetising flux linkage
$\psi_{im}$  Quadrature axis component of the magnetising flux linkage
$\psi_{id}$  Direct axis component of the rotor flux linkage
$\psi_{iq}$  Quadrature axis component of the rotor flux linkage
$\psi_{sd}$  Direct axis component of the stator flux linkage
$\psi_{sq}$  Quadrature axis component of the stator flux linkage
$\psi_Q$  Quadrature axis damper winding flux linkage
$\theta_i$  Rotor angle
$\zeta$  Saliency ratio
Subscripts
b  Base value
cd  Direct axis iron loss component
cq  Quadrature axis iron loss component
d  Direct axis
e  External
max  Maximum value
min  Minimum value
Loss  Losses
q  Quadrature axis
rb  Rotor back
rt  Rotor tooth
s  Stator, a quality related to stator
sb  Stator back
st  Stator tooth

Superscripts
g  General reference frame, general coordinates
s  Stator reference frame, stator coordinates
r  Rotor reference frame, rotor coordinates

Acronyms
AC  Alternating Current
ALA  Axially Laminated Anisotropic
CEMEP  Comité Européen de Constructeurs de Machines Electriques et d'Electronique
de Puissance (European committee of manufacturers of electrical machines and power electronics)
DTC  Direct Torque Control
EC  European Commission
EMI  Electro Magnetic Interference
FEM  Finite Element Method
GTO  Gate Turn Off (Thyristor)
IGBT  Insulated Gate Bipolar Transistor
IGCT  Integrated Gate Commutated Thyristor
IEA  International Energy Agency
IM  Induction Motor
FC  Frequency Converter
LUT  Lappeenranta University of Technology
TLA  Transversally Laminated Anisotropic
SynRM  Synchronous Reluctance Motor
VSD  Variable Speed Drive
1 INTRODUCTION

The squirrel cage induction motor (IM) in industrial drives has maintained its top position for decades. The era of frequency converters began as it was aimed at speed control of squirrel cage machines. The so called scalar control of IM was, in principle, easy compared to the vector control needed in magnetically unsymmetrical machines. The IM scalar control is still valid and the most popular alternative for most of the variable speed drive (VSD) applications. The development of power electronic motor controllers has, however, brought on the market new devices that are capable of driving all rotating field machines even in dynamically demanding applications. This development has brought new motor types into focus. The synchronous reluctance motor (SynRM) and its properties in possible industrial drives are investigated in this thesis.

1.1 Current AC drives and design of the motors for VSD

A variable speed drive consists of an electric motor and a frequency converter (FC) controlling the torque and the speed of the drive. Electric motors are by far the most important type of electric load because they are used in all sectors and in a wide range of applications. Electric motors use over half of all electricity consumed in the developed countries; typically motors consume 60% ... 80% of the electricity in the industrial sector and about 35% of the electricity in the tertiary sector. The largest part is consumed by traditional synchronous and induction motors but also new motor types are emerging in the industry. For example, permanent magnet synchronous motors and synchronous reluctance motors may improve both drive performance and efficiency. The family of the electromagnetic rotating machines is shown in Fig. 1.1.

![Figure 1.1. The family of electromagnetic machines used in industrial applications.](image)

Typical configurations of controlled AC drives given in (Leonhard 1995) are shown in Table 1.1. Some typical industrial applications are taken into account. The updated table gives a basis for discussion about different AC drive applications. The table considers the influence of the development of IGBTs and new IGCTs (Steimer 1999).

Modern converters offer for different motor types control properties that are superior to those of the conventional DC drives. This consequently brings up the question of the right motor type for certain applications. Such machine types as permanent magnet synchronous motors (PMSMs), synchronous reluctance motors (SynRMs) or permanent magnet assisted synchronous reluctance motors (PMASynRMs) drives, which is a combination of the first two types, may also be used extensively in the future.
Table 1.1. Typical configurations of controlled AC-drives (Leonhard 1995 updated)

<table>
<thead>
<tr>
<th>Machines</th>
<th>Converters</th>
<th>DC link converters</th>
<th>Cycloconverters with line commutation (GTO)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Voltage-source converters (VSI)</td>
<td>Current-source converters (CSI)</td>
</tr>
<tr>
<td>Synchronous motor</td>
<td>Transistor inverter (IGBT)</td>
<td>Thyristor inverter (GTO, IGCT or GCT)</td>
<td>Thyristor inverter (GTO)</td>
</tr>
<tr>
<td>with permanent magnet excitation</td>
<td>Low power (10 kW), very good dynamic performance (servo drives), 3 MW windmill position sensorless direct drives</td>
<td>Medium power (3 MW), high power density</td>
<td></td>
</tr>
<tr>
<td>Synchronous reluctance motor</td>
<td>Low to medium power (100 kW)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Squirrel-cage induction motor</td>
<td>Low to medium power (500 kW), high speed, very good dynamic performance (spindle and servo drives)</td>
<td>Medium to high power (5 MW), high dynamic performance (traction drives)</td>
<td></td>
</tr>
<tr>
<td>Synchronous motor</td>
<td>Shaft generators on ships (2 MW)</td>
<td>High power (20 MW), subsynchronous operation</td>
<td>High power (100 MW), limited speed control range</td>
</tr>
<tr>
<td>with field and damper windings</td>
<td>30 MW ACS6000SD (ABB)</td>
<td>High power 100 MW (ABB)</td>
<td>High power (20-40 MW), low speed, good dynamic performance</td>
</tr>
</tbody>
</table>

1.2 The present-day Induction Motor

The most common electric motor type for industrial use is the AC-three phase, low voltage, 4-pole, continuous duty, totally enclosed (IP55), fan cooled, asynchronous squirrel cage motor. Generally, it is estimated that these motors represent about 90% of the total electrical energy consumption by electric motors (Frost 1994). Induction motors have gained this prevailing position because of their low price, simple network connection or inverter drive, good availability, simple construction and high reliability. During the latest century the IM factories have driven up and extremely improved their productivity. This, of course, forms a major obstacle to the adopting of new motor types for production. Some superior properties are needed before any changes can be done. The use of permanent magnets in low speed industrial machines seems to supply this demand. The significance of the SynRM remains more ambivalent and doubtful. The advantages of the SynRM compared to the induction motor are the synchronous rotation speed, possibility of sensorless control and a possibility to achieve a better or at least as good efficiency as achieved with induction motors.
In the latest decades the induction motor has been developed to produce higher and higher efficiencies in direct network drives. The new material properties give the designers new opportunities and modern calculation tools like Finite Element Method (FEM) combined with circuit simulators may improve the design remarkably. This is extremely valuable for the designing of induction motors for variable speed drives. The use of 2D or 3D FEM not only makes it possible to produce the necessary initial machine parameters for the drive control but it also enables the designer to study the transient behaviour of the machine as e.g. the torque ripple. An example of this development is shown in Fig. 1.2. The Figure describes the improvement of the efficiency of the standard ABB motor series and of one high performance motor series. The oldest one of the motor series, MKB, manufactured by ASEA, came on the market in 1935. The newest one, M3000, has been in production since 2001.

![Figure 1.2. Improvement of the efficiency of ABB’s 4-pole motors according to IEC 34-2 (Sjöberg 1997). The limiting curve for the EC/CEMEP agreement of the highest efficiency class level EFF1 is also given (EC-CEMEP 1999).](image)

The three-phase squirrel cage induction motor consumes approximately about 34 % (700 TWh) of the total electricity in the European Union. About 80 % of these motors are small power range (≤ 15 kW) 4-pole standard motors. It has been evaluated that using high efficiency EC/CEMEP EFF1 class motors and FCs more than 100 TWh of electricity could be saved annually in the EU (Haataja 2002a).

Despite a general improvement of the efficiency of the IM, the efficiency values of induction motors produced by different manufacturers are still varying considerably, Fig. 1.3. The dots (● Δ – etc.) in Fig. 1.3 show the efficiency values of motors manufactured by the major motor factories and available on the market. These values are taken from the EURODEEM database and are given according to the IEC 34 standard. Also two entries from ABB Motors awarded with "IEA Award of Excellence" are shown in Fig. 1.3 (Haataja 1999a).
Figure 1.3. Four pole, standard induction motor efficiencies as a function of power for different manufacturers according to IEC 34-2 (EURODEEM 2000). The EC/CEMEP efficiency class limits for four-pole motors are also represented. The two distinguished dots indicate the IEA Hi-Motors competition results.

The accuracy of the efficiency determination depends on the test method used and the precision of the loss determination by the test method. The methods used are IEC34-2, which is often preferred by the European manufacturers, and the American IEEE 112, which gives more accurate results (IEC 34-2 1972, NEMA 1993, IEEE 1991, Haataja 1999b).

International standards set also many limitations to the design. For example, the maximum allowable direct line starting current limits the improvement of the motor performance. Seldom the motor is optimised for a VSD. The motors are normally designed for sinusoidal supply. There are no motor series designed specifically for inverter supply but standard motors are usually used also in variable speed drives. Fig. 1.4 e.g. introduces a modern induction motor rotor design with double cage and closed rotor slots.

Figure 1.4. The cross section area of a squirrel cage rotor for a machine with number of rotor bars $Q_r = 36$. 
This rotor type is not very well suited for the VSD using a pulse width modulated high switching frequency inverter. The double cage is introduced in order to limit the starting current in direct network start. Regarding the VSD this is not necessary. It is obvious that for variable speed drives especially different rotor types for IMs should be designed. When a standard squirrel cage motor is selected for a converter drive its continuous output power must – because of the increased losses – be derated in nominal operating conditions. With the lower output the increasing of the motor temperature is maintained at the originally designed level despite of the fact that the converter supply causes additional losses in the motor (Nee 1995, Haring 1999, Abrahamsen 2001). This deration is dependent on several factors in the total drive system, converter type and modulation, motor design, mechanical power transmission and control equipment. For example, in the case of an IM the direct line-starting current has no meaning at all and the measures to limit the starting current make the motor less efficient for the VSD. The rotor bars of the IM may thus be designed with a larger cross section area and thus reduce the rotor ohmic loss. In such a case it might be possible to select the rotor diameter a little bit smaller. By reducing the rotor diameter the depth of the stator slots may be increased and thus the stator resistance can be decreased by increasing the amount of copper in the stator slots. The reducing of the rotor outer diameter also results in lower friction losses, centrifugal force and noise.

The IM VSD has two main weaknesses: 1. Estimating the rotor actual speed, especially at low speeds, is very difficult and requires rotor speed encoders in demanding drives and 2. the using of standard induction motors optimised for line starting that do not provide the user with the optimum performance in the VSD.

1.3 Synchronous Reluctance Motor

In the following, the basics of the singly salient synchronous reluctance motors are introduced. The history of the SynRM is briefly described. Because the electromagnetic torque for the SynRM $T_{el}$ consists only of the reluctance torque, the two most important parameters are derived: the direct and the quadrature axes’ inductances $L_d$ and $L_q$. The thesis will consider only three-phase synchronous reluctance motors with stators similar to those of a standard three-phase IM.

The principle of using the reluctance differences to produce the torque has been known for over 160 years. Before the discovery of the rotating magnetic field by Tesla in 1887 this first reluctance motor was close to the doubly salient synchronous reluctance motor, which is nowadays known as the switched reluctance motor. The first rotating-magnetic-field synchronous motor was, however, introduced by Kostko not earlier than in 1923 (Kostko 1923).

In the literature there seems to be no uniform designation for singly salient synchronous reluctance motors. Frequently used names are: Reluctance motor (RM), Synchronous Reluctance Motor (SRM, Synchrel, SynRM) and Reluctance Synchronous Motor (RSM). In this thesis it is used the name and abbreviation Synchronous Reluctance Motor (SynRM).

Fig. 1.5(a) shows a cross-sectional view of a singly salient RM consisting of a non-salient stator and a two-pole salient rotor, both constructed of high-permeability magnetic material. The figure shows a three-phase stator winding although any number of phases is possible. Fig. 1.5(b) shows the cross-sectional view of a three-phase doubly salient RM.
In principle, the SynRM is similar to the traditional salient pole synchronous motor but does not have an excitation winding in the rotor. Before the development of today’s AC motor the variable speed drive technology motor was excited from a fixed frequency supply. In such a case, it is necessary that the SynRM includes a squirrel cage on the rotor to provide the starting torque. Otherwise, the rotor does not accelerate and synchronize with the supplying network. The squirrel cage is also needed as a damper winding in order to maintain synchronism under pulsating load torques. The inclusion of a cage for line starting may impose conflicting requirements on the rotor design.

Regarding the primitive models of these motors types the designing of a motor appeared to be a compromise between the needs of the cage winding and the need for a large saliency ratio. The line-start type “synchronous induction motors” have simple constructions, but the saliency ratio is too low to give a competitive performance compared to the IM. These early primitive models of SynRM used to have a low saliency ratio that was often achieved by removing some teeth from the conventional induction motor rotors. They were applied only for relatively small output power. The line-start type synchronous reluctance motors were developed particularly in the 1960’s (Lawrenson 1963, 1964, 1967, Lipo 1967). These early motor types are illustrated in Fig. 1.6. The design of these motor types is based on the induction motor of which the SynRM is a modification.

The basis for the design of the rotor is the IM rotor with a die-cast cage and a single layer flux barrier. However, this is a way of approaching that should be given up. In order to achieve a good performance for the SynRM the designing of the rotor should be reconsidered.
Regarding the VSD the SynRM belongs to the category of brushless synchronous AC machines, which are sinusoidal current driven machines and use a standard quasi-sinusoidal distributed AC stator winding and inverter. The other main types of brushless synchronous AC machines are the permanent magnet synchronous motors (PMSM). The synchronous motor operates, as the names suggest, at a synchronous speed determined by the stator excitation frequency and the number of poles. The different types of brushless synchronous AC machines are shown in Fig. 1.7. The dotted areas represent steel, the white areas represent air and the dark grey areas represent permanent magnets.

Because the permeability of the present-day permanent magnet materials is close to air, the IPM motor produces also some reluctance torque because the inductances of the direct or d- and quadrature or q- axes are not equal. Because the magnetizing inductance value is inversely proportional to the square of the pole-pair number, the reluctance torque may be significant only when the number of pole-pairs is small (1 ... 3). The magnetic asymmetry of the rotor determines the different axes in a SynRM. In this thesis, the d-axis is the direction, where the value of the magnetic circuit reluctance achieves the minimum and the q-axis is the direction, where the corresponding reluctance achieves the maximum. The d-axis reluctance consists mainly of the air-gap reluctance but the q-axis reluctance includes with this air-gap reluctance also the reluctance of the flux barriers.
The line-start SynRM can also be designed based on the SynRM construction where a cage or damper winding is placed on the flux barrier holes. The line-start SynRM rotor may thus be based also on the construction of a multiple barrier rotor with damper winding, Fig. 1.8. In this case, the rotor is primarily designed for a high performance SynRM and the damper windings have been added afterwards to the rotor construction.

![Figure 1.8](image_url). The multi-barrier SynRM construction with damper winding.

In frequency converter applications the damper cage fulfils the task of dampening the oscillations in speed and increasing the torque change speed. The cage may also improve the mechanical strength of the rotor. In converter-fed synchronous machines the time harmonic contents of the stator currents, the stator magneto-motive force spatial harmonic contents and the air-gap region permeance variations induce currents in the damper.

Today’s variable speed industrial drives are mostly based on standard one- or two-pole pair induction motors. These pole pair numbers are also suitable for the SynRM. The non-domestic applications for the SynRM in are
- industrial drives,
- electrical road vehicles, rail-road traction, propulsion and auxiliaries.

The advantages of the SynRM VSD compared to IM drives generally mentioned are (Miller 1991):
- simple rotor construction with no vital need for the rotor cage in speed controlled drives
- no rotor ohmic losses
- low inertia
- synchronous running with easy speed control without encoders
- easy field weakening compared to the PMSM
Generally, SynRM rotor losses are ignored because the rotor does not necessarily consist of a rotor cage. On the other hand, to achieve a high magnetising inductance the air-gap length of the SynRM must be small if a stator is used equal to the IM stator. The air-gap length may thus be smaller than that of the respective IM and this yields to iron losses in the rotor surface that are bigger than those of the IM. A sophisticated motor design and especially the use of stator semi-magnetic slot wedges diminishing the air-gap permeance harmonic losses on the rotor surface may increase the efficiency of the SynRM even to be higher than that of the IM. The SynRM has another advantage over the IM: it offers the possibility of producing high quality sensorless drives because of its synchronous running and large inductance ratio \( L_{sd} > L_{sq} \), which facilitates an accurate estimation of the rotor position. Considering the control accuracy in a sensorless control the SynRM is superior to the IM, especially at zero speed and at very low rotational speeds.

According to Table 1.1 the SyRM is suitable only for low or medium power applications. Traditionally, this motor type has been used in conveyor applications where line start synchronous reluctance motors are advantageous because of their synchronous steady state running. The other, poor properties of the models used on the market have just to be accepted in these applications.

According to the number of the papers published about SynRM in the latest decades, there seem to be an increasing interest towards the SynRM due the development of the modern AC motor drive technology. Research and development in the 90’s in the field of synchronous reluctance motors and drives have been focused on:

- design and optimisation of transversally or axially laminated rotor structures:
  Until 1990 research on the SynRM was concentrated almost completely on the use of ALA type rotors since they can produce a very high saliency ratio. Mainly because of the iron loss (two to three times that of the multiple-barrier rotors) and fabricating problems related to the ALA type rotors the focus of research is now moved to the use of TLA rotors (Vagati 1992).

- comparative analysis of synchronous reluctance motor and induction, permanent magnet, or switched reluctance motors:
  At low speed the torque of the IM is limited and the control algorithms is complex. On the other hand, at low speed the losses in the SynRM might be low enough to relieve the need for an external fan. For certain applications (e.g high temperature applications) the absence of magnets may be an advantage. Many qualities that are claimed to be characteristic for the switched reluctance motor are also offered by the the SynRM, but the SynRM uses a standard AC IM stator. Like the IM, it is robust and brushless, and since it operates with a rotating magnetic field an inherently smooth and quiet operation is achieved (Staton 1993b).

- vector control of SynRMs:
  The study of the control of SynRM has been focused on the torque vector control (Boldea 1991), including the effects of saturation and iron losses (Xu 1991) on the field oriented control (Matsuo 1994) and on the field weakening (Fratta 1994). The SynRM may thus be operated according to all known vector control methods

- five phase synchronous reluctance motor:
  By increasing the number of phases (>3) of the SynRM some advantages can be noticed: a reduced current per phase for higher reliability, a reduced common-mode voltage for less EMI, and an increasing of the redundancy in the machine winding and associated converter device for fault tolerance (Longya 1999).
1.4 Vector presentations of the IM and SynRM

If the SynRM carries a damper winding the rotor-co-ordinate equations are similar to those of the IM. The stator voltage equation of all rotating field machines in vector form can be given as

\[ u_s^r = R_s i_s^r + \frac{d \psi_s^r}{dt}, \quad (1.1) \]

which gives the stator flux linkage as a voltage integral

\[ \psi_s^r = \int (u_s^r - R_s i_s^r) dt, \quad (1.2) \]

where \( u_s^r, i_s^r, \) and \( \psi_s^r \) are the instantaneous stator voltage, current and flux linkage vectors in the stationary reference frame.

By multiplying the stator voltage vector by \( e^{-j\omega t} \), the stator voltage may be presented in a rotor reference frame as

\[ u_s^r = u_s^r e^{-j\omega t} = R_s i_s^r + \frac{d \psi_s^r}{dt} + j \omega \psi_s^r. \quad (1.3) \]

The product \( \omega \psi_s^r \) is defined by the angular speed of the rotor reference frame. In Eq. (1.3) \( \omega_r \) is the rotor electrical rotational speed.

In a general reference frame the co-ordinate axes rotate at a synchronous speed \( \omega_s \), and the Eq. (1.3) is written as

\[ u_s^g = R_s i_s^g + \frac{d \psi_s^g}{dt} + j \omega_s \psi_s^g. \quad (1.4) \]

The rotor equation in the general reference frame is then

\[ u_t^g = R_t i_t^g + \frac{d \psi_t^g}{dt} + j (\omega_s - \omega_r) \psi_t^g. \quad (1.5) \]

Ignoring iron loss components and using the voltage Eqs. (1.4) and (1.5) a common equivalent circuit for an IM and a SynRM may be drawn.
Figure 1.9. The IM and SynRM equivalent circuit in a rotating ($\omega_g$) general reference frame.

In the stator reference frame $\omega_g = 0$ and in the rotor co-ordinates $\omega_g = \omega_r$ and thus Eqs. (1.4) and (1.5) are simplified respectively. Since the magnetising inductance of the SynRM is dependent on the rotor position the rotor dq-co-ordinates may conveniently be used in the evaluation.

Using the two-axis theorem in the rotor dq-co-ordinates the Eq. (1.3) can be resolved also in direct and quadrature axis components as

$$u_{sd} = R_s i_{sd} + \frac{d\psi_{sd}}{dt} + \omega_r \psi_{sq}, \quad (1.6)$$

$$u_{sq} = R_s i_{sq} + \frac{d\psi_{sq}}{dt} - \omega_r \psi_{sd}. \quad (1.7)$$

The stator flux linkages in stator co-ordinates in an IM and in a SynRM with damper windings consist of the stator leakage and the air-gap fluxes as

$$\psi_s^g = L_{sg} i_s^g + L_{sm} (i_s^g + i_r^g), \quad (1.8)$$

where $L_{m}$ is the magnetising inductance tensor representing the different axes’ inductances. $L_{sg}$ is the stator leakage inductance tensor and $i_r^g$ is the rotor current vector (in the case of a SynRM $i_r^g = i_D^g$, which is the damper winding current).

In an IM we may abandon the tensor representation since the magnetising and leakage inductances of the IM are identical in the d- and q-axes directions

$$L_{md} = L_{mq}, \quad (1.9)$$

but for the SynRM the magnetising inductance (as well as the stator leakage inductance) is different in the d- and q-axes

$$L_{md} \gg L_{mq}. \quad (1.10)$$

The tensor representation may be left out in this case by using the traditional d-q-axis representation for the machine.
The stator leakage inductance is generally assumed to be independent of the rotor position and to be identical in the direct and quadrature axis direction. If the rotor structure contains cut-outs on the q-axis, Fig. 1.7b, the leakage inductance values of the axes are no longer, in principle, equal because the leakage inductance value is also a function of the air-gap length so

\[ L_{sq} \geq L_{sd}. \]  

(1.11)

The two equivalent circuits of the axes of the IM and SynRM in a synchronously rotating rotor frame (\( \omega_b = \omega_s \)) are shown in Fig. 1.10.

The vector diagrams of the IM in stationary reference frame fixed to stator and the SynRM in rotor oriented reference frame at steady state are shown in Fig. 1.11.
In the following the quantities of an AC system may be expressed in a dimensionless form, in so-called per unit (p.u.) values. This enables easy comparison of motors of different dimensions. The base values marked using subscript b are defined as

\[ I_b = \sqrt{2} I_N, \]  

\[ U_b = \sqrt{2} U_{\text{phase}} = \sqrt{2} \frac{U_N}{\sqrt{3}} \]  

\[ Z_b = \frac{U_b}{I_b}, \]  

\[ T_b = \frac{P_N}{\omega_N}, \]  

\[ t_b = \frac{T_N}{T_b}. \]  

The different parts of the impedance can then be expressed in per unit values as
The stator current armature reaction in an IM is to a large degree eliminated by the magnetomotive force of the rotor current. The small sum of the stator and the rotor current vectors is the air gap flux linkage magnetising current $i_m$. In the SynRM there are no armature reaction eliminating currents in the rotor. This clearly indicates that it is more difficult to produce a high power factor with the SynRM. The larger the d-axis inductance is the smaller the no load current of the SynRM remains. And when the q-axis inductance becomes smaller the q-axis current will have less influence on the air gap flux and the power factor of the machine will be better. The $L_d/L_q$ ratio is thus an extremely important factor determining the performance of the SynRM.

The electro-magnetic torque of all rotating field machines may be given as a vector presentation

$$T_e = \frac{3}{2} \rho (\psi_s \times i_s).$$  \hspace{1cm} (1.17)

The vector presentation, however, is not always very useful and thus the scalar version for the torque may be written using different reference frames as

$$T_e = \frac{3}{2} \rho (\psi_{sd} i_{sq} - \psi_{sq} i_{sd}).$$ \hspace{1cm} (1.18)
The torque of the IM may be calculated most conveniently in the stator co-ordinates using the torque

\[ T_e = \frac{3}{2} \rho (\psi_{q} i_{dx} - \psi_{sd} i_{sy}), \]  

(1.19)

where the x-axis currents \( i_{sx} \) is the direct axis current component and \( i_{sy} \) is the quadrature axis current component in the stator reference frame. The \( x,y \)-axes are fixed to the stator.

For a SynRM the stator flux linkage may be given as

\[ \psi_s = L_d i_{sd} + jL_q i_{sq} = \psi_{sd} + j\psi_{sq}. \]  

(1.20)

The inductances \( L_d \) and \( L_q \) are variables and are dependent on self and cross saturation of the axes. Cross saturation is the magnetic coupling between the fictitious d- and q-axis windings of the machine. These aspects will be studied in chapter 2.

Fig 1.13 gives for the stator current vector \( i_s \) and flux linkage \( \psi_s \) components

\[ \begin{align*}
  \psi_{sd} &= L_d k_s \cos \kappa, \\
  \psi_{sq} &= L_q k_s \sin \kappa, \\
  i_{sd} &= k_s |i_s| \cos \kappa, \\
  i_{sq} &= k_s |i_s| \sin \kappa.
\end{align*} \]  

(1.21)

Using the equivalence \( \sin \alpha \cos \beta = \frac{1}{2} \left[ \sin(\alpha - \beta) + \sin(\alpha + \beta) \right] \), the torque Eq. (1.17) gives

\[ T_e = \frac{3}{4} \rho (L_d - L_q) k_s^2 \sin 2\kappa. \]  

(1.22)

The inductance difference between the d- and q-axis inductances is not constant. Due to self and cross saturation it varies with both the d- and q-axis currents of the machine.

The maximum theoretical torque in a SynRM with a known current is thus found when the current-d-axis angle is \( \pi/4 \).
The SynRM maximum torque producing load angle $\delta_{\text{max}}$ can be solved with a constant voltage supply, if the magnetising current $i_m$ and the stator flux linkage modulus $|\Psi_s|$ are given. The stator current modulus is assumed to be unlimited here. The stator flux linkage components and the stator current components in the rotor reference frame can be expressed as a function of the stator flux linkage load angle $\delta$.

$$\Psi_{sd} = |\Psi_s| \cos(\delta),$$  \hspace{1cm} (1.23)

$$\Psi_{sq} = |\Psi_s| \sin(\delta),$$  \hspace{1cm} (1.24)

$$i_{sd} = \frac{1}{L_d} |\Psi_s| \cos(\delta),$$  \hspace{1cm} (1.25)

$$i_{sq} = \frac{1}{L_q} |\Psi_s| \sin(\delta).$$  \hspace{1cm} (1.26)

The torque can be expressed as a function of the load angle:

$$|\tau| = \frac{3}{2} p |\Psi_s| \times i_s = \frac{3}{2} p \left( |\Psi_s| \cos(\delta) - \frac{1}{L_d} |\Psi_s| \sin(\delta) - |\Psi_s| \sin(\delta) - \frac{1}{L_d} |\Psi_s| \cos(\delta) \right),$$  \hspace{1cm} (1.27)

$$= \frac{3}{2} p |\Psi_s|^2 \left( 1 - \frac{1}{L_d} \right) \sin(\delta) \cos(\delta),$$  \hspace{1cm} (1.28)

$$= \frac{3}{2} p |\Psi_s|^2 \left( \frac{L_d - L_q}{L_q L_d} \right) \frac{1}{2} \sin(2\delta),$$  \hspace{1cm} (1.29)

which corresponds to the reluctance power part of the salient pole synchronous machine power. The more traditional expression is the expression for p.u. power $p_{\text{out}}$

$$p_{\text{out}} = \omega |\Psi_s| = \frac{3}{2} p |\Psi_s|^2 \left( \frac{L_d - L_q}{2 L_d L_q \omega} \right) \sin(2\delta),$$  \hspace{1cm} (1.30)

where $|\Psi_s| = \frac{\omega |\Psi_s|}{U_b}$.  \hspace{1cm} (1.31)

If the inductances are assumed to be independent from the load angle the maximum torque producing load angle can be found by differentiating Eq. (1.28) with respect to the load angle and finding zero for the derivative
\[
\frac{\partial T_s}{\partial \delta} = \frac{3}{2} p \left( \sin(\delta) \cdot -\left| \psi_s \right|^2 \left( \frac{1}{L_q} - \frac{1}{L_d} \right) \sin(\delta) + \cos(\delta) \left( \left| \psi_s \right|^2 \left( \frac{1}{L_q} - \frac{1}{L_d} \right) \cos(\delta) \right) \right)
\] (1.32)

Setting (1.32) to zero yields
\[
2 \cdot \left| \psi_s \right|^2 \left( \frac{1}{L_q} - \frac{1}{L_d} \right) \cos^2(\delta) - \left| \psi_s \right|^2 \left( \frac{1}{L_q} - \frac{1}{L_d} \right) = 0.
\] (1.33)

The solution is
\[
\cos(\delta) = \sqrt{\frac{1}{2}}.
\] (1.34)

This gives the maximum torque producing power angle \( \delta = \pi/4 \).

The torque in the per unit value as a function of the load-angle and current angle is given in Figs. 1.14a and 1.14b. The stator flux linkage used in the calculation presented is based on a 230 V grid voltage 50 Hz frequency. The parameters used in the calculations are listed in Table 1.2.

<p>| Table 1.2. The values of the calculations |</p>
<table>
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<th>Parameter</th>
<th>Value</th>
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</tr>
<tr>
<td>Current</td>
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<tr>
<td>Motor speed</td>
<td>1500 rpm</td>
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<td>Pole number</td>
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<tr>
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<tr>
<td>Stator leakage inductance</td>
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<tr>
<td>Iron + mechanical losses</td>
<td>890 W</td>
</tr>
<tr>
<td>Quadrature-axis inductance</td>
<td>7 mH</td>
</tr>
<tr>
<td>( L_q )</td>
<td>0.2 p.u.</td>
</tr>
</tbody>
</table>

In the torque calculations the self and mutual saturation of the inductances as well as the effect of the iron losses are ignored.
Figure 1.14. The calculated reluctance torque as a function of a) the load-angle and b) the current angle. The saliency ratio $L_d/L_q$ is a parameter.

The calculated torque values do not take into account saturation or iron loss effects and therefore the given torque values are too optimistic. In a constant voltage supply the stator current angle with respect to the rotor d-axis producing the maximum torque is large and differs significantly from the current supply theoretical value $\kappa = \pi/4$ as calculated in Eq. (1.22).

If the effect of the stator resistance on the power factor value is ignored, the power factor $\cos \varphi$ can be estimated from Fig. 1.11b as
\[ \varphi \approx \delta + \frac{\pi}{2} - \kappa. \]  

(1.35)

At the maximum SynRM torque load-angle the power factor angle reaches the value of

\[ \varphi_{t,\text{max}} \approx \delta_{t,\text{max}} + \frac{\pi}{2} - \kappa = \frac{\pi}{4} + \frac{\pi}{2} - \kappa = \frac{3\pi}{4} - \kappa. \]  

(1.36)

At the motor operation area the current angle is in the range of

\[ \kappa \in \left[0, \frac{\pi}{2}\right], \]  

(1.37)

At the maximum torque the SynRM power factor is thus less than

\[ \delta_{t,\text{max}} < \cos\left(\frac{\pi}{4}\right) = 0.707. \]  

(1.38)

If the stator resistance and the iron losses are ignored and the inductances are assumed to be not depending of the load angle, the maximum power factor \( \varphi_{t,\text{max}} \) can be found from the equations as

\[ \cos\varphi_{t,\text{max}} = \frac{T_e \omega}{mpu i_s^2}, \]  

(1.39)

Using Eqs (1.17), (1.31) and (1.39) the power factor is

\[ \cos \varphi = \frac{(L_d - L_q) i_{ad} i_{sq}}{\sqrt{(L_d i_{ad})^2 + (L_q i_{sq})^2} \sqrt{i_{ad}^2 + i_{sq}^2}}. \]  

(1.40)

The maximum power factor producing current angle can be found by differentiating Eq. (1.40) with respect to the current axes ratio \( i_{sd}/i_{sq} \) and finding zero for the derivative

\[ i_{sq} = \frac{i_{sd}}{L_q}. \]  

(1.41)

According to Eq. (1.40) the power factor achieves its maximum when the current angle is

\[ \kappa = \arctan \frac{L_d}{L_q}. \]  

(1.42)

and the maximum power factor is

\[ \cos \varphi_{\text{max}} = \frac{L_d - L_q}{L_d + L_q}. \]  

(1.43)
Fig. 1.15a and 1.15b show the respective power factor values as a function of the load-angle and current vector angle. The saliency ratio is a parameter.

Figure 1.15. The calculated power factor as a function of a) the load-angle and b) current angle. The saliency ratio $L_d/L_q$ in the calculations is a parameter. The leakage inductance $L_{mσ}$ has been ignored.

From Figs. 1.14 and 1.15a it can be noticed that the maximum torque and maximum power factor values are achieved at different load angles. The Figs. 1.15a and 1.15b show also that the
maximum power factor values are achieved at lower load angles and higher current angles as the saliency ratio increase.

The 30 kW four-pole IM per unit static torque in direct network starting as a function of per unit slip is illustrated in Fig. 1.16.

![Figure 1.16](image1.png)

Figure 1.16. The 30 kW four-pole IM torque per unit and power factor values as a function of the motor slip.

From Fig. 1.16 it can be noticed that the IM maximum torque may be achieved at a different slip than the maximum power factor.

The calculated power factors for the 30 kW four-pole IM and SynRM as a function of the motor output power are illustrated in Fig. 1.17. The inductance parameters in the calculation of the SynRM are set so that the quadrature axis inductance is kept constant and the direct axis inductance limits the no-load current of the SynRM to be equal to the no load current value of the IM at saliency ratio value of 10.

![Figure 1.17](image2.png)

Figure 1.17. Calculated power factors as a function of the motor partial load for a) the IM and b) SynRM. For the SynRM the saliency ratio \(L_d/L_q\) in the calculations is a parameter. The leakage inductances \(L_{\sigma}\) have been ignored. The highest power factor values are found using only very small air-gap lengths – smaller than the IM air-gap length.
From Fig. 1.17 it can be noticed that a SynRM with a high saliency ratio achieves a good power factor also at lower partial loads. To achieve for the SynRM a nominal point power factor equal to that of the IM the d-axis magnetising inductance of the SynRM must be higher than the magnetising inductance of the IM. This aspect will be discussed more in detail in chapters 2 and 3.

The vector diagram in Fig. 1.18 illustrates the SynRM power factor value when the saliency ratio is \( L_d/L_q = 5 \), the load angle is \( \delta = 37.5^\circ \) and the current angle is \( \kappa = 75.4^\circ \). These values have been marked with small circles in Figs. 1.16a and 1.16b.

The SynRMs efficiency at different saliency ratios is estimated as

\[
\eta = \frac{P_{Out}}{P_{Out} + P_{Loss}}, \quad (1.44)
\]

\[
\eta = \frac{T_e \omega}{T_e \omega + m I_s^2 R_s + P_{Fe+Mech}}, \quad (1.45)
\]

\[
\eta = \frac{m U_s^2 (L_d - L_q) I_d I_q \sin(2\delta)}{m U_s^2 (L_d - L_q) I_d I_q \sin(2\delta) + m I_s^2 R_s + P_{Fe+Mech}}, \quad (1.46)
\]

To simplify the evaluation in this case the loss component \( P_{Fe+Mech} \) consists of the no-load iron losses and the mechanical losses and the effects of the stray load losses are ignored. The load dependent iron losses are ignored. The calculated results as a function of the load angle and current angle are shown in Fig. 1.19. The values used in the calculations are given in Table 1.2. Because the quadrature axis inductance is quite independent of the air-gap length, this axis inductance is chosen for the initial parameter where the d-axis inductance is calculated using the saliency ratio.
Figure 1.19. The calculated efficiency as a function of a) the load-angle and b) current angle. The saliency ratio $L_d/L_q$ in the calculations is a parameter.

The calculated efficiency values are relative optimistic. This may be explained by the fact that, firstly, the torque values were calculated using current values which do not take into account the effect of iron losses and, secondly, the inductance values used ignore saturation.
1.5 SynRM rotor constructions

When designing a rotor for a high performance SynRM it should be found the largest possible inductance ratio to guarantee the best power factor and the best efficiency for the machine. No primitive solutions can thus be allowed and the solution should most probably consist of multi flux barrier features. Fig. 1.20a. introduces an approach suggesting a design which produces a quite small q-axis inductance because of the large air gap. But, at the same time the d-axis inductance remains small and thus the inductance ratio remains in the rang of 2 … 3 which does not produce good enough properties for the motor. This rotor type does not bring the satisfactory performance results and can not compete e.g. with the performance of the IM.

The synchronous reluctance motor needs a high saliency ratio to be competitive with the induction motor. To achieve a high saliency ratio there are, in principle, two different possible ways of manufacturing the synchronous reluctance motor rotor: with transversal laminations (TLA) Fig. 1.20(b) or with axial laminations (ALA) Fig. 1.20(c).

Transversely laminated rotors (TLA) are made of standard flat laminations, where the flux barriers are punched in order to obtain the preferred flux paths. In the ALA type rotor laminations are interleaved with nonmagnetic material. The ALA type rotor has been introduced in the literature since 1923 (Kostko 1923, Douglas 1956, Cruickshank 1966, El-Antably 1985). The TLA type rotor is described e.g. in (Kurscheidt 1961, Brickman 1965).

The ribs of the TLA type rotor fundamentally reduce the q-axis reluctance and so this rotor type has inherently a lower saliency ratio than the same shape ALA-type motor where no ribs are needed. Magnetic tangential webs have the advantage that they make the rotor surface smooth and thus decrease the volume of the air-gap flux harmonics. The flux barriers reduce the q-axis flux and increase thus the saliency ratio. The TLA rotor type is probably the only model that can be manufactured at moderate costs and for this reason factories producing squirrel cage IMs are concentrating their interest on this type. Regarding the rotor surface losses caused by the air gap spatial harmonics the thin transversal laminations also are beneficial. However, with this rotor type it may not be possible to reach the largest possible inductance ratios.
The ALA-type rotor structure makes it possible to use numerous flux barriers and to leave the tangential and inner ribs out and thus to achieve the highest possible inductance ratios. In the ALA type rotor numerous flux barriers are used to decrease the air gap flux spatial harmonic contents and to reduce the torque ripple and iron losses. Stator slot openings create air gap flux density variations that readily induce the eddy currents in the ALA-structure. Thus, the laminations must be thin and highly resistive. However, the ALA-rotor is more difficult and probably much more expensive to manufacture than a squirrel cage rotor.

Bomela (1999) suggests the skewing of the SynRM rotor by one slot pitch to reduce the torque ripple drastically (from more than 20 % to less than 5 %). This skewing has only a little effect on the average torque. The sheets of the TLA-type rotor are much easier to skew than the interleaved sheets of the ALA-type rotor. The torque ripple can also be reduced by chording the stator windings. Based on (Bomela 1999) chording has little effect on the torque ripple of the SynRM while it significantly reduces the average torque. On the other hand, this analysis considers only rotors with a number of flux barriers that is less than the number of the stator slots. The number of flux barriers of the ALA-type rotor is not that limited as that of the TLA-type rotor.

For reasons of manufacturing the TLA type motor appears to be the better alternative for the induction motor than the ALA type motor. The TLA type motor can be built at low costs and its rotor can be skewed to decrease the torque ripple.

In recent years a number of studies has been done to compare the performance of induction motors and synchronous reluctance motors. The studies are mainly focusing on the motor efficiency and power factor. The feasible SynRM rotor types that are comparable with the IM are the transversally laminated (TLA) and the axially laminated (ALA) rotors. Due to its ribs the TLA-rotor has a lower saliency ratio than the ALA-type. However, the TLA-motor appears to be the better alternative for the IM since it can be manufactured using the same kind of punching tools as used for the manufacturing of the rotor laminations of an IM. The TLA skewed rotor needs a relatively small air gap to improve the inductance ratio. Table 1.3 gives some extracts from the IM vs. SynRM comparisons.

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<th>Pole Number</th>
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<th>Power factor</th>
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</tbody>
</table>
Generally, the studied SynRMs are based on the frame of a small or medium power four-pole standard induction motor and both the TLA and ALA type rotor are used. In the reference papers the SynRMs are driven by current controlled converters with a position sensor. In a considerable number of the papers the efficiency of the SynRM is considered to be better and the power factor lower than those of the competing IM. The power factor values of the SynRMs characterize the saliency ratio of the motor. Staton (1993b) has fed the IM with a sinusoidal supply at 11 kW power, but when supplied from an inverter the motor power has been derated to 6 kW to provide the variable speed 10:1 speed ratio with a constant torque.

Despite of the values given in the papers (Table 1.3) it is not at all very straightforward to reach any conclusions concerning the performance superiority between the SynRM and the IM. The performance of the motors are very much dependent on the control method used in the drive. The results given for the SynRMs are reached using some current control. For the SynRMs shown in Table 1.3, the following control strategies were used [Betz 1993]:

1) Maximum Torque Control (MTC), where maximum torque/ampere is obtained by setting the load angle $\delta = \pi/4$.

2) Maximum Rate of Change of Torque Control (MRCTC), where the load angle is set to be $\delta = \tan^{-1} \zeta$.

3) Maximum Power Factor Control (MPFC), where the load angle is set to be $\delta = \tan^{-1} \zeta^{0.5}$.

4) Constant Current in Inductive Axis Control (CCIAC), where the constant current is maintained in the d-axis and the torque is manipulated by the q-axis current.

The control strategies used for the IM are not mentioned in the papers, which makes the comparison even more difficult.

In most of the papers the power factor of the SynRM is lower compared to the power factor of the IM. On the other hand, the major part of the papers come to the conclusion that the efficiency of the SynRM is better than that of the IM. The articles, however, are quite old and for this reason the efficiencies of the competing induction motors seem not to be very high. The efficiencies of the induction motors have enhanced during the latest years and thus the competition in efficiency has grown even tougher. The efficiency of a high efficiency IM is difficult to reach with a SynRM.
1.6 Outline of the thesis

The main goal of this thesis is to present a theoretical and practical SynRM vs. IM comparison and to design a synchronous reluctance motor for an industrial frequency converter drive within the constraints given by the initiator of the work. One of the main constraints is that the stator of the original induction motor must be utilized for the designing of the SynRM. Another obvious constraint is to study more specifically the performance of the TLA-rotor type because it offers some important manufacturing benefits compared to the ALA-rotor.

The research work of this thesis is structured as follows:

- studying of methods to design and analyse the performance of the synchronous reluctance motor.
- comparing the SynRM and induction motor performances.
- designing a technically competitive synchronous reluctance motor for variable speed application.

The thesis is divided into six chapters. The fundaments, the goals and the motivation for this work are given in chapter 1. Chapter 2 introduces a background theory of the iron loss mechanism and the behaviour of the dq-axes inductances in reluctance machines. Chapter 3 considers the possibilities of replacing four-pole induction motors by synchronous reluctance motors in the power range from 1.1 kW up to 90 kW in variable speed drives. Chapter 4 discusses the synchronous reluctance performance analysis using the finite element method and gives a comparison between the measured and the calculated values. The rotor design evolution of the SynRM using the finite element method is also given. An experimental evaluation of the synchronous reluctance motor and comparison with the induction motor torque output is given in chapter 5. Chapter 6 provides a summary of the work. Ideas for further investigation are suggested.

1.7 Scientific contributions of the thesis

This thesis shows that using the stator of the same size three-phase induction motor the efficiency of the transversally laminated SynRM is equivalent to that of a corresponding high efficiency IM in variable speed drives. However, the power factor on the SynRm is somewhat lower than the power factor of an IM. By redesigning the stator it might be possible to reach more competitive values for the power factor.

Using the FEM designed multi-layer transversally laminated rotor with damper windings it is possible to design a directly network driven motor without degrading the motor efficiency or power factor compared to the performance of the IM.

The power factor of the SynRM at rated load is smaller than that of the IM. However, at lower partial load this difference decreases and this, probably, brings that the SynRM gets a better power factor in comparison with the IM.

The big rotor inductance ratio of the SynRM allows a better estimating of the rotor position. This appears to be very advantageous for the designing of the rotor position sensor-less motor drive.

The calculation methods developed in the thesis give accurate results. Comparison between the calculated and measured values shows that the design methods are reliable.

A new application of the IEEE 112 measurement method is developed and used especially for determination of stray load losses in laboratory measurements.
2 SYNCHRONOUS RELUCTANCE MOTOR

In this chapter the effects of the self and cross saturation and iron losses on the synchronous reluctance motor performance are described. The cross saturation between the d- and q-axes means that the axes inductances may not be considered independent. A more thorough introduction to cross saturation is given in chapter 2.2.

Two main rotor types (ALA and TLA) and their properties are studied. Also the effect of the rotor position with respect to the stator slot openings is described.

The direct and the quadrature axes inductances $L_d$ and $L_q$ introduced in chapter 1 consist of the magnetising inductances $L_{md}$ and $L_{mq}$ and the leakage flux inductances $L_{sd}$ and $L_{sq}$ of the respective axes as

$$L_d = L_{md} + L_{sd},$$

$$L_q = L_{mq} + L_{sq},$$

Here, it is first assumed that there is no cross-saturation between the axes: $L_{md} = f(I_{d,0})$, $L_{mq} = f(0, I_{q})$.

If cross saturation is taken into account the inductances may be expressed as $L_{md} = f(I_{d}, I_{q})$, $L_{mq} = f(I_{d}, I_{q})$. The effect of cross saturation will be analysed in section 2.2. If the iron loss effects are taken into account the magnetising currents $i_{md}$, $i_{mq}$ are not equivalent to $i_d$, $i_q$. The effect of the iron losses will be analysed in section 2.3.

2.1 Inductance analysis of the SynRM

To estimate the magnetising inductances analytical or FEM calculations may be used. The latter method offers the possibility to solve the flux path more exactly and to take into account also the rotor position and the self and cross saturations.

The fundamental magnetising inductance $L_m$ of a saturated, uniform air-gap AC machine may be expressed as (Vogt 1996)

$$L_m = \frac{m \mu_0}{\pi \delta_s} DL \left( \frac{N}{p} \right)^2 \xi_1^2,$$

where

- $\mu_0$ - the permeability of vacuum,
- $m$ - the number of phases,
- $\delta_s$ - the effective air-gap,
- $D$ - the stator bore diameter,
- $L'$ - the stator stack effective length,
- $N$ - the number of turns in a phase,
- $p$ - the number of pole pairs,
- $\xi_1$ - the fundamental winding factor.

The effective air-gap length $\delta_s$ takes normally into account also the effect of the self-saturation in the iron parts as...
\[ \delta_e = \frac{V_{\text{tot}}}{V_\delta} \delta_e, \quad (2.4) \]

\[ \delta_s = \frac{V_{\text{sh}} + V_{\text{rb}} + 2(V_\delta + V_{\text{st}} + V_{\text{rt}})}{2V_\delta} \delta_e, \quad (2.5) \]

where

- \( V_{\text{tot}} \) - the total magnetic voltage (mv) of the machine,
- \( V_\delta \) - mv in the air-gap,
- \( V_{\text{sh}} \) - mv in the air-gap,
- \( V_{\text{rb}} \) - mv in the rotor yoke,
- \( V_{\text{st}} \) - mv in the stator tooth,
- \( V_{\text{rt}} \) - mv in the rotor tooth,
- \( \delta_e \) - the equivalent air-gap,

but it may also include the cross saturation effects and thus produce the right inductance values in a cross-saturated machine as well.

The equivalent air-gap \( \delta_e \) is calculated by multiplying the mechanical air-gap length \( \delta_m \) by the Carter factor \( k_C \) (Carter 1901).

\[ \delta_e = k_C \delta_m. \quad (2.6) \]

The Carter factor accounts for the effect that the flux density drops under the stator and rotor slots, which can be seen from Fig. 2.2b. If a rotor with no tangential ribs is used (as in ALA type rotors), also the effect of the rotor “slot” opening should be considered. A wide “cut-out” in the rotor surface affects the respective axis value in a same way. This would make the calculation of the Carter factor or the effective air-gap length more complicated because the replacement of a rotating slot opening should be considered. In the case of the TLA type motor without cut-outs, the rotor is closed and thus the rotor surface is smooth.

As it has been shown in Eq. 1.17 the torque production of the SynRM is proportional to the difference of the magnetising inductances of the axes. Thus, if the SynRM stator is kept equivalent to that of the IM the length of the active parts, the winding factor, air-gap diameter and the number of turns in the series per stator winding are the same. The main limiting parameter for the SynRM magnetising inductance, according to Eq. (2.3), is the air-gap length.

To maximise the d-axis inductance the flux guides should ideally have the same shape as the flux path shown in Fig. 2.1a. And to minimise the q-axis flux linkage the flux barriers should be perpendicular to the flux lines, as it is shown in Fig. 2.1b. The risk of self-saturation occurring on the d-axis flux path is high as the d-axis flux density in the d-axis iron parts is high. The flux barriers reduce the q-axis flux so that the risk of saturation on the q-axis flux path is remarkably lower. This effect is illustrated in Figs. 2.1c and 2.1.d. The flux solutions are calculated for the experimental SynRM introduced later in the thesis. The nominal values of the experimental machine are, \( U_N = 400 \, \text{V}, \, I_N = 66 \, \text{A}, \, I_{dN} = 10 \, \text{A}, \, I_{qN} = 65 \, \text{A}, \, D_\delta = 215 \, \text{mm}, \, \delta = 0.4 \, \text{mm}. \)
According to Fig. 2.1c, the most saturated areas on the d-axis flux-path are the stator teeth. There occurs slight saturation also on the rotor’s first three flux guides from the centre and in the stator yoke between the d-axis poles. The saturation in the innermost flux guide is lower than in the next two ones. The respective stator tooth meets the maximum flux and is the most heavily saturated. This means that to decrease the saturation on the d-axis the amount of iron in the stator should be increased. However, as in fact the amount of the stator copper should also be increased, this makes optimisation difficult. In the q-axis (Fig. 2.1d) only the rotor ribs and webs are saturated.

In no-load the fundamental current in the rotor bars of an IM is very low and thus the rotor situation corresponds to that of the SynRM. If the number and the width of IM rotor bars are equal to the SynRM flux guides and if the air-gap length is kept the same it may be assumed that also the d-axis inductances will be the same.

\[ L_{m(IM)} \approx L_{m(d(SynRM))} \]  

(2.7)

In the following, the cross saturation effect is taken into account using the so-called reluctivity matrix. To estimate the reluctivity matrix used in the cross-saturation FEM calculations first the fluxes in the d-axis direction and the q-axis direction are calculated using the appropriate current values. The reluctivities of each element are stored in the reluctivity matrices. Both matrices are then unified to one cross-saturation reluctivity matrix by selecting the maximum reluct-
tivity for each element. This, in practice, enables the use of different permeabilities in the d- and q-axes directions in the FEM-calculations. The results achieved (see Fig. 2.20) with this method seem to coincide well with the experimental torque measurement results. Thus, in the FEM calculation the d- axis inductance $L_{md}(i_d, i_q)$ can be found from

$$L_{md}(i_d, i_q) = \frac{\psi_{md}}{i_d} = \frac{3}{2} \xi_i N_{ph} \tau_p \hat{b}_{1d}(\hat{i}_d, \hat{i}_q),$$

(2.8)

where

- $\hat{i}_{md}$ - the peak value of the d- axis stator current,
- $\hat{i}_{mq}$ - the peak value of the q- axis stator current,
- $\xi_i$ - the fundamental winding factor,
- $N_{ph}$ - the number of turns in series per phase,
- $\tau_p$ - the pole pitch,
- $\hat{b}_{1d}$ - the peak value of the fundamental wave of the d-axis air-gap flux density,

The pole pitch $\tau_p$ is determined as

$$\tau_p = \frac{\pi \cdot D}{2 \cdot p},$$

(2.9)

where

- $D$ - the stator bore diameter,
- $p$ - the pole pair number.

The field plots of the solved FEM-problem and the air gap flux density distribution for the TLA-type motor are illustrated in Fig. 2.2.

![Field plots of the solved FEM-problem and the air gap flux density distribution](image)

Figure 2.2. a) The solved FEM-problem of the prototype motor and b) the air-gap normal flux density distribution (distorted curve) and the fundamental wave of the air-gap flux density (sine wave). In the figure $I_d = 10$ A, which correspond with the no-load current at the used motor model air-gap length (0.3 mm).
In a rough analytical estimation for the q-axis reluctance the equation used for the d-axis may be used by increasing the q-axis air-gap length as to contain also the sum thickness of the flux barriers and the effects of the saturated iron ribs and webs.

The analytical estimation of the q-axis inductance is difficult because the path of the q-axis flux is not that straightforward. Also the permeability of the q-axis iron is complicated to determine.

Using equation (2.3) the q-axis inductance may be written

\[
L_{\text{mq}} \approx \frac{\mu_0}{\pi} \frac{m}{\delta_s} + \sum_{n=1}^{N} D l_b \left( \frac{N}{\rho} \right)^2 \xi_1 \xi_2 \xi_3 ,
\]  

(2.10)

where \( l_b \) is the length of each flux barrier and \( n \) is the number of flux barriers. This equation does not take into account the effect of saturation in the ribs and webs or cut-outs. Therefore, the q-axis inductance value of the transversally laminated type structure using Eq. (2.10) gives too low values compared to the results given by the FEM calculations.

Taking \( L_{\text{md}} \) and \( L_{\text{mq}} \) from equations (2.10 and 2.11) gives a possibility to determine the inductance ratio. According to Soong (1993), however, the intrinsic saliency ratio \( \xi_i \), which is the maximum possible saliency ratio for a given motor geometry, may be simplified defined as

\[
\xi_i = \frac{L_m}{L_q} = \frac{r}{p \delta_s}.
\]  

(2.11)

where \( r \) is the radius of the rotor.

Similarly, the intrinsic magnetising saliency ratio \( \xi_{mi} \)

\[
\xi_{mi} = \frac{L_m}{L_{\text{mq}}} = \frac{ar}{p \delta_s}.
\]  

(2.12)

where \( a \) is the ratio of the insulation thickness \( w_{\text{ins}} \) to the sum of the insulation and lamination thickness \( (w_{\text{ins}} + w_{\text{lam}}) \). This intrinsic saliency ratio is the maximum saliency ratio theoretically achievable with a given motor geometry. Practical motors generally achieve, depending on their design, unsaturated saliency ratios that are only 20 … 40 % of the given ratio.

The q-axis inductive may thus be evaluated using the saliency ratio equations shown above and calculating the d-axis inductance as

\[
L_{\text{mq}} = \frac{L_{\text{md}}}{\xi_{mi}}.
\]  

(2.13)

The magnetic field solution and the air-gap flux density distribution for a TLA-type motor are illustrated in Fig. 2.3.
Figure 2.3. a) The q-axis flux and b) the air-gap flux density distribution (distorted curve) and the fundamental wave of the air-gap flux density (sine wave). In the figure \( I_q = 55 \, \text{A} \).

Table 2.1. Comparison between analytical formulas and FEM-analysis of ALA-type and TLA-type SynRM rotors. \( I_d = I_q \), air-gap length is 0.8 mm, insulation ratio \( w = 0.5 \), \( r = 107.2 \, \text{mm} \), pole pair number \( p = 2 \).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Analytical</th>
<th>FEM (ALA)</th>
<th>FEM (TLA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_{md} ) [mH]</td>
<td>130</td>
<td>105</td>
<td>120</td>
</tr>
<tr>
<td>( L_{mq} ) [mH]</td>
<td>Eq. (2.10)</td>
<td>4.4</td>
<td>2.3</td>
</tr>
<tr>
<td>( L_{mq} ) [mH]</td>
<td>Eq. (2.13)</td>
<td>3.9</td>
<td>2.3</td>
</tr>
</tbody>
</table>

According to the result of Table 2.1 the d-axis inductances for TLA and ALA type rotors are quite equivalent, as for the q-axis inductance the difference is remarkable. It may be assumed that the higher d-axis inductance of the TLA type rotor is caused by the smoother rotor surface, which decreases the effect of air-gap harmonics. This effect is illustrated in Fig. 2.4 which shows the air-gap flux density and its harmonic content at a current value of 10 A for TLA and ALA type rotors.

Figure 2.4. The air-gap flux densities for a) a TLA type and b) a ALA type SynRM rotor using a 10 A peak current and an 0.4 mm air-gap length.

The corresponding harmonic flux density amplitudes are shown in Table 2.2.
Table 2.2. The air-gap flux density values for different harmonics

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>TLA $\hat{B}$ [T]</th>
<th>ALA $\hat{B}$ [T]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>0.878</td>
<td>0.799</td>
</tr>
<tr>
<td>5.</td>
<td>0.018</td>
<td>0.073</td>
</tr>
<tr>
<td>7.</td>
<td>0.032</td>
<td>0.011</td>
</tr>
<tr>
<td>11.</td>
<td>0.024</td>
<td>0.133</td>
</tr>
<tr>
<td>13.</td>
<td>0.258</td>
<td>0.127</td>
</tr>
</tbody>
</table>

According to Fig. 2.4 and Table 2.2 the tangential ribs in the TLA type rotor seem to reduce the superharmonic contents of the air-gap flux density.

Because especially the d-axis inductance is very sensitive to saturation it is used a very low current value for the comparison between the analytical and FEM-calculations. According to the results, Eqns. (2.10) and (2.13) give quite equal results for the unsaturated inductances, especially for the TLA type rotor. With the analytical equations the q-axis inductance adopts values that are much higher than the FEM calculated respective values. This is referred mainly to the ribs and webs, which decrease the reluctance.

2.2 Effect of cross saturation

If the iron is magnetically linear, the flux generation processes in d-axis and q-axis are independent, so

$$\psi_{md} = L_{md}i_{md}, \quad \text{(2.14)}$$
$$\psi_{mq} = L_{mq}i_{mq}. \quad \text{(2.15)}$$

The magneto-motive force acting on the d-axis affects, however, the ability of the q-axis magneto-motive force to create flux and vice versa, i.e the flux generation processes in the d- and q-axis are not independent of each other.

In this case, the modelling of the cross-saturation may be given as:

$$\psi_{md} = \psi_{md}(i_{md}, i_{mq}), \quad \text{(2.16)}$$
$$\psi_{mq} = \psi_{mq}(i_{mq}, i_{md}). \quad \text{(2.17)}$$

Assuming that the machine is symmetrical

$$\psi_{md} = 0, \text{when } i_{md} = 0 \quad \text{even though } i_{mq} \neq 0, \quad \text{(2.18)}$$
$$\psi_{mq} = 0, \text{when } i_{mq} = 0 \quad \text{even though } i_{md} \neq 0. \quad \text{(2.19)}$$

Thus, the functions (2.14) and (2.15) may be defined in the following forms:

$$\psi_{md} = L_{md}(i_{md}, i_{mq})i_{md}, \quad \text{(2.20)}$$
$$\psi_{mq} = L_{mq}(i_{mq}, i_{md})i_{mq}. \quad \text{(2.21)}$$
The FEM calculated effects of the self and the mutual inductance cross saturations are here given only for test motor 5. This TLA-type motor is introduced later in chapter 4. The magnetising inductance values are shown in Figs. 2.5a and 2.5b. The currents in the figures are rms values.

Figure 2.5. FEM calculated self- and cross-saturation effects for the test SynRM 5 a) d-axis and b) q-axis inductances. The q- and d-axis currents respectively are used as parameters.
These figures indicate clearly that the effect of cross saturation is significant especially for the d-axis inductance and may not usually be neglected in the analysis of the synchronous reluctance motor behaviour.

2.3 Stator leakage inductance

The leakage inductances of the SynRM are generally defined applying the same analytical equations as used for the induction motors (Richter 1954, 1967). The reasons for this are that in the SynRM it may be used a stator equivalent to the IM stator and, on the other hand, these traditional equations give accurate enough results, especially, since the per unit stray inductance is small compared to the magnetising inductances.

Because at steady state the rotor of the SynRM rotates at synchronous speed there exists neither rotor current nor rotor leakage flux. Thus, the leakage inductance may be expressed only for the stator. Of course, there occurs some kind of leakage flux also in the rotor because of the conducting iron sheets or real damper windings. However, at synchronous speed there is no need to study this phenomenon in detail.

According to the equations by Richter (1963, 1967) for the stray inductances and the parameters shown in Appendix A the results are given in Figure 2.6. These results are valid for the geometry of the original 30 kW four-pole ABB squirrel cage motor. The air-gap length is a variable.

![Figure 2.6. Four-pole ABB 30 kW induction motor’s stator leakage inductance as a function of the air-gap length. The original air-gap length of the motor is 0.8 mm.](image)

The leakage inductance is proportional to the length of the air-gap. As it is shown in Fig. 2.6, decreasing the air-gap length from the initial value of 0.8 mm down to 0.3 mm would increase the total leakage flux inductance of the initial induction motor stator by about 40 %. On the other hand, in the pilot SynRM the air-gap width on the d-axis is not equivalent to the respective value on the q-axis because of the cut-out. The main target of the rotor cut-out is to decrease the magnetising inductance of the respective axis and increase thus also the reluctance torque. The cut-out, however, decreases also the stator d-axis leakage inductance and decreases thus the motor power factor slightly.

In the prototype motor the q-axis maximum air-gap length is 10 mm, as for the d-axis the corresponding value is only 0.4 mm. The d-axis magnetising currents lie mainly in the q-axis area.
Because the q-axis air-gap length is so wide, the d-axis leakage inductance consists mainly of the air-gap length independent inductance parameters, which are the slot- and the coil-end leakage flux inductances. In assuming this the d-axis leakage inductance in the case of the prototype motor is about 70% lower than the corresponding value of the d-axis. However, the q-axis air-gap is not so long in the whole q-axis pole area but only one third of the slots are in the q-axis cut-out area and thus this effect is remarkably reduced.

Since the leakage inductance is small compared to the magnetising inductances a constant value corresponding to the d-axis air-gap length for \( L_{so} \) is used in the analytical evaluations. When using the FEM-approach the variation of the leakage in different axes is automatically taken into account except for the effects of the end windings leakage. The end winding leakage inductance is in these cases assumed to be a constant.

2.4 Effects of iron losses

For motor vector control purposes the stator iron losses may usually be neglected. However, for the designing a SynRM the iron losses are an important factor because they may decrease torque production in a considerable way. For this reason, the iron loss resistances are introduced in the equivalent circuits. The effects of the iron losses on the torque production of synchronous reluctance motors were discussed in Refs (Longya 1991, Betz 1992, Levi 1995).

The equivalent circuits including the iron loss components are shown in Fig. 2.7. In the figure the resistances \( R_d \) and \( R_{eq} \) represent the equivalent steady state core losses in the motor (Longya 1991). If the iron loss components are included the direct and quadrature axis current components are

\[
\begin{align*}
    i_{sd} &= i_{md} - i_{cd}, \\
    i_{sq} &= i_{mq} + i_{cq},
\end{align*}
\]

(2.22)

(2.23)

where \( i_{cd} \) and \( i_{cq} \) express the iron loss components.

![Diagram](image-url)
The loop analysis of the parallel branch of each equivalent circuit gives the total steady state core loss as

\[ P_{Fe} = \omega^2 \left( \frac{\psi_{md}^2}{R_{eq}} + \frac{\psi_{mq}^2}{R_{cd}} \right). \]  

(2.24)

If the saliency ratio is high the core loss contribution resulting from \( \psi_{mq} \) may be neglected, thus

\[ P_{Fe} \approx \omega^2 \left( \frac{\psi_{md}^2}{R_{eq}} \right). \]  

(2.25)

When \( R_{cd} \) is assumed infinite the d-axis equivalent circuits of Fig. 2.7a may be modified to correspond that of Fig. 1.10c. Thus, the magnetising d-axis current \( i_{md} \) is now equal to the stator d axis current \( i_{sd} \). In the following approximations

\[ i_{md} \approx i_{sd}, \text{ when } \psi_{ad} >> \psi_{mq}, \]  

(2.26)

is used.

Fig. 2.8 shows the effects of the iron loss current components on the SynRM vector diagram. To illustrate the production of iron loss the d- and q-axis currents are given also in the stator voltage \( u_s \) oriented co-ordinates.
According to Fig. 2.8, the internal magnetising current angle $\tan(i_{\text{md}}/i_{\text{mq}})$ is by the degree of $\Delta \kappa$ smaller than the external current angle $\kappa = \tan(i_d/i_q)$. This decrease of the current angle is caused by the iron losses and it thus decreases the torque producing capability of the SynRM. On the other hand, the power factor improves.

The iron loss may be taken into account by subtracting the iron loss current $i_{cdq}$ from the torque production magnetising current $i_{dqm}$

$$i_{cdq} = \frac{\psi_{md}}{R_{eq}}.$$  \hspace{1cm} (2.27)

The iron losses reduce the production of the reluctance torque and the new torque is found as

$$T_{\text{rel}} = m \left( p(L_{md} - L_{mq})_d (I_q - I_mq) = m \left( p(L_{md} - L_{mq}) I_d I_{mq} \right).$$  \hspace{1cm} (2.28)

Considering the iron losses, the rotor position $\theta_1$ and the fundamental mutual inductance between d- and q- axis, the magnetising flux linkages are given by

$$\begin{bmatrix} \psi_{md}(\theta_1) \\ \psi_{mq}(\theta_1) \end{bmatrix} = \begin{bmatrix} L_{md}(\theta_1) & L_{mdq}(\theta_1) \\ -L_{mdq}(\theta_1) & L_{mq}(\theta_1) \end{bmatrix} \begin{bmatrix} i_{md} \\ i_{mq} \end{bmatrix},$$  \hspace{1cm} (2.29)

where

- $L_{mdq}$ the fundamental mutual inductance between the d- and q-axes,
- $\theta_1$ the rotor angle with respect to the stator xy-co-ordinates.
The instantaneous torque $T_e$ in the rotor reference frame including the iron loss components is

$$T_e = \frac{3}{2} p (\psi_{md}^i m_q - \psi_{md}^i m_d) \cdot$$  \hspace{1cm} (2.30)

The iron losses have two effects; firstly, they reduce the output power and, secondly, they increase the external current angle.

There are two methods to solve iron losses in the FEM calculations; the time-harmonic or the time-stepping method.

The time-harmonic method is a relatively fast method to analyse the machine performance at steady state. The equation used in FEM to calculate the iron loss power density $dP'_{fe}$ in steady state is

$$dP'_{fe} = \left( k_h \hat{B}^2 f + \pi^2 \frac{\sigma d^2}{6} (\hat{B} f)^2 + k_e ((\hat{B} f)^{1.5} \cdot 8.67) \right) k_f \cdot$$  \hspace{1cm} (2.31)

where

$\hat{B}$ - the maximum fundamental flux density of the element under investigation

$k_h$ - the hysteresis loss coefficient,

$\sigma$ - the conductivity,

$k_e$ - the excess loss coefficient,

$d$ - the sheet thickness and

$k_f$ - the stacking factor.

The loss constants $k_h$ and $k_e$ are determined by curve fitting to measured iron loss data measured on a single-sheet test rig (Mueller 1995).

Table 2.4 shows the coefficients used and determined from the curve fitting the information that is supplied by the sheet manufacturer (Surahammar 1998).

<table>
<thead>
<tr>
<th>Coefficient</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material DK-70</td>
<td></td>
</tr>
<tr>
<td>Hysteresis loss $k_h$ [WsT^{-2}m^{-3}]</td>
<td>152</td>
</tr>
<tr>
<td>Conductivity $\sigma$ (\Omega m)^{-1}</td>
<td>4 \cdot 10^6</td>
</tr>
<tr>
<td>Excess loss $k_e$ [W(Ts^{-1})^{3/2}m^{-3}]</td>
<td>2.32</td>
</tr>
<tr>
<td>Sheet thickness $d$ [m]</td>
<td>0.5 \cdot 10^{-3}</td>
</tr>
<tr>
<td>Stacking factor $k_f$</td>
<td>0.98</td>
</tr>
</tbody>
</table>

The iron loss calculation can be done also in a periodic state (time-stepped calculation over one complete period) including also the harmonics effect.

$$\frac{1}{T} \int_0^T dP'_{fe} dt = k_h \hat{B}^2 f k_f + \frac{1}{T} \int_0^T \left[ \pi^2 \frac{\sigma d^2}{12} \left( \frac{dB}{dt} (t) \right)^2 + k_e \left( \frac{dB}{dt} (t) \right)^{1.5} \right] k_f \cdot$$  \hspace{1cm} (2.32)
The motor iron loss is found by integrating Eqs. (2.31) or (2.32) over the total magnetic circuit volume.

In the time-stepping calculation one complete period must be resolved, which correspondingly increases the calculation time and the requirements for the computer memory. The FEM calculated iron loss values for both stator and rotor iron of the IM and the TLA-type SynRM are presented in Table 2.5.

<table>
<thead>
<tr>
<th></th>
<th>Total iron loss 50 Hz (time-harmonic) [W]</th>
<th>Stator iron loss (time-stepping) [W]</th>
<th>Rotor iron loss (time-stepping) [W]</th>
<th>Total iron loss (time-stepping) [W]</th>
</tr>
</thead>
<tbody>
<tr>
<td>No-load (IM)</td>
<td>274</td>
<td>284</td>
<td>22</td>
<td>306 (422)</td>
</tr>
<tr>
<td>Full-load (IM)</td>
<td>270</td>
<td>350</td>
<td>32</td>
<td>382 (496)</td>
</tr>
<tr>
<td>No-load (SynRM)</td>
<td>275.5</td>
<td>485</td>
<td>110</td>
<td>595 (690)</td>
</tr>
<tr>
<td>Full-load (SynRM)</td>
<td>275.5</td>
<td>796</td>
<td>242</td>
<td>1038 (1058)</td>
</tr>
</tbody>
</table>

The time-harmonic calculation results obtained for the IM and the SynRM are quite equal because the calculations do not take into account the effects of the harmonic fields. The majority of the harmonic iron losses occur on the rotor surface, where losses are caused by the flux pulsation under the stator slots. The air-gap length used for the SynRM is 0.4 mm and 0.8 mm for the IM and thus the surface losses are much more intense in the SynRM than in the IM. The SynRM stator slots include semi-magnetic slot-wedges, which reduce the permeance harmonics on the rotor surface. The time-harmonic calculation method ignores these pulsation losses and therefore gives far too optimistic results for the SynRM. According to the results, the time-harmonic iron loss values of the IM are about 35 % lower compared to the measured iron losses without additional losses. For the SynRM the corresponding error at no-load is about 70 %.

Bertotti (1991) mentions several factors that explain the lower calculated iron loss results for the IM: the occurring of eddy current losses in the rotor cage, and the actual deterioration of the material properties during the lamination punching and core assemblage, which have not been modelled, and also the fact that the calculations have been carried out using a two-dimensional model, where the machine end effects have been ignored.

The transient calculated iron losses for both the no-load and full load test for the IM are about 25 % lower than the measured losses. This relation is also lower for the SynRM iron losses and thus the correlation between the transient FEM calculation and the measured values is quite good. The measurements of the motor iron losses are evaluated later in chapter 5. According to the calculated values, the flux density does not change a lot as a function of the load. Therefore, it may be assumed that the correlation coefficients do not change dramatically as a function of the load.

### 2.5 Supplying of the SynRM using a VSD

In chapter 1, the torque and the other motor performance parameters have been evaluated assuming the inductances to be constant. In the evaluation of the stator flux, the voltage drop in the stator windings have been taken into account and the currents of the axes are calculated using the fluxes and inductances of the axes. Actually, the inductances of the axes saturate as a function of the flux of the axes and thus the saliency ratio does not stay constant as a function of the torque. In the following, the performance of the pilot SynRM is studied using the measured
currents of the axes and the saturated model of the inductances of the axes. The inductance models used are based on the FEM calculations obtained from the pilot motor introduced later in chapter 5.

In the VSD, a constant d-axis current may be used at speeds below the base speed to keep the flux constant (constant flux control). The so-called constant current angle control offers an alternative control method. The current angle control is defined as the control of the current vector angle $\kappa$ of the machine to obtain the best performance in terms of torque-to-current ratio or power factor (Trübenbach 1993, Betz 1992). The VSD system used in the measurements in chapter 5 is based on the constant flux linkage direct torque control, where the stator flux linkage in the stator reference frame is estimated as in Eq. 1.2.

If the stator resistance is neglected the stator voltage squared is

$$u^2 = \omega^2 [(L_d i_{sd})^2 + (L_q i_{sq})^2] = \omega^2 |\psi_s|^2.$$  \hspace{1cm} (2.33)

The equation is rearranged to

$$\frac{i_{sd}^2}{L_d} + \frac{i_{sq}^2}{L_q} = \left( \frac{u}{\omega} \right)^2.$$  \hspace{1cm} (2.34)

This is an equation of an ellipse in $i_{sd}, i_{sq}$ plane centred at (0, 0) with axes

$$2a = 2(u/\omega) / L_q \quad \text{major}$$
$$2b = 2(u/\omega) / L_d \quad \text{minor}$$

The axes are inversely proportional to the angular frequency $\omega$.

Varying the stator frequency and voltage is the technique preferred in most variable-speed IM and SynRM drive applications. As it is shown in Eq. 2.33 the voltage of an electrical machine is linearly dependent on the angular frequency $\omega$. If a balance set of three-phase sinusoidal voltages at a frequency $f = \omega/2\pi$ are applied to the stator, it results in a balanced set of currents, which establishes a flux-density distribution $B_\delta$ in the air-gap with the following properties:

1. it has a constant amplitude and
2. it rotates with a constant speed, also called the synchronous speed.

The synchronous speed in a $p$-pole motor, supplied by frequency $f$, can be obtained as

$$\omega_b = \frac{2\pi f}{p},$$  \hspace{1cm} (2.35)

which is synchronized to the frequency $f$ of the applied voltages and currents to the stator windings.

At a certain frequency, $\omega_b$, the voltage $u$ reaches the maximum voltage available from the inverter. Below this frequency, the flux modulus is kept constant but, if it is desired to increase the speed to be above this frequency, the flux modulus must be decreased. The speed range below $\omega_b$ is called the base speed area, the constant flux area or the constant torque area. The
speed range above $\omega_b$ is traditionally called the field weakening area or the constant power area. The boundary frequency $\omega_b$ is called the base frequency of the nominal frequency. The behaviour of ideal drive characteristics stator voltage $u$ ignoring the voltage drop in the stator resistance, motor flux $\psi$, output torque $T$ and output power $P_{out}$ are illustrated in Fig. 2.9.

![Graph](image1)

Figure 2.9. Ideal drive characteristics a) input voltage $u$, b) motor flux $\psi$, c) output torque $T$ and d) output power $P_{out}$. The $\omega_b$ is the rated speed.

Fig. 2.10 shows the envelope curves of the stator current vector $i_s$, the stator flux linkage vector $\psi_s$ and the stator back electro magnetic voltage vector $e_m$ for the load angle $\delta \in [0, \pi]$.

![Graph](image2)

Figure 2.10. Envelope curves for the stator current vector $i_s$, the stator flux linkage vector $\psi_s$ and the stator back electro magnetic voltage vector $e_m$. The vectors are not in scale in the figure.

In Fig. 2.10 the back electro motive voltage draws an ellipse because of the voltage drop in the stator windings as a function of torque. On the other hand, the control system keeps the flux linkage constant. The stator current envelope curve is an ellipse because of the magnetic asymmetry of the machine. The increasing voltage drop in the stator windings reduces the magnetising current as the torque increases and thus also the magnetising flux decreases.
Results for ALA- and TLA-rotors using no cross saturation

In the following, the TLA- and ALA types SynRM rotors will be analysed using the measured current vector values as a function of the measured torques and the FEM calculated magnetising inductances as a function of the currents of the axes. The lay-outs of the rotors are shown in Fig. 2.11.

Figure 2.11. The lay-outs of the studied a) ALA-type rotor and b) TLA type rotor. The ALA-rotor is manufactured of quite thick ferromagnetic and non-ferromagnetic half-pipes having the same length as the stator stack. The TLA-rotor is manufactured of thin round laminations having the radial and tangential ribs that maintain the form of the lamination.

These two rotor types have a different lay-out and also the materials used are different. The shares of the iron compared to the flux barriers are quite equal but the ALA-type rotor does not have any cut-outs and the number of the flux barriers is higher. No semi-magnetic slot wedges are used for the ALA-type rotor.

Because the ALA-type rotor is manufactured using a solid steel core and 3 mm thick iron and stainless steel sheets, consequently, the effect of the iron losses for this rotor type will be big. The TLA type of rotor is manufactured using 0.5 mm thick insulated electrical steel sheets.

The impact of cross saturation is assumed to be lower in the ALA-type rotor because the q-axis inductance and thus the corresponding flux is also smaller than in the TLA-type rotor.

Fig. 2.12 shows several examples of the behaviour of the axes current vectors in the ALA- and the TLA-type test motors as a function of the per unit torques. The constant flux linkage control method is applied. The presented estimation of the stator flux linkage \( \psi_s \) is based on the current dependent FEM calculated inductance values of the axes, however, ignoring cross saturation.
Figure 2.12. The measured stator direct and quadrature axes current rms components $I_d$ and $I_q$ using FC for a) the ALA-type and for b) the TLA-type rotor SynRM as a function of per unit torque. The equations of the trend-lines are also shown. The estimated stator and direct axes magnetising flux linkages per unit values are also given. The current measurement is based on the motor line current measured total value, which is divided into the d- and q-axis values according to the knowledge obtained from the DTC-controller.
It may be noticed from Fig. 2.12 that for both motor types the d-axis flux linkage decreases slightly (by about 2 … 2.5 %) as a function of the torque, as it is described in Fig. 2.10. For the ALA type motor the stator flux linkage reduces slightly as for the TLA type motor it slightly increases as a function of the torque. According to Fig. 2.10, the stator flux linkage should stay constant or slightly be reduced due to the stator resistance voltage drop as the torque increases. The stator flux error in Fig. 2.12 results from the fact that the cross saturation and the load dependent iron losses (additional losses) and the voltage drop in the stator windings have been ignored.

The FEM calculated axes inductances for the ALA-type rotor with 0.3 mm air-gap length and for the TLA-type rotor with slot-wedges and with air-gap length 0.4 mm is shown in Fig. 2.13. The figure and the equations presented do not include the effect of cross saturation.

![Figure 2.13](image-url)
Even though the ALA-type rotor has a smaller air-gap length it has lower d-axis inductances compared to the TLA-type rotor, which may be explained by the non-ferromagnetic sheets in the rotor construction. This, according to Carter’s theory, increases the air-gap length. Because the ALA-type rotor has no radial or tangential ribs the q-axis inductance is significantly lower compared to the TLA-type rotor.

Using the axes currents shown in Fig 2.12 and the inductance equations shown in Fig. 2.13 the following torque estimation as a function of load angle may be presented, Fig. 2.14. The estimated saliency ratios and measured torque values are also shown.

Figure 2.14. The measured and calculated per unit torque values as a function of the estimated load angle for a) the ALA type and b) the TLA-type SynRM. The torque dependent saliency ratio is also given.
Using the measured currents vector values and the FEM estimated inductance values without taking account the effect of cross saturation or iron losses the relative error between the measured and the estimated torque value is the biggest on the high torque value.

It may be noticed from Fig. 2.14 that the saliency ratios do not stay constant as a function of the torque. As the torque increases the direct axis current reduces and thus the value of the d-axis inductance increases likewise. According to Fig. 2.13, the q-axis inductance reduces much less even though the q-axis current changes much more than the d-axis current.

The control method used above (SynRM-DTC) is based on constant stator flux linkage. The corresponding current angle as a function of the torque is given in Fig. 2.15. Also the power factor using Eq. (1.36) as a function of per unit torque is shown.

Figure 2.15. The estimation of the current angle and power factor as a function of the torque. a) the ALA-rotor, b) the TLA rotor. The high power factor values are obtained with high direct axis magnetising inductances (small air-gap) and high inductance ratios. The high surface losses in the ALA-rotor are not taken into account for the calculating of the power factor.
Thus, if the constant stator flux linkage control method is used the current angle increases as a function of the torque and the power factor gets high values also at low torque values. Because of the higher saliency ratio the ALA-type motor gets higher power factor values than the TLA-type motor.

**Result for ALA- and TLA-rotors using cross saturation**

Since the above given calculated values for the torque are quite different from the measured values the effects of cross saturation and iron losses on the torque are investigated in the following.

The cross saturation effect is analysed using the inductance values given in Fig. 2.5. The calculation results of the stator and the d-axis flux using the measured axes current values of Fig. 2.12 are shown in Fig. 2.16.

![Figure 2.16. The estimated stator and direct axes magnetising flux linkages in per unit values for the TLA-type SynRM as a function of per unit torque. The flux linkages have been calculated using the corresponding motor currents and inductances. This indicates that the reluctance matrix method taking the cross-saturation into account gives accurate results. The stator flux linkage stays almost constant and corresponds thus well to the DTC-control, which tends to keep the flux linkage a constant. Only the voltage drop in the stator resistance causes the flux linkage to drop. Using the estimated cross-saturated inductance values with the estimated current values the stator flux stays constant compared to the situation in Fig. 2.12b. The effect of the iron loss may be taken into account using the iron loss current $i_{eq}$ from Eq. (2.27) and the calculated iron losses as a function of the current.](image)

**Rotor position effect**

The effect of the rotor position on the self-inductance of the axis is analysed. Rotor positions, where the d-axis is exactly under the stator tooth or exactly under a slot opening are illustrated in Fig. 2.17.
According to obtained calculation results, the effect of the rotor position on the axis self-inductance values is minor in the transversally laminated type rotor. This is mainly due to the tangential rib, which equalizes the effect of stator slot reluctance.

There are used two different methods to analyse the effect of the rotor position on the cross saturation and thus on the torque production: the rotating of the stator current vector keeping the rotor locked or the rotating of both the stator current vector and the rotor.

In the locked rotor model the effect of the rotor position is taken into account by turning the rotor by half a slot pitch and by defining the average torque using the maximum and minimum torque values as

\[ T_{\text{average}} = T_{\text{max}} \cdot \frac{\tau_{\text{tooth}}}{\tau_{\text{slot opening}} + \tau_{\text{tooth}}} + T_{\text{min}} \cdot \frac{\tau_{\text{slot opening}}}{\tau_{\text{slot opening}} + \tau_{\text{tooth}}}, \]  

where the \( \tau_{\text{slot opening}} \) is the width of the stator slot opening in the stator surface and \( \tau_{\text{tooth}} \) the width of the stator tooth as it is illustrated in Fig.2.18.

Fig. 2.19 shows an example of the torque calculated with FEM with the rotating current vector in the locked rotor model in two different rotor positions. A torque curve calculated using \( L_d \) and \( L_q \) without cross saturation is also given. The self-saturation is taken into account.
Figure 2.19. The torque values as a function of the current angle. The marks $T_{\text{min}}$ and $T_{\text{max}}$ indicate the rotor positions where the $d$-axis inductance reaches its maximum or minimum values. A measured torque point using the external current angle value of $80^\circ$ as well as a calculated torque curve using $L_d$ and $L_q$ without cross saturation is also given.

According to Fig. 2.19, the difference between the maximum and the minimum torque values at a nominal current value of 55.6 A seems to reach its biggest value at a current angle of about 60 degrees where the gap between the maximum and average torque value is about 0.04 p.u. If cross saturation is ignored ($T(L_d(I_d), L_q(0, I_q))$) the torque gets lower values at smaller current angles but bigger values as the current angle increases. The average torque values differ only slightly from the measured values at the rated load point.

The torque ripple using the locked rotor torque calculation method may be evaluated as

$$\text{Torque ripple} = \frac{T_{\text{max}} - T_{\text{min}}}{T_{\text{average}}} \cdot 100\% \quad (2.37)$$

Using the parameters of Fig. 2.19 the torque ripple is at nominal current 15.9 %, which is lower than when the rotating rotor model is used, as it is introduced next. The locked rotor method has the disadvantage that it does not allow to evaluate the effect of skewing the rotor.

In the following, the effects of the iron losses are evaluated using the FEM calculated iron loss values as a function of the per unit current. The FEM calculated $P_{\text{Fe(FEM)}}$ results are compared to the measured iron loss values, Fig. 2.20.
Using the iron loss line equation from Fig. 2.20 and the estimation of the d-axis flux linkage \( \psi_{md} \) from Fig. 2.12, the iron loss resistance \( R_{cl} \) may be calculated as:

\[
R_{cl} = \frac{\omega \psi_{md}^2}{P_{Fe(FEM)}}.
\]  

(2.38)

The calculated iron loss resistance and iron loss current values are given in Fig. 2.21.

According to the result shown in Fig. 2.21, the iron loss resistance does not stay constant as a function of the torque. This may be due to the increasing leakage flux, which increases the volume of the load dependent iron losses. On the other hand the iron losses increase the iron loss current component as a function of the load.

Using the iron loss current component the behaviour of the internal magnetising current angle as a function of the torque is shown in Fig. 2.22.
Figure 2.22. Curves of the external current angle $\kappa$ and the internal current angle $\kappa_m$ values as a function of the per unit torque.

Even though the iron loss increases as a function of the torque, the gap between the external and the internal current angles reduces. This is because the share of the iron loss current component compared to the total current component decreases.

The iron loss current component saturates the motor iron but it does not produce the torque. This saturation effect of the iron loss component is not present in the equivalent circuit of the SynRM in Fig. 2.7. In the following, this saturation effect of the iron loss is taken into account in the calculation model. The model uses the motor reluctivity matrix, which is produced by using the motor total input current. In the torque production the absence of the iron loss component is taken into account by decreasing the torque producing current vector modulus by the iron loss component, Fig 2.23b. The comparing results without the iron loss effect are shown in Fig. 2.23a.
Figure 2.23. The torque values as a function of the current angle when a) the iron loss effect is ignored and b) the iron loss effect is taken into account. The measured torque values as a function of the external current angle are also shown.

When it is used the reluctivity matrix, which takes into account the total input current saturation effect, and the magnetising current, where the iron loss current component is taken out, the results obtained are quite accurate compared to the measured values.

Similarly, the reluctivity matrix and the reduced current vector methods may be used in the rotating rotor model. In this model the estimated internal current angle is used. The calculated results are shown in Fig. 2.24.

Figure 2.24. The FEM calculated torque values as a function of the rotor angle. The iron loss effects are taken into account in the current modulus and in the current angle. The comparing measured torque value lines are also shown.
Even though some estimation of the torque ripple in the locked rotor model can be calculated fairly easily if only the maximum and minimum torque values for a certain current vector and rotor positions are found the rotating rotor model has the advantage over the locked rotor model that it offers the possibility studying the rotor geometry effects on the torque performance in detail as a function of the rotor mechanical rotation. This enables the designing of a rotor that gives less torque ripple. Chapter 4.2.3 introduces the method in more detail.

Fig. 2.25 compares the results of the rotating rotor model method compared with the results of the non-rotating method and the measured values. The torque calculation marks used in the figure are

- \( T(L_{md}(I_d,0),L_{mq}(0,I_q)) \) Torque calculated using Eq. (1.22) using axis self inductances.
- \( T(L_{md}(I_d,I_q),L_{mq}(I_d,I_q)) \) As previous but using axes cross saturations.
- \( T(L_{md}(I_d,I_q),L_{mq}(I_d,I_q),I_{mq}) \) Torque calculated using Eq. (2.28) taking into account the effect of iron loss current.
- \( T(I_s,\kappa_m,\theta) \) Torque achieved from FEM rotating rotor and current vector model. The iron loss effect is taken into account in the current angle value used and in the current modulus. The saturation of the total current is taken into account using the appropriate reluctivity matrix.

Figure 2.25. The calculated and measured torque values as a function of the per unit current. The calculation method is a parameter.

According to Fig. 2.25 the optimal calculated torque values compared to the measured values are achieved with the rotating rotor model. Also the method where the iron loss current component is taken into account in the calculation gives quite reasonable results.

The effect of the cross saturation and the iron losses on the saliency ratio is illustrated in Fig. 2.26. In the case of rotor rotation, the saliency ratio is evaluated by using the q-axis inductances from Fig. 2.5 and by setting the d-axis inductance to be a variable. The d-axis inductance is thus calculated from torque equation (1.22).
Figure 2.26. The saliency ratio as a function of the measured per unit current. The cross saturation and the cross saturation with the effect of iron losses in the current angle and the rotor rotation are parameters. The leakage flux inductances are set constant.

In Fig. 2.26 the saliency ratio increases as a function of the current when the cross saturation and the iron losses with cross saturation are taken into account. The effect of the cross saturation on the saliency ratio seems to be more significant than the effect of the evaluated iron losses.

Because the load-dependent additional losses, especially for the ALA-type rotor, are very considerable the measured efficiency curves as a function of the per unit torque for these two rotors are shown below, Fig. 2.27. The ALA-rotor is manufactured of 3 mm thick steel and stainless steel plates. The ALA-rotor core is of solid material. The TLA-rotor is manufactured of 0.5 mm electrical steel laminations. According to the results in Fig. 2.27 the ALA-rotor should be manufactured of much thinner and more resistive material than the material of the prototype rotor. Also skewing of the ALA rotor might be of great importance in minimising the rotor losses.

Figure 2.27. The measured efficiency values of the ALA-type and the TLA-type SynRM as a function of the per unit torque (Haataja 2000).
Even though the saliency ratio of the ALA-motor is very high the motor efficiency is rather poor. This is due to the high eddy current losses in the thick laminations of the rotor with no skewing. The stator slot openings cause large magnetic flux density variations on the edges of the thick ALA-laminations thus causing large losses. The ALA-rotor should be manufactured using very thin, high resistivity and possibly skewed laminations in order to avoid excessive rotor losses. Increasing the air gap length of the test-ALA-motor decreases the motor power factor but improves the efficiency because of the reduced iron losses. The final result, however, is not at all satisfactory. The ALA-rotor certainly offers a high potential in theory but probably the manufacturing difficulties will efficiently arrest its success in the future. At least, the ALA-rotor type requires a manufacturing technique that is totally different from that of the IM and the TLA-rotor SynRM.

According to the study, the ALA-rotor produces a very high inductance ratio and thus a good power factor. However, a careful design and the selection of the materials, especially the insulation material, must be performed in order to avoid excessive rotor losses and a poor efficiency.

2.6 Conclusion

The SynRM needs a large inductance ratio to produce good properties. Especially the d-axis inductance must be large and this requires a small air gap. It is easier to produce a large inductance ratio by using the ALA-type rotor, but especially this rotor type suffers from rotor surface eddy current losses. Using suitable materials and a geometry preventing the eddy currents might give very good properties for a SynRM equipped with this rotor type. The TLA-rotor is, however, much more easier to manufacture than the ALA-rotor and is, consequently, of greater interest for the electrical machines manufacturing industry. Because of the easy manufacturing of the TLA-rotor type this type will be investigated in this thesis in the next paragraphs.
3 ANALYSIS OF SYNCHRONOUS RELUCTANCE MOTORS BASED ON THE RATED VALUES OF INDUCTION MOTORS

This chapter introduces a technical comparison between synchronous reluctance motors and three-phase induction motors in the power range of 1.1 kW up to 90 kW. The comparison is based on the existing data about commercial IMs and calculations of the SynRM. The present level of the performance parameters of both motor types is given.

3.1 Induction motor efficiency

Temperature-rise is one of the main limiting factors that determine the motor rating. This is mainly because of the temperature tolerance of the insulation. The cooling and the losses in the machine are the factors that affect the temperature. Cooling depends on the type of the motor frame and the cooling fan. Induction motors are typically totally enclosed IP 55 types that have their own cooling fans attached onto the motor shaft (IC 01 41).

The losses in induction machines may be sorted into five groups: iron losses, ohmic losses in the stator and the rotor, mechanical losses and additional losses. Fig. 3.1 shows the typical distribution of the individual losses of induction motors.

![Figure 3.1. Typical loss distribution of 4-pole totally enclosed fan-cooled squirrel-cage motors (Auinger 1997).](image)

In Fig. 3.1 the additional load losses are assumed to be equal to the estimated 0.5 % of the power input according to the standard IEC 34-2. Because the base data used in Fig. 3.1 are from the 1980’s the table does not illustrate exactly the loss distribution of today’s motors. This is clearly indicated by the share of iron losses. The iron material quality and manufacturing accuracy of standard motors have been improved and, consequently, the share of iron loss of the ABB motors is lower than the respective iron loss curve. However, to use these data in the following overall evaluation of different losses in different motors will be no mistake.

Because of the large number of manufacturers and the wide variation of efficiencies obtained by different manufacturers the products of only one manufacturer were selected for this study.
The reason for selecting ABB motors is their leading position on the motor market, and the product’s high level of motor efficiency in the studied power range 1.1 kW ... 90 kW. In Fig. 3.2 the motor data of the ABB products are indicated.

![Figure 3.2. The efficiencies of the ABB four-pole, standard induction motors as a function of the power. The data are collected from the EURODEEM database (EURODEEM 2000).](image)

### 3.2 SynRM comparison with induction motor

It is obvious that there is not yet achieved a similar level of experience in the designing of synchronous reluctance motors as it is attained e.g. in the designing of induction machines. There are no catalogues available about SynRMs because their volume of commercial production is minor. One fundamental element in this thesis has been the designing and studying of a SynRM using a stator frame and winding equivalent to that of the high efficiency 30 kW IM by ABB. For that reason, the general comparison of the SynRM and IM presented here is based on existing data of commercial IMs. The stator stacks and the air-gap lengths, the used material qualities and the numbers of turns in the stator slots are given values that are used for commercial IMs. Most of the IM values used in the general evaluation are based on the EURODEEM database.

At first, an overview of some preceding studies is presented. The comparison uses the values for the same stator frame, windings and air-gaps utilising the data given in chapters 1 and 2. The evaluation is based on the efficiencies of the motors, estimations of typical loss distributions at each power level, no-load and full-load current values and power factors. In the evaluation a virtual 400 V star connection is assumed to be irrespective of the actual connection of the motor.

The IM stator inductance at a given air-gap may be evaluated using the equation

\[
L_s = L_{md} + L_{sc} \approx \frac{1}{\omega} \sqrt{\left(\frac{U}{I_0}\right)^2 - R_s^2},
\]

(3.1)
where $L_{\sigma}$ is the stator leakage inductance, $U$ the rated phase voltage, $I_0$ the no-load current and $R_s$ the stator resistance. The results of Eq. (3.1) are illustrated below in Fig. 3.7. Eq. (3.1) does not take into account the effect of iron losses but the error thus made is minor because of the low power factor of the no load current. The no-load current consists of the magnetising current $I_m$ and the iron loss current $I_{Fe}$.

$$I_0 = I_m + I_{Fe}. \quad (3.2)$$

The share of the magnetising current of the no-load current is big. This share is evaluated as a function of the motor power by comparing the iron loss resistance $R_{Fe}$ to the magnetising inductance $L_m$ at the rated frequency 50 Hz of the motor. The value of the iron loss resistance is estimated by calculating the iron losses $P_{Fe}$ from the total losses using Fig. 3.1 and estimating the magnetising voltage to be 98% of the total voltage as

$$R_{Fe} = \frac{(0.98U)^2}{P_{Fe}} m \quad (3.3)$$

The magnetising inductance is calculated by using the equation presented below in Fig. 3.7. The ratio $I_m/I_0$ is shown in Fig. 3.3.

![Figure 3.3. The ratio $I_m/I_0$ as a function of the motor power.](image)

It can be noticed from Fig. 3.3 that the relative influence of the iron loss resistance decreases as a function of motor power size. In the following the average value of 0.96 is used to calculate the magnetising current from the total no-load current so that $I_m = 0.96 I_0$.

A curve of the ABB 400 V, 50 Hz and four-pole IM no-load currents as a function of the motor power is shown in Fig. 3.4.
Figure 3.4. No-load currents of the ABB 400 V, 50 Hz four-pole motor as a function of the motor power.

The rated currents of the respective IM are shown in Fig. 3.5.

The stator resistance is evaluated utilising the efficiency values of the ABB motor shown in Fig 3.2 and a state of art estimation of typical loss distribution in Fig. 3.1.

According to the estimation shown above the stator resistance may be calculated using equation

\[ I_0 = -0.002P^2 + 0.76P + 1.16 \]

Figure 3.5. Rated currents of the ABB induction motor as a function of the motor power.

\[ I_N = 1.76P + 1.58 \]
\[ R_s = \frac{P_N \left( \frac{1}{\eta_N} - 1 \right) \chi_{\text{Cus}}}{m \cdot I_N^2}, \]  

(3.4)

where \( I_N \) is the rated current, \( \chi_{\text{Cus}} \) the share of the stator ohmic losses and \( m \) the number of phases. The results based on the values of the ABB catalogue standard motors are shown in Fig. 3.6.

The estimated magnetising inductance \( L_m \) used in the evaluation is calculated using no-load current values of ABB motors (Fig. 3.4) as

\[ L_m = \frac{E_d}{\omega I_m} \approx \frac{0.98U}{\omega 0.96I_0}. \]  

(3.5)

The coefficient 0.98 takes into account the voltage drop in the stator winding and the share on the magnetising current 0.96 from the total no-load current. Fig. 3.7 illustrates the estimated results of the magnetising inductance \( L_m \) of the ABB induction motors and the stator inductance with leakage flux inductance \( L_d = L_m + L_{\text{leak}} \) based on Eq. (3.1).
The SynRM torque values with different saliency ratios $\zeta$ are estimated as

$$T_{\text{SynRM}} = m \cdot p \cdot I_d \cdot I_q \left( L_m - \frac{L_m}{\zeta} \right), \quad (3.6)$$

where the axes’ currents are calculated as

$$I_d = I_0 \quad (3.7)$$

$$I_q = \sqrt{I^2 - I_0^2} \quad (3.8)$$

The calculated torque values using current Eqs. (3.7) and (3.8) from Figs. 3.4 and 3.5 and the magnetising inductance given in Fig. 3.7 are shown in Fig. 3.8. The saliency ratio of the SynRM is a parameter.
Figure 3.8. The torque values of different four-pole, 50 Hz synchronous reluctance motors compared to the torque of same size induction motor torque. The induction motor power lies on the x-axis. The synchronous reluctance motors have d-axis magnetising inductances equal to the magnetising inductance of the original IM. The current values of the axes are assumed to be according to Eqs (3.6) and (3.7). The saliency ratio \( L_d/L_q \) is a parameter.

Fig. 3.8 proves that if it is used a SynRM d-axis inductance having the same size as that of the IM the same current level does not produce a high enough torque in the SynRM unless the saliency ratio is unrealistically high. To lower the d-axis current and to produce more torque the d-axis magnetising inductance of the SynRM must thus be increased.

If the stator of an IM is used in a SynRM the air-gap length of the SynRM should be decreased to achieve a comparative torque per current ratio. By decreasing the air-gap length the d-axis inductance increases and the d-axis current decreases. The decreasing of the d-axis current improves also the motor power factor. The main disadvantage of the smaller air-gap is that especially the rotor surface iron losses increase because of the intensified spatial harmonics.

In the following the magnetising inductance is estimated as a function of the air-gap length. The total magnetising inductance \( L_m \) is

\[
L_m = \frac{E}{I_m\omega} = \frac{\mu_0 m}{\pi} \frac{DL}{\delta_k} \left( \frac{N}{p} \right)^2 \xi_1^2. \tag{3.9}
\]

Decreasing of the air-gap affects almost inversely the magnetising inductance. The accuracy of this assumption deteriorates from the impact of the effective air-gap length, which takes into account the Carter coefficient and the mmf of the ferromagnetic parts. The relative ratio between the real air-gap \( \delta \) and the effective air-gap \( \delta_k \) decreases as a function of the motor power.

This inverse air-gap effect is used to convert the magnetising inductance values given in Fig. 3.7 to the decreased air-gap length \( \delta_k \) as
The air-gap length $\delta_2$ follows Richter’s air-gap length $\delta_1$ equation for four-pole motors

$$\delta_1 = 0.1 + 0.145\sqrt[3]{P}$$

but the length is scaled so that the air-gap length is 0.4 mm at 30 kW and the minimum value for the air gap length at 1.1 kW is 0.2 mm. The adjusted air-gap length is shown in Fig. 3.9. The fixing of the 30 kW motor air-gap length is based on the results from the pilot motor presented in chapter 5.

The respective magnetising inductance values calculated using the decreased air-gap length ($\delta_2$) are given in Fig. 3.10. The proportional value of the initial air-gap value $\delta_1$ shown in Fig. 3.7 is also given.
Figure 3.10. The calculated machine magnetising or d-axis inductance values using the initial ($\delta_1$) and the adjusted air-gap length ($\delta_2$) as a function of the power.

The SynRM torque production using the decreased air-gap $\delta_2$ is shown in Fig. 3.11. In the calculation the d-axis magnetising flux $\psi_{md}$ is set equal with the initial motor as

$$I_d(\delta_2) = I_d(\delta_1) \frac{L_m(\delta_1)}{L_m(\delta_2)}.$$  \hspace{1cm} (3.12)

Figure 3.11. The SynRM torque using the decreased air-gap length and initial d-axis flux as a function of the IM power. The saliency ratio is a parameter ($L_d/L_q = 50, 10, 5$). The IM catalogue torque values are also shown. The induction motor power lies on the x-axis.
According to Fig. 3.11 the decreased air-gap produces a reluctance torque that is high enough to compete with the original IM. The results illustrated in Fig. 3.11 ignore both the effect of saturation and iron losses, which both decrease the reluctance torque. For this reason the results obtained are somewhat optimistic.

The effect of the air-gap length to the SynRM vector diagram has been illustrated in Fig. 3.12.

![Vector Diagram](image)

Figure 3.12. Comparison of the vector diagrams of the SynRM at the original air-gap length \((\delta_1)\) and the decreased air-gap length \(\delta_2\). The absolute value of the current and the d-axis flux are set constants.

According to Fig. 3.12 the decreasing air-gap length increases the motor power factor and the motor output torque assuming that the increase of iron losses is minor.

The share of the air-gap reluctance of the total q-axis reluctance is low and thus the decreasing of the air-gap length does not affect remarkably the quadrature axis inductance meanwhile the d-axis inductance increases rapidly. Thus, decreasing the air-gap length increases the SynRM saliency ratio, the torque per current ratio and the power factor. In this analysis the effect of the increasing iron losses caused by the harmonics are ignored. Decreasing the air-gap length will increase remarkably the effect of the air-gap harmonics. This again increases both no-load and load dependent losses of the motor. On the other hand, the decreasing air-gap length increases also the stator leakage inductance, which further has a negative effect on the motor power factor.

### 3.3 Maximum torque

The maximum torque is one of the performance elements of the motor. The estimated maximum torque values for the SynRM using Eqs. (1.29) and (1.34) are represented in Fig. 3.13. The respective relative value for the IMs maximum torque is between 2.4 … 3.2 (average 2.9).
According to the results shown in Fig. 3.13 the maximum torque is generally lower than in IMs. Normally, according to IEC34, the maximal torque must be at least 160 % of the rated torque. The saliency ratios, which produce at least 160 % of the rated torque, are illustrated in Fig. 3.14. The figure shows also the corresponding maximum torque.
According to Fig. 3.14, to produce a high enough maximum torque in larger motors ($P_N > 15$ kW) a saliency ratio higher than 10 is needed if the initial IM air-gap length is used. In smaller sizes the SynRM is thus much more competitive than in higher power ranges.

### 3.4 SynRM power factor

In the following the 30 kW IM and SynRM power factors have been analysed using motor per-unit components. The per-unit values express the quantities in a dimensionless form. The currents of the axes have been defined as

\[
\begin{align*}
    i_s &= \frac{I_s}{I_{SN}}, \\
    \phi_s &= \cos \varphi, \\
    i_q &= \frac{I_q}{I_{SN}}, \\
    \phi_q &= \sin \varphi, \\
    L_d &= \frac{1}{i_d}, \\
    L_q &= \frac{1}{i_q},
\end{align*}
\]

Using the IM rated load power factor value from Fig. 3.18, the 30 kW motor power-producing current $i_q = 0.84$, so the magnetising current value is $i_d = 0.54$. The magnetising flux is set to $\psi_m = 1.0$ p.u and thus the magnetising inductance is $L_m = 1.84$ p.u. The stator resistance $R_s$ is 0.01 p.u. of the motor total impedance.

Fig. 3.15b proves that using the same current and inductance parameters and neglecting the stator resistance component, the SynRM in Fig. 3.15a with a saliency ratio of 10 gets a power factor value of 0.75, and if the stator resistance is included the value is 0.76 in.

Figure 3.15. A vector diagram of a SynRM for determining the power factor using corresponding IM per-unit parameter values. The iron loss current components are not considered. If the stator resistance is ignored the power factor $\cos \varphi = 0.75$ (Fig. a.) and if it is included $\cos \varphi = 0.76$ (Fig. b).
The SynRM power factor as a function of the saliency ratio neglecting stator resistance is represented in Fig. 3.16.

![Diagram](image)

Figure 3.16. The SynRM rated point power factor as a function of the saliency ratio. The d-axis magnetising inductance has been set equal to the IM magnetising inductance.

According to Figs. 3.15 and 3.16, if the d-axis current is kept constant the power factor of the SynRM could not be better than that of the corresponding value of the IM. To improve the SynRM power factor to a level better than that of the corresponding IM the d-axis magnetising current vector should be smaller.

To calculate the SynRM power factor while including the possibility of taking into account the stator resistance following equation (Feng 1993) is applied

$$
\cos \varphi = \frac{\omega Q \sin \kappa \cos \kappa - L_q \sin \kappa \cos \kappa}{\sqrt{\omega^2 \left( L_q^2 \sin^2 \kappa + L_d^2 \cos^2 \kappa \right) + R_s^2 + \omega^2 R_s \sin \kappa \cos \kappa \left( L_d - L_q \right)}}
$$  \hspace{1cm} (3.17)

where the current angle $\kappa$ is the angle between the rotor d-axis and stator current vector and assuming the q-axis flux minor.

It can be noticed from Eq. (3.17) that the power factor of the SynRM increases as the d-axis inductance and the stator resistance increase. The increase of the latter component is not advisable because it increases the stator ohmic losses. The decrease of the q-axis inductance improves also the power factor. The value of the trigonometric factor $\sin \kappa \cos \kappa$ reaches its maximum at the angle 45°. Because the angle between the d-axis flux vector and the stator current vector affects heavily also the saturation level of the inductances the angle 45° may not produce the maximum power factor as it has been shown in Fig. 1.16.

The stator current is a function of the power factor value and thus the power factor affects the stator windings ohmic losses. The stator ohmic losses make up the largest share of the IM losses as shown in Fig. 3.1. The power factor of an IM increases as a function of the load and motor rated power. This effect is illustrated in Fig. 3.17.
As it is shown in Fig. 3.17 the IM full load power factor varies between 0.8 and 0.88 in the power range of 1.1 … 90 kW and increases as a function of the load and rated power. The effect of the pole-pair number on the power factors is minor if the pole-pair number is kept small ($p \leq 3$).

Using Eq. (1.36) and the IM power factor data of Fig. 3.17 the SynRM current angle values at rated load as a function of the motor power are represented in Fig. 3.18.

An estimation of the SynRM power factor is given in Fig. 3.19 using Eq. (3.17) at 50 Hz. The results are represented in Fig. 3.19a without the stator resistance and 3.19b with the stator resistance. The current angle equation given in Fig 3.18, the inductance equation given in Fig 3.11 and the stator resistance equation given in Fig. 3.6 are used. Also the rated load values of the corresponding ABB induction motor and the power factor values using Eq. (1.39) are included in Fig. 3.19.
Figure 3.19. The IM and SynRM power factor as a function of the motor power and different saliency ratios using Eqs. (1.39) and (3.17). a) ignoring the stator resistance and b) including the effect of the stator resistance. The d-axis magnetising inductance is calculated for the initial IM air-gap.

If the air-gap length is kept equal to the IM air-gap according to Fig. 3.19 a saliency ratio as high as 50 is needed to attain the power factor level of the induction motor. The difference between the IM and SynRM power factor becomes only slightly smaller at smaller loads. The IM power factor is increasing as a function of the motor power but, according to Fig 3.19b, the power factor of the SynRM seems to stay quite stable in the given power range. At a low saliency ratio ($\zeta=5$) and ignoring the stator resistances Eq. (3.17) gives values that are quite equal to the values of Eq. (1.39). As the saliency ratio increases also the difference between these two
equations increases. At the same time, the effect of stator resistance decreases because the difference of the used current angle in Eq. (3.17) and the maximum power factor current angle of Eq. (1.38) is becoming bigger. For saliency ratio 5 the maximum power factor current angle according to Eq. (1.38) is 66 electrical degrees and according to Fig. 3.19 the average current angle is about 56 electrical degrees. For saliency ratio 10 the current angle according to Eq. (1.38) is 72 degrees and for saliency ratio 50 the corresponding value is 82 degrees while the respective value using Eq. (3.17) has stayed constant at 56 degrees.

3.5 Efficiency

The efficiency of the SynRM might be assumed to be higher than the efficiency of a corresponding induction motor because of the absence of the rotor ohmic losses. Since, however, the power factor of an IM is often better than the power factor of a SynRM and the per unit slip of a larger IM is small the discussion brought up here cannot be straightforward. According to the results presented in the previous section the torque production of a SynRM is heavily proportional to the d-axis magnetising inductance of the motor. According to Eq. (3.9) the increasing of the d-axis inductance can be achieved by decreasing the air-gap length, by increasing the number of stator winding turns or by increasing the size of the rotor.

Since the IM has been developed already for a considerable long time it can be assumed that the thickness of the stator yoke has reached its optimum value. Because the load current values of the SynRMs are assumed equal to those of the respective IM increasing the stator bore is not here investigated.

If the length and the cross sectional dimensions of the SynRM stator structure are kept the same as those of the corresponding IM also the shape of the stator slot is assumed to be equivalent to that of the corresponding IM. Increasing the number of the stator winding turns would increase the inductance values but it also decreases the cross sectional area of the winding wires. The reduction of the stator conductor cross-sectional area would with no change in load current increase the stator ohmic losses. Because the stator ohmic losses increase when using equal slot cross section with the IM this eventuality has not been investigated in this thesis. At equal load current the stator ohmic loss will increase also in the case where the stator stack length is increased and the number of coils is kept equal. Increasing the stator stack length with the same number of coils increases the inductance. This, however, causes that also the stator winding length increases. The increase of the coil length increases the winding resistance and thus with the equivalent load current also the stator ohmic loss.

According to Fig. 3.10 the air-gap of an IM increases as a function of the power. To increase the SynRM d-axis inductance value the air-gap length of the SynRM must be made smaller than that of the corresponding IM. The reduction of the air-gap length brings some disadvantages. One major problem is the increase of iron and windage losses. If the stator slot openings are kept equal to the corresponding IM stator values, fitting a semi-magnetic slots wedge may reduce both of the losses. The semi-magnetic slot wedges reduces the harmonics in the air-gap flux density distribution and thus the magnitude of iron losses. The slot wedges also make the stator surface smoother and reduce the windage loss.

If it is assumed that the stator slot wedges smoothen the harmonic content of the air-gap flux to an equal level as that of the corresponding IM, then the rotor ohmic loss is the only element that may increase the SynRM efficiency to become higher than the corresponding IM efficiency. On the other hand, it may be assumed that the SynRM has generally a lower power factor than the corresponding IM. The SynRM torque per current ratio cannot be higher than that of the respec-
tive IM. It can also be assumed that the SynRM with a lower air-gap length and a similar stator construction has more air-gap flux harmonics than the corresponding IM. From this point of view, the advantage from the absence of rotor ohmic loss must be remarkable to achieve an efficiency higher than the efficiency of a competing IM. This proves that the SynRM may compete successfully with the IM at low rated powers.

In the examination of the impact of rotor ohmic losses on SynRM efficiency it has been assumed that the iron or the mechanical losses do not differ from the corresponding IM values. Also it has been assumed that the saliency ratio is so high that the motor power factor is equal to the IM value.

Fig. 3.20 shows the absolute reducing effect of rotor loss to ABB four-pole IM efficiency as a function of the motor power.

![Figure 3.20: The effect of four-pole IM rotor ohmic losses on the ABB motor efficiencies.](image)

According to Fig. 3.20 the effect of the rotor ohmic losses on the motor efficiency decreases as the motor rated power increases. The impact of the rotor ohmic losses at a power range below 15 kW are more significant, about 3.0 %-unit than at higher powers. The impact of the rotor losses on the motor efficiency is illustrated in Fig. 3.21. The figure shows also the efficiency limit for EFF1.
If the air-gap length is kept the same as that of the IM, according to Fig. 3.9 more current must be fed to the SynRM to achieve the torque needed. Note, that in this assumption the effect of the optimum current control has been left out of the perusal and thus this allegation is not included to this study.

On the other hand, the increased iron losses caused by the air-gap harmonics and the increased stator ohmic losses caused by the lower power factor reduce the effect of the higher SynRM efficiency compared to the IM. The IM rotor ohmic losses and the lower power factor at lower power range area make the SynRM a marginally superior alternative to the IM, especially at low powers. The main reason for this is that the air-gap length in this power range is small also in the IM. At a higher power range area the power factor of the IM becomes higher, the relative slip value smaller and the air-gap length bigger which clearly render the advantages of the SynRM ineffective, when evaluating the motors efficiency.

When induction motors are fed by converters the rotor current losses and stator iron losses are increased. The harmonics in variable speed drives increase the iron losses of the IM compared to the sinusoidally supplied motor. Reduction of the harmonic contents may be done on both the converter side and the motor side. From the converter side motor time harmonic losses can be reduced by increasing the switching frequency or by filtering the motor current. On the other hand, the converter losses will increase when the switching frequency is increased.

The motor space harmonics consist of the permeance and the winding harmonics. There are several methods of reducing the permeance harmonics: making the stator slot opening as small as possible, increasing the air gap length, modifying the shape of the slot opening and using a semi-magnetic slot-wedge. The winding harmonics may be reduced by increasing the number of slots per pole and phase \( q \) or by spanning the windings.
3.6 Conclusion

The advantages of the SynRM compared to the induction motor result mainly from the synchronous rotational speed and the high $L_d/L_q$ ratio. The lighter rotor of the SynRM makes it marginally suitable for servo motor drives where a high power per weight ratio is required. However, the maximum torque of the SynRM is normally not large enough for servo applications where very high maximum torque values are needed. At the same time, the SynRM rotor construction is weaker than that of the IM. Generally, the absence of the rotor bars is also considered to be one of the benefits of the SynRM. However, according to the experience of the author, the addition of damper windings or bars to the rotor decrease the torque oscillations and may, in some cases, increase the mechanical strength of the rotor. Because the only currents induced to the rotor bars are caused by the air-gap harmonics, the share of those losses remains small. If no damper windings are used, the SynRM calls always for a converter supply.

The SynRM needs a smaller air-gap than the IM to produce an acceptable fundamental power factor. This is a major disadvantage that may considerably complicate the use of the SynRM applications where the IM air-gap is increased according to the regulations. Such applications are e.g. the explosion proof applications.

The smaller air-gap causes some other disadvantages. The efficiency of the motor suffers from the rotor surface losses. Therefore, it is suggested that better iron materials should be used for the SynRM than for the IM.

The heat sources in the SynRM rotor are the iron losses caused by the air-gap flux harmonics. Thus, if the air-gap harmonics can be decreased to a tolerable level using for example semi-magnetic slot wedges it can be assumed that the heat rise of a SynRM rotor is lower than that of the corresponding IM rotor. This is an advantage compared to the IM because, especially in closed induction motors the rotor is always more difficult to cool than the stator. If a standard induction motor is selected for a converter drive the continuous output power must be derated. With the lower output temperature rise of the motor is maintained at the original level in spite of the fact that converter supply has caused additional losses in the motor. The reduction level depends on several factors in the converter and motor system, which are the converter type and modulation, the motor design and control equipment. Each of these factors can occur in several variants. According to converter manufacturer’s own guidelines this derating varies from 0% up to 20% at normal frequency. If the induction motor is loaded below its normal frequency this derating volume must increase up to 50% or an external cooling system must be used.

When the supply frequency of the IM is lowered in an FC drive the p.u. slip increases and weakens the efficiency of the IM. In this case, the synchronous running of the SynRM is a benefit. At its lowest speeds the SynRM will be superior with respect to its efficiency.

More accurate rotor speed control is enabled with the SynRM synchronous speed and rotor high saliency ratio $L_{dm}/L_{qm}$ than it is possible with the IM with slip. Thus, with the SynRM high quality speed control without a rotor position encoder for very low speeds and even for zero speed is possible. In non steady state operation with the fluctuations of the load torque the absence of a damper winding sets high challenges for the motor control system.
4 DESIGN OF A SYNCHRONOUS RELUCTANCE MOTOR

The aim of this chapter is to find a rotor design suitable for the TLA-SynRM using the same stator as it is designed for the standard four-pole 30 kW induction motor. The variables investigated here are the length of the air-gap, the width and location of the flux barriers and the width of the tangential and radial rotor ribs. Also the width of the rotor q-axis cut-out are analysed. The design is carried out using the two-dimensional finite element method, 2D FEM. The SynRM rotor structure must be designed in such a way that it is possible to achieve the maximum torque per current ratio with minimum torque ripple. The calculation method used to evaluate the torque takes into account the effect of cross saturation between the d- and q-axis inductances, but the effect of the iron losses on torque production is ignored.

4.1 Calculation model

The different rotor structures are analysed utilising commercial FEM software packages. The dimensions of the stator are kept the same as in the original 30 kW four-pole IM. The design is based on a 48-slot machine with full pitch stator windings. The number of turns in series per phase is also kept equal to the corresponding induction motor value.

Fig. 4.1 shows the variables investigated in the FEM model. The arc radius origin \( x_2, y_2 \) is also shown in the figure. The selection of this origin is based on the flux line solution of a four-pole machine. According to the figure selecting this point makes the flux guides about to conform the form of flux lines, see e.g. Fig. 2.1.

The parameters shown in Fig. 4.1 are
1. \( r_r \) - rotor radius
2. \( r_n \) - \( n^{th} \) arc radius
3. \( r_t \) - tangential rib width
4. \( r_{tr} \) - radial rib width
5. \( e_n \) - \( n^{th} \) flux barrier width
6. \( \tau \) - d-axis pole span
7. \( \tau_p \) - d-axis pole pitch
With parameters number 2 and 5 it is possible to investigate the effect of the flux barrier number as well as the location and width on the motor torque. In the transversally laminated type the rotor tangential and radial ribs (parameters 3 and 4) must be set non-zero for mechanical reasons. These parameters have a strong negative effect on the SynRM motor properties. The effect of the air-gap length is studied by changing the rotor radius (parameter 1). The bore of the stator is fixed. The effect of rotor q-axis cut-out is studied by changing the d-axis pole span value.

The nominal torque and the torque ripple are analysed using a rotating FEM model. The fundamental air-gap flux wave iron losses are investigated in chapter 2.4 using both magneto-dynamic and magneto-transient FEM models.

In order to achieve a low q-axis inductance the ribs of the transverse laminated type rotor should be as thin as mechanically possible. A rule of thumb for the punching of the lamination should be considered, which is that the rib width should not be less than the thickness of the laminations itself (Kamper 1996), which, in this case, is 0.5 mm.

4.2 Calculation

The FEM solvers used to analyse the different SynRM rotor designs are non-linear, two-dimensional, magneto-static solver in a Cartesian frame. The 48-stator-slot structure has four slots per pole and per phase. The slot pitch is thus 7.5 mechanical degrees. In the following investigations the effects of the different parameters on the torque production are studied with the aid of stepped calculations keeping a constant rms current 55 A and current angle of 60° electrical degrees. Because the angular step is 0.5 degrees totally 15 simulation steps must be calculated to achieve one slot pitch.

The only modification in the stator done is the addition of semi-magnetic slot wedges. The wedges are used to make the air-gap region permeance function more uniform since more flux is led under the stator slot opening than without wedges. The relative permeability of the magnetic slot wedge varies between $\mu_r = 2 \ldots 4$ and it has a high resistivity. Choosing the material is a question of optimisation between the power factor of the machine and the harmonic losses on the rotor surface. Increasing the permeability of the slot-wedge flattens the permeance function, but also increases the leakage flux. Figs. 4.2a and 4.2b show the effect of the slot-wedge in comparison with a normal slot opening.

![Figure 4.2. a) Flux plot in a stator slot section without a semi magnetic wedge. b) Flux plot in a stator slot section as a result of using a wedge (Pyrhönen 1993). The relative permeability of the slot wedge used in the figure is $\mu_r=10$.](image)
In the FEM models there are used several layers of elements across the air-gap. This enables the producing of similar triangle elements in the motor air-gap when the rotor is rotated which then increases the accuracy of calculating the motor torque. The number of elements is high because the air-gap of the SynRM is small. The number of element points in the layers is 180 per pole. It may thus be used the Maxwell stress tensor method to calculate the rotor electromagnetic torque. However, the calculation of the torque of a SynRM is remarkably more demanding than the calculation of the torque of an IM because of the very small air-gap length of the SynRM.

The angle from the middle of a stator tooth to the middle of the next stator tooth is 7.5°. In the original induction motor the stator winding is a full pitch single layer winding and each slot carries 13 conductors. The stator stack length is 205 mm, the air-gap diameter 215 mm and stator outer diameter 323 mm. For the initial SynRM prototype an air-gap length of 0.3 mm is used. The original induction motor has an air-gap of 0.8 mm.

The length of the 4-pole motor fundamental wave is 180° in geometrical degrees along the air-gap. By selecting the rotation angle to be 0.5° the arc of one pole pair (90°) is divided into 180 parts for the FEM-analysis. In the following sections (4.2.1) … (4.2.5) the tangential rib thickness is kept constant at 2 mm.

When the rotor rotates through the angle from the centre of one stator tooth to the centre of the next stator tooth, all the ripple effects of the torque caused by permeance variations may be studied. 15 different rotor positions are needed to calculate the required 7.5° distance. The torque wave for the 7.5° distance along the air-gap is constructed of the torque values calculated in the different rotor positions. The torque ripple is then calculated with respect to the mean value of the torque.

The calculation procedure is the following (Salo 2000):
1. Draw the stator geometry, build the stator mesh and set the boundary conditions.
2. Draw the rotor geometry, build the rotor mesh and set the boundary conditions.
3. Copy the rotor mesh and rotate it by 0.5°. Set the boundary conditions. Join the stator mesh to the rotor mesh. Repeat this 15 times. The result is 15 joined meshes, one for each rotor position.
4. Calculate the winding currents for each rotor position to maintain the dq-reference frame current as a constant
5. Set the FEM-problems for each rotor position.
6. Solve the FEM-problem for each rotor position with the 2D-FEM-calculation.
7. Calculate the torque of the solved FEM-problem for each rotor position.
8. Construct the torque wave from the torque values calculated in the different rotor positions, when the rotor rotates through the angle from the centre of one stator tooth to the centre of the next stator tooth.

4.2.1 Influence of the single flux barrier width

The effect of one single flux barrier width has been investigated by changing the width of one barrier as it is described in Fig. 4.3. At each step the flux guide width is increased by 2 mm.
The calculated torque values as a function of the rotor angle are shown in Fig. 4.4. The single barrier width is a parameter. It may generally be assumed that the torque production volume first increases from the initial value as a function of the barrier width. This is because the d-axis flux path reluctance is not remarkably increasing while the q-axis flux path reluctance increases rapidly. Of course, there exists a limit after which the torque starts to decrease while the flux barrier gets larger.

According to Fig. 4.4 the maximum and average level of the torque is increasing as a function of the barrier width but, on the other hand, also the torque ripple increases. Using this constant current vector model the air-gap flux varies a lot mainly because of the saturation in the stator teeth and tangential rib. This saturation and thus also the torque ripple effect increase when the flux barrier width grows, which then increases the irregularity in the flux distribution in the air-gap. This irregularity in the air-gap flux density is also a function of the current amplitude, so that at lower current vector amplitudes the saturation in the stator teeth decreases and thus also the torque ripple effect decreases.
The data shown in Fig 4.4 are analysed by calculating the average, maximum and minimum torque values from one slot pitch in Fig. 4.5.

![Graph showing torque values](image)

**Figure 4.5.** The maximum, minimum and average per unit torque values as a function of one flux barrier width. The torque of the original IM corresponds to unity p.u. torque. \( I = 55 \) A and the current vector angle is 60° in dq-co-ordinates.

While increasing the single flux barrier width the average torque first increases from the initial 2 mm value of the flux-barrier-width 0.25 p.u. up to 0.6 p.u. at 8 mm width value. After this, the slope of the torque value increase seems to “saturate”. As the flux barrier width increases, an increasing part of the flux is starting to go through the radial rib, which also starts to saturate as a function of the increasing flux. 7

According to Fig. 4.5 the absolute difference between the maximum and minimum torque, which means the torque ripple, seems to reach its lowest value at the flux barrier width 4 mm. While the q-axis inductance is decreasing as a function of the flux barrier width also the d-axis inductance is decreasing because of the increasing saturation of the rib. This flux plot situation has been illustrated in Fig. 4.6, where flux plots at barrier widths of 6 mm and of 18 mm are shown. The amount of flux lines through the barrier in the first mentioned case is 9 as the corresponding value in the latter case is only 6. The current angle vectors used in these figures are equal.

![Flux plots](image)

**Figure 4.6.** The flux plots with flux barrier widths of a) 6 mm and b) 18 mm without radial ribs.
According to Fig. 4.6, with the increase of the width of the flux barrier the flux is channelled to go through the tangential rib path or back to the stator using the shorter way. Even though the current vectors used in Fig. 4.6 are equal the place of the maximum flux seems to move towards the d-axis direction. The same effect can also be noticed from Fig. 4.3, where the place of the rotor angle producing the maximum torque is also moving as a function of the flux barrier width.

4.2.2 Influence of the flux barrier location

The effect of the location of the one single flux barrier with constant width is investigated by moving the barrier as shown in Fig 4.7. The moving is done so that at each step the d-axis flux guide width is increased by 2 mm so that the flux barrier moves 2 mm towards to the flux barrier arc radius initial point. The width of the flux barrier used here is set equal to the width given in previous chapter (6 mm). The flux barrier width used is also equal to the stator slot average width (6 mm). The average stator slot width is calculated using the slot bottom width and the largest slot width.

Figure 4.7. The direction of the flux barrier movement.

The relationships between the flux barrier movement and the calculation numbers are clarified in Fig. 4.8. The width of each line is 2 mm which makes that a flux barrier occupies the area of three arcs. The flux barrier number code 1-2-3 may be interpreted so that the barrier end is located between lines 1, 2, 3, 4 as it is shown in Fig. 4.8.
According to Fig. 4.8 the flux barrier is for the first time totally under the stator slot in this initial rotor position when using the barrier address numbers 2 – 6. Because there is a wide variation possibility of flux barrier numbers only some of the flux barrier torque values as a function of the rotor angle are shown in Fig. 4.9. The number code means that e.g. flux barrier 1-2-3 is located between the lines 1 and 4.

In Fig. 4.9 the most ripple-less torque over one slot pitch rotation is achieved using barrier number 4-5-6 and 8-9-10. None of the flux barrier positions produces a totally smooth torque. This means that in order to avoid a torque ripple when using several flux barriers the positions of the flux barriers should be selected so that their sum torque ripple is minimised.
The torque produced by a single-flux-barrier-rotor as a function of the barrier location is represented in Fig. 4.10. The data shown in Fig. 4.10 are analysed by calculating the average, maximum and minimum torque values over one slot pitch.

According to Fig. 4.10, in a one-flux-barrier-rotor-motor the torque ripple reaches its minimum at the line numbers 4 and 8. This is because the cogging torques in both ends of the flux barrier compensate each other in these flux barrier positions. The main flux variations are also minimised in these points.

As the flux barrier starts to move - between calculation points 1 to 9 - the average torque seems to increase from the value 0.5 p.u up to 0.6 p.u. After that point has been reached the respective torque value starts to decrease. It can be assumed that after calculation point 9 the d-axis flux does not remarkably increase and most of the q-axis flux meets the flux barrier. The flux plot situations in different calculation points are shown in Fig. 4.11. The share of the flux lines going through the barrier of the total flux lines is given in Table 4.1.

<table>
<thead>
<tr>
<th>Flux barrier location</th>
<th>Share of the flux lines going through the single flux barrier</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-2-3</td>
<td>9/17 (53 %)</td>
</tr>
<tr>
<td>6-7-8</td>
<td>9/18 (50 %)</td>
</tr>
<tr>
<td>15-16-17</td>
<td>7/18 (39 %)</td>
</tr>
<tr>
<td>28-29-30</td>
<td>2/18 (10 %)</td>
</tr>
</tbody>
</table>
According to the results from Fig. 4.12 and Table 4.1 the share of the flux lines going through the flux barrier is higher if the barrier is located closer to the d-axis. On the other hand, it may be assumed that if the barrier is located closer to the d-axis, the flux path in the d-axis centre is becoming narrower, which decreases the axis inductance.

### 4.2.3 Influence of the flux barrier number and width

When the width of all the individual flux barriers are kept equal the following torque values for the TLA-rotor as a function of the barrier amount number can be calculated. According to the results reached in the previous chapter the main d-axis flux guide is selected to reach the point number 3, which gives the main flux guide width of 10 mm. The other flux barriers and flux guides are selected suitably to produce the overall insulation ratio determined in the following. The width of the flux barriers is presented as a function of the insulation ratio $W_{\text{ins}}$. The results are shown in Fig. 4.12.

\[
W_{\text{tot}} = \frac{W_{\text{ins}}}{W_{\text{ins}} + W_{\text{iron}}},
\]

where

- $W_{\text{ins}}$ - the sum of the widths of the flux barrier layers
- $W_{\text{iron}}$ - the sum of the widths of iron layers (flux guides).

Figure 4.12. The torque as a function of the insulation ratio $W_{\text{ins}}$. The flux barrier number $n_r$ is a parameter. The flux guide widths and the flux barrier widths vary according to the insulation ratio. The main flux guide in the d-axis centre is kept constant 10 mm. The torque of the original IM corresponds to unity p.u. torque. $I = 55$ A and the current vector angle is $60^\circ$ in dq-co-ordinates.

According to the constant current vector analysis of Fig. 4.12 the maximum torque is achieved using an insulation ratio between 0.35 and 0.45. The number of insulation layers seems to have only a small effect on the maximum torque with the used current vector. In Fig. 4.12 the maximum torque is achieved by using four flux barriers.
According to Boldea (1991) values in the range 0.33 … 0.40 are recommended. Lipo (1991) suggests the use of value 0.33 in order to reduce the rotor iron loss. Soong (1995) used the value 0.5 to find a reasonable performance. All of these reference recommendations are for the axially laminated type rotor.

In the following the widths of the flux barriers are set equal to the width of the flux guide. Thus, selecting of four flux barriers sets the width of one barrier used in the model to be 6 mm \( W_{\text{tot}} = 0.40 \), which is, according to the result presented in chapter 4.3, a recommendable value for the flux barrier width.

Four separate flux barriers are selected in the final rotor. The torque ripple then consists of the sum of the cogging torques of these four barriers.

In the following the torque ripple using different flux barrier combinations is investigated. The number of the flux barriers inside the rotor is limited to three. The fourth flux barrier is located on the surface of the rotor forming the q-axis cut-out. The results of the torque ripple as a function of the flux barrier combinations is shown in Fig. 4.13. The flux barrier combination marks used in the figure may be presented as:

Combination 1: flux barriers on arcs 1-2-3, 9-10-11, 17-18-19
Combination \( n \): flux barriers on arcs \( n-(n + 1)-(n + 2), (n + 8)-(n + 9)-(n + 10), (n + 2 \cdot 8)-(n + 2 \cdot 8 + 1)-(n + 2 \cdot 8 + 2) \).

![Figure 4.13. The ripple of the different flux barrier combinations. The torque of the original IM corresponds to unity p.u. torque. \( I = 55 \) A and the current vector angle is 60° in dq-co-ordinates.](image)

Using the flux barrier arcs 3-4-5, 11-12-13 and 19-20-21 the total flux ripple remains quite low as it is shown in Fig. 4.14. The lowest torque ripple value is achieved using the combination 7 which are the arcs 7-8-9, 15-16-17 and 23-24-25, but in this case the fourth flux barrier would be located too close to the rotor surface. Thus, selection number 3 is chosen. The torque ripple curves of the selected four flux barriers are illustrated in Fig. 4.14.
Figure 4.14. The torque curves of four separate flux barriers and the sum curve of these. The torque of the original IM corresponds to unity p.u. torque. $I = 55$ A and the current vector angle is 60° in dq-coordinates.

According to Fig. 4.14 the torque ripple of the selected flux barrier combination is 21.6%. Skewing the rotor slightly may reduce this torque ripple. The calculation results of the effect of skewing on the torque ripple are given in Fig. 4.15 as a function of the skewing angle.

Figure 4.15. The torque ripple of the selected flux barrier combination as a function of the skewing angle. The torque of the original IM corresponds to unity p.u. torque. $I = 55$ A and the current vector angle is 60° in dq-coordinates.

According to Fig. 4.15 the minimum torque ripple is achieved at a skewing angle of 4.0 degrees, which is practically the value of one stator slot pitch ($\frac{2\pi R}{Qs I} = 3.9°$). The torque
ripple using a rotor skewed by four degrees is 2.8 % with the current vector used. For a 1500 rpm machine this is a reasonably good value and may compete well with the IM torque ripple.

Fig. 4.16 shows the value for the corresponding IM torque ripple at nominal load. The IM has no rotor skewing. Evaluation is done using the magneto-transient FEM model.

Figure 4.16. The torque ripple of the IM at nominal load (194 Nm). The ripple is about 18 % of the nominal torque.

4.2.4 Influence of the pole span on the pole pitch ratio $\tau / \tau_p$

The torque-current-ratio increases by increasing the air-gap width in the $q$-axis direction. With this method the $d$-axis flux path reluctance is kept quite stable as the $q$-axis reluctance increases. The limit for the pole-span-to-pole-pitch-ratio is the value where the ratio between the $d$- and $q$-axes inductances starts to decrease. Fig. 4.17 shows the rotor structure with a pole span.

Figure 4.17. The rotor structure with a pole span caused by the $q$-axis cut-out.
The torque values as a function of the pole span angle are shown in Fig. 4.18.

Figure 4.18. The behaviour of the torque as a function of the pole span angle. The torque of the original IM corresponds to unity p.u. torque. $I = 55$ A and the current vector angle is 60º in dq-co-ordinates.

According to Fig. 4.18 the maximum torque-current-ratio is achieved using a cut-out producing at the pole span ratio 0.67. This gives a result which is quite similar to the corresponding value introduced by Boldea (1996). The pole span increases the nominal torque by about 10 %. The respective torque ripple values are shown in Fig. 4.19.

Figure 4.19. The variations of the torque ripple as a function of the pole span angle without skewing. In this figure the value of the torque ripple is slightly differing from the corresponding value in Fig. 4.15 because in the case here the torque ripple is calculated using three inner flux barriers at the same time.

The minimum torque ripple is achieved at a higher pole span angle than the corresponding maximum average torque. Because the difference of the pole span angles between these two extreme values is quite small it is recommended to select a lower torque ripple value. The fact is that a higher pole span angle brings a wider flux guide and thus more rigidity in the rotor structure.
4.2.5 Influence of the air-gap length

The air-gap length affects the direct axis inductance, the stator leakage flux inductance, the iron losses and the mechanical loss values. Decreasing the air-gap length increases effectively the d-axis inductance. The air-gap length effect on the q-axis inductance is minor because of its relatively large reluctance consisting of the rotor inner flux barriers. Fig. 4.20 illustrates the effect of the air-gap length on the axis self-inductance. The rotor type shown in this figure is axially laminated and the lay-out of the rotor is the same as that described in Fig. 2.9a. The estimation of the d-axis current values producing a constant d-axis flux are marked by means of circles.

![Figure 4.20. The axially laminated type SynRM d- and q-axis inductances as a function of the currents of the axes. The air-gap length is a parameter. The white circles indicate the current values corresponding to the constant d-axis flux.](image)

4.2.6 Influence of the mechanical strutting

In the transversally type SynRM rotor structure the mechanically weakest items are the tangential and the radial ribs. The general rule is that wider flux guides make the rotor mechanically stronger and more braced and thus the number and the width of the flux barriers are mechanically limited (Kamper 1996). The widths of the tangential and radial ribs must increase as a function of the motor size because the mechanical forces inside the motor increase.

The q-axis flux path reluctance of the transversally laminated reluctance motor consist of the air-gap, the individual radial ribs, the tangential rib and the sum of the equal flux barrier reluctances. In the following the effect of different rib widths on the torque production is investigated.

**Radial rib**

The model has a 2 mm wide tangential rib and a cut-out. The effect of the radial rib width on the average torque and to torque ripple is described in Table 4.2.
Table 4.2. The average and torque ripple as a function of the width of the radial rib. The tangential rib is constant 2 mm.

<table>
<thead>
<tr>
<th>Width of radial rib [mm]</th>
<th>Average torque p.u.</th>
<th>Torque ripple [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1.02</td>
<td>17.1</td>
</tr>
<tr>
<td>1</td>
<td>1.01</td>
<td>17.9</td>
</tr>
<tr>
<td>2</td>
<td>0.99</td>
<td>19.3</td>
</tr>
<tr>
<td>4</td>
<td>0.97</td>
<td>21.9</td>
</tr>
</tbody>
</table>

The increase of the radial rib increases also the q-axis inductance and decreases thus the average torque. The increase of the q-axis inductance decrease also the saliency ratio and increases thus also the effect of the cross saturation as it is mentioned in chapter 4.1.

The torque curves of different tangential rib widths as a function of the rotation angle are represented in Fig. 4.21.

![Torque Curves](image)

Figure 4.21. The torque curves as a function of the rotation angle. The radial rib width is a parameter and the tangential rib width is constant 2 mm. The torque of the original IM corresponds to unity p.u. torque. \( I = 55 \text{ A} \) and the current vector angle is 60º in dq-co-ordinates. The rotor has four flux barriers and no skewing.

The radial rib width is selected to be 1 mm. This also fulfils the general rule mentioned above, that the width must by bigger than the lamination thickness.

**Tangential rib**

Table 4.3 gives the results for the rotor tangential rib as a function of the rib \( r_{tr} \) width. In Fig. 2.11 the lay-out of the rotor model is introduced. The rotor is an ALA-type when the tangential rib width is 0 mm. The model has no radial rib and has six flux barriers.

Table 4.3. The average and torque ripple as a function of tangential rib width. The model used has no radial rib as it is shown in Fig. 2.11a.

<table>
<thead>
<tr>
<th>Width of tangential rib [mm]</th>
<th>Average torque p.u.</th>
<th>Torque ripple [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1.02</td>
<td>42.6</td>
</tr>
<tr>
<td>1</td>
<td>1.00</td>
<td>30.2</td>
</tr>
<tr>
<td>2</td>
<td>0.96</td>
<td>24.7</td>
</tr>
<tr>
<td>3</td>
<td>0.93</td>
<td>17.8</td>
</tr>
</tbody>
</table>
According to Table 4.3 the average torque decreases as a function of the tangential rib thickness but the torque ripple reduces. It is shown in Fig. 2.4 that the reduce of the air-gap harmonics contents causes a decreasing of the torque ripple.

The torque curves of different tangential rib widths as a function of the rotation angle are represented in Fig. 4.22.

Figure 4.22. The torque curves as a function of the rotation angle. The tangential rib width is a parameter and the model does not contain a radial rib. The torque of the original IM corresponds to unity p.u. torque. \( I = 55 \text{ A}, \) and the current vector angle is 60º in dq-co-ordinates. The rotor has six flux barriers and no skewing.

4.3 Conclusion

According to the calculations the maximum torque was achieved in this case using four flux barriers. The minimum torque ripple with these four flux barriers was achieved using an insulation ratio of 0.4, which was obtained using a flux barrier width of 6 mm and a flux guide width of 10 mm. The fourth flux barrier consists of the rotor surface cut-out in the q-axis area. The width of the cut-out was selected compromising the maximum torque productivity and the torque ripple. The widths of the radial ribs and the tangential rib were chosen to be three times as large (1.5 mm) as the thickness of the iron steel sheet used in the machine (0.5 mm) mainly to maximise the mechanical ruggedness. The selected wide ribs reduce the torque ripple but decrease also the torque productivity.
5 COMPARISON OF THE INDUCTION MOTOR AND SYNCHRONOUS RELUCTANCE MOTOR BASED ON THE TEST RESULTS

This chapter introduces the results of the laboratory measurements. Chapters 2 and 4 already discussed some of the results by comparing them with the calculated values. Because there is no established standard for the measuring of an inverter fed SynRM or induction motor, the IEEE 112 B method for sinusoidally fed poly-phase induction motor is applied in this work. This U.S. standard is recommended for efficiency measurements in the power range of 0.75... 190 kW. The calculation forms for input-output tests of motors with segregation of losses and smoothing of stray load losses are shown in Appendix C. The forms follow the IEEE 112 Std 112-1996 method B. The iron and mechanical losses of both the induction motors and SynRMs are measured using sinusoidal voltage. This is explained more detailed in the following.

A DC-machine is used to load the motor. The motor torque is measured by a torque transducer. The DC-machine has an electronic control that sets the load torque and holds it steady while readings are being taken.

The measuring consists of several steps, the rated load temperature test, the load test and no-load test. The apparent total loss (input minus output) is segregated into various components with stray-load loss defined as the difference between the apparent total loss and the sum of the conventional losses. The value of the stray-load loss thus determined is plotted vs. the torque squared, and a linear regression is used to reduce the effect of random errors in the test measurements. The smoothed stray-load loss data are used to calculate the final value of the total loss and the efficiency.

5.1 Description of the test set-up

The SynRM rotor structure is designed and manufactured based on the standard 30 kW four-pole induction machine stator. Both the induction and SynR-motor are supplied by a prototype sensorless direct torque controlled (DTC) variable speed drive.

The SynRM rotor is transverse laminated. To eliminate cogging the rotor is skewed by one stator slot pitch. The designed rotor photo is shown in appendix B.

All motor configurations are tested in the laboratory in order to evaluate the effects of the changes made, Fig. 5.1. The length of the air-gap and the effect of stator slot-wedges are the variables that are investigated.
The identification numbers and the constructions of the motors are given in Table 5.1.

Table 5.1. The test motors

<table>
<thead>
<tr>
<th>Motor no.</th>
<th>Features</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Sinusoidally fed standard high efficiency induction motor, air-gap 0.8 mm</td>
</tr>
<tr>
<td>2</td>
<td>Inverter fed standard induction motor, air-gap 0.8 mm</td>
</tr>
<tr>
<td>3</td>
<td>Inverter fed standard induction motor + semi magnetic slot-wedge, air-gap 0.8 mm</td>
</tr>
<tr>
<td>4</td>
<td>SynRM, air-gap length 0.3 mm</td>
</tr>
<tr>
<td>5</td>
<td>SynRM, air-gap length 0.3 mm + semi magnetic slot-wedge</td>
</tr>
<tr>
<td>6</td>
<td>SynRM, air-gap length 0.4 mm + semi magnetic slot-wedge</td>
</tr>
<tr>
<td>7</td>
<td>SynRM, air-gap length 0.4 mm + semi magnetic slot-wedge + damper windings</td>
</tr>
</tbody>
</table>

Motor number 1 is the original standard induction motor, the stator construction of which is used also for the SynRM construction. The effect of the variable speed drive on the efficiency of the standard induction motor is investigated in motor drive system 2. To run the motor it is used an ABB ACS600 frequency converter, which applies the direct torque control (DTC) system.

The effect of semi magnetic slot wedges is investigated in both cases, the IM (motor 3) and the SynRM (motors 5 and 6).

The effect of damper windings in the SynRM is investigated with motor 7.
To achieve power factors that are high enough for the SynRM it must be selected an air-gap length that is much smaller than the corresponding value of the IM. Thus the effect of the air-gap length has been investigated only for SynRMs (motors 5 and 6).

**Instrumentation accuracy**

The IEEE 1112 instrumentation tolerances, the instruments used and their accuracy values are summarised in Table 5.2. The GPI-bus transfers and the computer handles the data in digital mode and there the error should be negligible.

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Unit</th>
<th>IEEE 112 (Method B)</th>
<th>Instrument</th>
<th>Accuracy</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>W</td>
<td>± 0.2 %</td>
<td>Power Analyser Norma 6100</td>
<td>± (0,05 + 0,005) %</td>
</tr>
<tr>
<td>Current</td>
<td>I</td>
<td>± 0.2 %</td>
<td>Power Analyser Norma 6100</td>
<td>± (0,05 + 0,005) %</td>
</tr>
<tr>
<td>Voltage</td>
<td>V</td>
<td>± 0.2 %</td>
<td>Power Analyser Norma 6100</td>
<td>± (0,05 + 0,005) %</td>
</tr>
<tr>
<td>Frequency</td>
<td>Hz</td>
<td>± 0.1% * f₀</td>
<td>Power Analyser Norma 6100</td>
<td>± 0,01 %</td>
</tr>
<tr>
<td>Rotational speed</td>
<td>r/min</td>
<td>± 1 %</td>
<td>Vibro-Meter TM 214</td>
<td>≤± 0,1 % of the measuring range</td>
</tr>
<tr>
<td>Torque</td>
<td>Nm</td>
<td>± 0.2 %</td>
<td>TM 285 (500 Nm)</td>
<td>≤± 0,1 % of the measuring range</td>
</tr>
<tr>
<td>Resistance</td>
<td>Ω</td>
<td>± 0.2 %</td>
<td>Fluke 8840</td>
<td>± 0,0010 % + 0,0003</td>
</tr>
<tr>
<td>Temperature</td>
<td>°C</td>
<td>± 1</td>
<td>DTM 10 Wallac</td>
<td>± 0,05</td>
</tr>
</tbody>
</table>

According to Table 5.2 the accuracy of the instruments used are acceptable.

### 5.2 No-load test

The test is performed using sinusoidal voltages to evaluate the iron loss at no-load and the mechanical loss at rated synchronous speed of the motors. With this method of evaluating the iron loss values the harmonic losses caused by the inverter supply are not considered. These losses are assumed to be included in the load dependent iron losses - to the additional losses. Thus the no-load loss values for test motor 1 with sinusoidal supply and test motor 2 with converter supply are equal.

Because the direct on-line starting torque of the SynRM without rotor cage is very low the motor should be forced to run at rated synchronous speed before connection to sinusoidal voltage supply can be done. The synchronous speed is achieved with an external motor running on the same shaft. To carry out the measurements it is used a DC-motor supplied by a converter as an external working machine. After the SynRM is connected to sinusoidal voltage the DC-motor is decoupled using an external coupler.

It is possible to run a SynRM at synchronous speed with sinusoidal voltage due to the high saliency ratio of the motor. During the test runs it was noticed that it is easier to synchronize the SynRM to the sinusoidal supply at higher rotating speeds. This may be caused by the increased
energy carrying in the inertia mass and by the smaller relative slip between the rotating magnetic field and the rotor.

Because the ambient temperature in the laboratory is kept constant the no-load tests are carried out in each case with a cold motor. It is assumed that applying this method of measuring the accuracy needed for this test will be achieved. For this reason, it is unnecessary to do any temperature or input power stabilisation before the no-load measurement readings can be started. The reading of the input power is equal to the sum of the no-load motor losses. Subtracting the stator ohmic loss (at the temperature of this test) from the input gives the sum of the friction, windage, and core losses. Fig. 5.2 shows the measured values for the induction motor, Fig. 5.2(a), and for the SynRMs, Fig. 5.2(b), at no-load.

\[
P_{\text{Motor 1}} = 0.0011*U^2 + 0.59*U + 259.29
\]

\[
P_{\text{Motor 3}} = 0.003*U^2 - 0.0007*U + 181.75
\]

\[
P_{\text{Motor 4}} = 0.0054*U^2 - 1.95*U + 1087.1
\]

\[
P_{\text{Motor 5}} = 0.0021*U^2 + 1.36*U - 0.80
\]

\[
P_{\text{Motor 6}} = 0.0025*U^2 + 0.93*U + 35.94
\]

Figure 5.2. No-load test measured values using sinusoidal voltage for a) the induction motor and b) the SynRMs.
The iron losses can be calculated from the equations shown in Figs. 5.2(a) and 5.2(b) after the mechanical losses have been determined. For the evaluation of the iron losses a 400 V supply voltage is used.

The core loss is separated from the friction and windage losses by reading the voltage, the current, and the power input at the rated frequency and at voltage ratings from the rated voltage down to a point where further voltage reduction starts to increase the current. The power input minus the stator ohmic loss is plotted vs. voltage, and the curves obtained are extrapolated to zero voltage. The interception with the zero voltage axis is the sum of the friction and windage losses. In this test the mechanical loss is determined more accurately by plotting the input power minus the stator ohmic loss against the voltage squared for lower voltage range, figure 5.3.

\[
P_{\text{Motor 1}} = 0.0042*U^2 + 270.23
\]

\[
P_{\text{Motor 3}} = 0.003*U^2 + 182.38
\]

\[
P_{\text{Motor 6}} = 0.0045*U^2 + 189.11
\]

\[
P_{\text{Motor 5}} = 0.0037*U^2 + 201.81
\]

\[
P_{\text{Motor 4}} = 0.0034*U^2 + 300.15
\]

Figure 5.3. Motor core and mechanical loss curves against the voltage squared for a) the induction motors and b) the SynRMs.
The results of the no-load tests are given in Table 5.3. The values are evaluated utilising the curve fittings of figures 5.2(a) … 5.3(b).

<table>
<thead>
<tr>
<th>Measured values</th>
<th>Motor 1</th>
<th>Motor 2</th>
<th>Motor 3</th>
<th>Motor 4</th>
<th>Motor 5</th>
<th>Motor 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator ohmic loss [ W ]</td>
<td>88</td>
<td>88</td>
<td>101</td>
<td>24</td>
<td>22</td>
<td>41</td>
</tr>
<tr>
<td>Stator and rotor iron loss [ W ]</td>
<td>422</td>
<td>422</td>
<td>461</td>
<td>870</td>
<td>605</td>
<td>690</td>
</tr>
<tr>
<td>Mechanical loss [ W ]</td>
<td>270</td>
<td>270</td>
<td>182</td>
<td>300</td>
<td>202</td>
<td>189</td>
</tr>
<tr>
<td>Total no-load loss [ W ]</td>
<td>780</td>
<td>780</td>
<td>744</td>
<td>1194</td>
<td>829</td>
<td>920</td>
</tr>
<tr>
<td>No-load current [ A ]</td>
<td>19.8</td>
<td>19.8</td>
<td>20.3</td>
<td>10.9</td>
<td>10.5</td>
<td>12.2</td>
</tr>
</tbody>
</table>

Table 5.3 proves that the total no-load losses of the tested SynRMs are higher than those of the IMs. This is mainly caused by the higher no-load iron losses. The smallest no-load losses for the SynRMs are achieved with the air-gap length 0.3 mm using semi-magnetic slot wedges.

According to Fig 5.3(b) in the SynRM constructions the inserting of the semi-magnetic slot wedges and also the increase of the air-gap length decrease the mechanical losses. A smooth and bigger air-gap decreases the friction of air.

In the case of the induction motor, the semi-magnetic slot wedges (motor 3) decrease the motor power factor and increase thus the stator current. With a higher no-load current the no-load iron losses seem to increase. In the induction motor, which has a relatively big air-gap length, the effect of the wedges on the rotor surface iron loss is quite negligible. The mechanical losses seem to differ in a considerable way. However, the big differences may also be explained by slight changes in the mechanics due to several rotor changes.

The SynRM rotor without damper windings is about 20 % lighter than the IM rotor because it does not contain a rotor cage and, therefore, the mass of iron is smaller. A lighter rotor has the benefit that it decreases the mechanical losses of the motor. When comparing the motors having a semi magnetic slot-wedge it may be noticed that, even though the rotor surface of the IM is smoother than that of the SynRM and even though the SynRM has a smaller air-gap length the mechanical losses are quite equal. The same observation may be done regarding the motors having no semi magnetic slot-wedge. The cut-outs (Fig. 4) on the SynRM rotor surface may increase the windage losses because the air-gap length is not smooth.

In the SynRM the effect of the slot wedges on the no-load iron losses is remarkable because the air-gap of the SynRM is narrower compared to that of the induction motor. Increasing the air-gap length of the SynRM increases also the no-load current. During the test a couple of stator slot wedges are lost which may be the reason for the increasing of the iron losses. On the other hand, the load dependent iron losses and also the mechanical losses decrease.

5.3 Rated load temperature test

For the performing of the load tests the motors are coupled to a DC-machine that is operated at rated current until it was thermally stable. Because the temperature of the SynRM prototype DTC controller behaves unstably at high load the measured values do not reach the 125 % value in any of the measured cases.

The measured motors have quite high efficiencies and, therefore, it is important that the actual temperature of operation is monitored and used for evaluation. The reduced losses of the motor
reduce the temperature rise in the motor. Otherwise, the insulation class temperature values for the windings should be used for the determining. High efficiency motors generally operate beneath the permitted temperature rise.

The stator ohmic loss $P_{Cu,S}$ for three-phase machines is equal to equation

$$P_{Cu,S} = 1.5 I^2 R_s$$

(5.1)

where $I$ is the measured rms current per line terminal at a specified load and $R_s$ is the DC resistance between any two line terminals corrected to the specified temperature. In this test the measurement of the temperature is done after a rapid shutdown. The stator resistance behaviour is measured as a function of time. These readings and a fitted curve (with a mathematical formula) are been plotted in figure 5.4.

![Figure 5.4. The resistance values as a function of time measured after a heating test for the sinusoidally supplied IM.](image)

This stator resistance measurement is unique because in all measured motors it is used the same stator which is warmed in a same time and using the same current. The only temperature correction that should be done is caused by different ambient temperatures.

### 5.4 Tests under load

The motor is loaded by a DC-machine regenerating drive. The load tests are carried out to determine the efficiency, the fundamental power factor, the speed, the current, and the temperature rise of the motor. The measured values of the power factors for the induction motors and SynRMs as a function of power are shown in Fig. 5.5.
According to Fig. 5.5, at a lower load range the power factors of the SynRMs are higher compared to the values of the induction motors. The benefit of the small d-axis air-gap is emphasized in this area.

According to the results the power factors of variable frequency supplied induction motors are higher than the power factor of the corresponding sinusoidally supplied induction motor below the rated power area. The semi-magnetic slot-wedges decrease the harmonic losses but this effect remains quite minor in the range of higher power. Above this power the fundamental power factors of VF driven motors seem to fall.

Compared to the induction motor with a frequency converter supply the power factors of the SynRMs behave quite similarly. The value starts to decrease at a certain load point. For the SynRMs this load point appears at a lower power than for the induction motor drives. According to figure 5.5, motor 5 with semi magnetic slot-wedge has a higher power factor than the respective motor 4 without semi magnetic slot-wedge. It is assumed that the power factor increases as a result of the increased saliency ratio. The d-axis inductance increases more than the q-axis inductance. Because of the rotor cut-out the leakage flux inductance in the q-axis is somewhat smaller than in the d-axis.

The increased air-gap length in motor 6 decreases the power factor because of the decreased d-axis inductance and thus decreased saliency ratio. Because of the decreased d-axis inductance more d-axis current is needed, which then reduces the current angle $\kappa$.

The measured differences of efficiencies between the induction motor with FC or sinusoidal voltage supply are shown in Fig. 5.6.
As it can be noticed from Fig. 5.6 the efficiencies of the motors 2 and 3 with converter supply are decreasing compared to the efficiency of the motor with sinusoidal supply, especially at full load. This efficiency decrease of the converter fed motors is caused by two phenomena. The harmonic losses in the motor increase remarkably. Also the copper losses increase since the motor is running in a slight field weakening at its nominal speed. This field weakening is typical for DTC inverters that maintain a certain voltage reserve for dynamic changes in the torque. Converter fed motors reach their highest efficiency values at partial load values that are lower than those of the sinusoidally supplied motor. This is caused by the load dependent losses that are more considerable in the FC fed motors than in the sinusoidally fed motor. The efficiency maximum partial load point is dependent on the share of the load independent losses $P_I$ (core loss and mechanical loss) and load-dependent losses $P_L$ (stator and rotor $I^2R$ loss and stray load loss) (Auinger 1987). The efficiency reaches its maximum value at the partial load point $P^*$

$$P^* = \sqrt{\frac{P_I}{P_L}P_N}.$$  \hspace{1cm} (5.2)

The difference between the standard motor with and without semi-magnetic slot-wedges in converter supply is interesting to note. The semi-magnetic slot-wedges affect both the no-load and the load dependent losses. The stator semi-magnetic slot-wedges smoothen the stator inner surface, and decrease thus the windage losses. They also smoothen the flux density distribution on the rotor surface, as it is discussed earlier in chapter 4.2, and reduce thus the permeance harmonics generated rotor surface losses.
The measured differences between the SynRMs with FC supply are shown in Fig. 5.7. The effect of semi-magnetic slot wedges (motors 5 and 6) and the increase of the air-gap length (motor 6) are illustrated.

Fig. 5.7 shows that the efficiencies of motors 5 and 6 are quite equal. The interesting aspect of the difference between the motor without semi-magnetic slot-wedges (motor 4) and the motors with semi-magnetic slot-wedges (motors 5 and 6) is analysed more detailed below. The increase of the air-gap length moves the efficiency maximum to a higher partial load.

![Efficiency vs. Motor Power](image)

Figure 5.7. The efficiencies of the SynRM prototypes as a function of the motor power.

**Stray-load loss**

The stray-load loss will be separately calculated for each load point by subtracting from the apparent total loss the stator ohmic loss at rated temperature of the test, the core loss, the friction and windage loss, and the rotor ohmic loss corresponding to the measured values of the slip. Calculation will be done for each of the approximately equally-spaced load points.

Smooth stray-load loss data are obtained by using a linear regression analysis based on the expression of the stray-load loss as a function of the square of the load torque

\[ W_{LL} = AT^2 + C, \]  

(5.3)

where \( W_{LL} \) is the stray-load loss as plotted vs. torque squared, \( T \) is the torque, \( A \) is the slope and \( C \) is the intercept with the zero torque line.

If the slope is negative or if the correlation factor \( r \) is less than 0.9, the worst point should be deleted and then the regression should be repeated. If this increases \( r \) to 0.9 or larger, the second regression should be used; if not, or if the slope is still negative, the test is unsatisfactory. The linear regression data for SynR-motors are shown in Fig. 5.8.
Motor 5 = 0.019*\(T^2\) + 229.43
\(r^2 = 0.82\)

Motor 6 = 0.012*\(T^2\) - 87.60
\(r^2 = 0.91\)

Motor 4 = 0.018*\(T^2\) - 628.38
\(r^2 = 0.90\)

The stray-load losses vs. torque squared of the measured motor are shown in Fig. 5.9. Fig. 5.9 illustrates the measured effect of the semi-magnetic slot-wedges and of the air-gap length on the stray load additional losses.

Figure 5.8. Plot of Stray-Load loss vs. Torque Squared for 30 kW SynRM.

Figure 5.9. The corrected stray-load losses of the different motors as a function of the motor output power.
Table 5.4 gives an abstract of the components, origin, type and locations of the load dependent additional losses.

Table 5.4. Leakage flux (load) stray losses (Schwarz 1964)

<table>
<thead>
<tr>
<th>Component</th>
<th>Origin</th>
<th>Type and location</th>
</tr>
</thead>
<tbody>
<tr>
<td>Surface losses</td>
<td>Gap leakage (harmonic) flux</td>
<td>Stator and rotor core losses</td>
</tr>
<tr>
<td>Tooth-pulsation losses</td>
<td>Gap leakage (harmonic) flux</td>
<td>Stator and rotor core losses</td>
</tr>
<tr>
<td>Tooth-pulsation, squirrel-cage circulating</td>
<td>Gap leakage (harmonic) flux</td>
<td>Rotor ohmic losses</td>
</tr>
<tr>
<td>current losses</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Stator-harmonic, circulating current losses</td>
<td>Gap leakage (harmonic) flux</td>
<td>Rotor ohmic losses</td>
</tr>
<tr>
<td>Stator-slot eddy-current losses</td>
<td>Slot leakage flux</td>
<td>Stator ohmic losses</td>
</tr>
<tr>
<td>Rotor-slot eddy-current losses</td>
<td>Slot leakage flux</td>
<td>Abnormal rotor $I^2R$ loss at high slip</td>
</tr>
<tr>
<td>Stator-overhang eddy-current losses</td>
<td>Overhang leakage flux</td>
<td>Stator core loss</td>
</tr>
<tr>
<td>Rotor-overhang eddy-current losses</td>
<td>Overhang leakage flux</td>
<td>Abnormal rotor core loss at high slip</td>
</tr>
</tbody>
</table>

According to Fig. 5.9 the increase of the air-gap length in the SynRM (motor 6) decreases the additional losses. If the air-gap length is kept constant (motors 4 and 5) the effect of the semi-magnetic slot-wedges on the additional losses is minor. SynRM additional losses are mainly caused by the gap leakage harmonic flux. The effect of the slot harmonics on the SynRM additional losses can be ignored because there are no current conductors in the rotor.

Because the stators of these motors are equal the air-gap flux harmonic contents of the induction motor are smaller compared to the respective SynRM harmonic contents, mainly because of the wider air-gap length. The rotor surface of both motors are smooth and the air-gap length is constant, except in the cut-out areas of the SynRM q-axes.

Table 5.5 shows the results of the measured different losses based on the load test. The values are evaluated using the measuring report of IEEE 112 method B. The measuring table of motor 6 is given in appendix C.
Table 5.5. Estimated and IEEE temperature corrected motor data at rated power 30 kW

<table>
<thead>
<tr>
<th></th>
<th>Motor 1</th>
<th>Motor 2</th>
<th>Motor 3</th>
<th>Motor 4</th>
<th>Motor 5</th>
<th>Motor 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Efficiency [ % ]</td>
<td>92.4</td>
<td>91.9</td>
<td>92.1</td>
<td>90.5</td>
<td>91.6</td>
<td>92.0</td>
</tr>
<tr>
<td>Power factor</td>
<td>0.80</td>
<td>0.78</td>
<td>0.79</td>
<td>0.77</td>
<td>0.82</td>
<td>0.72</td>
</tr>
<tr>
<td>Nominal Current [A]</td>
<td>55</td>
<td>61</td>
<td>61</td>
<td>65</td>
<td>65</td>
<td>66</td>
</tr>
<tr>
<td>Iron loss [ W ]</td>
<td>422</td>
<td>422</td>
<td>461</td>
<td>870</td>
<td>605</td>
<td>690</td>
</tr>
<tr>
<td>Mechanical loss [ W ]</td>
<td>270</td>
<td>270</td>
<td>182</td>
<td>300</td>
<td>202</td>
<td>189</td>
</tr>
<tr>
<td>Stator ohmic loss [W]</td>
<td>1051</td>
<td>1192</td>
<td>1210</td>
<td>1334</td>
<td>1312</td>
<td>1373</td>
</tr>
<tr>
<td>Rotor ohmic loss [W]</td>
<td>650</td>
<td>552</td>
<td>584</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Additional loss [W]</td>
<td>74</td>
<td>213</td>
<td>145</td>
<td>635</td>
<td>633</td>
<td>368</td>
</tr>
<tr>
<td>Load dependent losses [W]</td>
<td>1775</td>
<td>1957</td>
<td>1939</td>
<td>1969</td>
<td>1945</td>
<td>1740</td>
</tr>
<tr>
<td>Total loss [W]</td>
<td>2467</td>
<td>2649</td>
<td>2581</td>
<td>3140</td>
<td>2753</td>
<td>2619</td>
</tr>
</tbody>
</table>

According to Table 5.5 the total losses of the best SynRM construction give values equal to the losses of the respective induction motor construction in the converter supply. The largest losses are measured with the SynRM construction with the smallest air-gap length and without semi-magnetic slot wedges to dampen the air gap spatial harmonics. With the slot wedges the total losses of the frequency converter supplied IM can be slightly reduced compared to the losses of the frequency converter supplied IM without slot wedges. Using slot wedges the iron loss on no-load are higher and the power factor on load remains higher. The slot wedges seem to reduce the windage losses in every case. However, the mechanical losses are easily affected by the bearing arrangement of the motor. The same motor frame was used in every measurement. The bearings were changed when necessary. This easily causes tens of watts of variation in the mechanical losses in the motors.

As it can be noticed from Table 5.5, there is no significant difference between the total losses of the best SynRM (motor 6) and the losses of the standard induction motor in converter supply. This is mainly the result of factors that are caused by the absence of the rotor ohmic losses. The smallest load losses are achieved with the air-gap length 0.4 mm and with semi-magnetic slot wedges.

According to Table 5.5, when the motor air-gap length is kept the same the semi-magnetic slot wedges decrease the mechanical and load depended iron losses of both the IM and SynRM. In the case of the SynRM the semi-magnetic slot wedges reduce the no-load iron losses but in the case of the IM the no-load iron losses increase.

In a variable speed drive the standard 30 kW induction motor efficiency without semi-magnetic slot-wedges (motor 2) is quite equal to the efficiency of the best SynRM with slot wedges (motor 6). The SynRM without slot-wedges has an efficiency that is 1.7 %-units lower at rated power.

5.5 The SynRM with damper winding

The starting and pull-out performances of the SynRM with damper winding construction introduced in Fig. 1.8 are shown in Figs. 5.10. In the following the motor is called motor 7.
Figure 5.10. SynRM with damper windings a) measured starting torque at no-load b) the pull-out torque test at lower voltage (100 V).

Fig. 5.10a shows that the starting time is quite long and the torque contains oscillation. The poor starting performance is an indication of the fact that the rotor cage and rotor magnetic path are not at all optimal for an asynchronous start. According to Fig. 5.11 the rotor seems to behave similarly compared to induction motors. The flux is not penetrating the rotor to a large degree at high slip.

Figure 5.11 SynRM with damper windings direct grid fed start-up flux penetration, blocked rotor.

The results of the SynRM with damper winding no-load test are illustrated in Figs. 5.12 … 5.13.
The network supply load test for the SynRM without damper windings was not done since it is impossible to start the motor directly to the network. Also, without dampening the motor does not maintain synchronism. Thus, the evaluation of the damper windings effect in direct supply can not be discussed here. The differences between the FC fed SynRM without damper windings and the network supplied SynRM with damper winding are shown in Figs. 5.14a and 5.14b. Both motors have semi-magnetic slot wedges and the same air-gap length.
According to Fig. 5.14 the SynRM with damper winding has a higher or equivalent power factor and corresponding efficiency values compared to the SynRM without damper winding and fed by the DTC FC in the load area below 50% of the nominal load. At higher load the power factor of motor 6 is lower due to the flux reduction caused by the voltage drop in the FC. Thus, to produce the same torque higher stator current is needed. This increases mainly the q-axis current, which reduces the power factor.
Because of the higher power factor the stator ohmic losses are lower in motor 7 but there occurs load dependent additional losses in the damper winding as it is shown in Fig. 5.15.

![Figure 5.15. The corrected stray-load loss as a function of motor output power. The damper winding is a parameter.](image)

**5.6 Conclusion**

In this chapter it is shown that the IEEE 112 test method B is an acceptable method for the evaluating of the share of losses in both induction and synchronous reluctance motors in converter supply. The effect of semi-magnetic slot-wedges in the stator slot openings has been investigated as a parameter for both motor types. According to the results obtained the slot wedges decrease the total losses in both motor types in a variable speed drive. The effect of the air-gap length has been examined for the SynRM. The increased SynRM air-gap length decreases also the total losses of the motor. In a variable speed drive, the losses of the SynRM with slot-wedges remain lower than those of the standard induction motor without slot-wedges. At rated load the iron and additional losses of the SynRM are over 60% higher than those of the standard induction motor in variable speed drives. The induction motors rotor ohmic losses are higher than the “extra” iron losses of the SynRM. Despite of this, the total losses of the SynRM are rather equal to or larger than the losses of the IM. This is explained by the fact that the lower power factor of the SynRM increases the stator ohmic losses of the SynRM. According to the results obtained in this study it may be concluded that in order to achieve a SynRM producing an efficiency that is considerably higher compared to the original induction motor efficiency the rotor magnetic circuit of the SynRM should be manufactured of a better iron material.
6 CONCLUSION

The aim of this research work was to study the properties of the SynRM and to see whether it can compete in performance with high efficiency standard induction motors in power ranges typical for low voltage IMs used in industrial applications.

The benefit of the SynRM with respect to the IM is that there are no copper losses in the rotor. Leaving the rotor cage out is thus, in principle, possible. However, the presence of the rotor cage might help in several ways: It binds the rotor laminations together to produce a firm construction; it also helps to control the machine during transients acting as a damper winding; and it makes a direct network start possible.

With respect to the motor performance – especially the full load power factor – the absence of the rotor currents is not only a benefit but also a problem. In larger induction motors the power factor is high – in the range of 0.9. This is possible since the rotor current acts as a compensating current for the stator armature reaction. The sum magneto-motive force of the stator and the rotor magnetises the IM air-gap. Since there is no current in the SynRM rotor, especially the quadrature axis inductance of the SynRM should be extremely small. Otherwise, the stator armature reaction inevitably produces for the SynRM – by turning the air-gap flux linkage away from the d-axis – a power factor that is lower than the power factor of a good IM.

The low quadrature axis inductance brings forward some problems concerning the power electronic control of the SynRM. E.g. a high switching frequency is needed to avoid a large current ripple in the q-axis current.

The power factor of both motor types may be increased by maximising the direct axis magnetising inductance. Specifically the magnetising inductance of the SynRM should be large. This, unfortunately, requires a smaller air-gap than it has been used traditionally in IMs. The smaller air-gap brings some disadvantages with it. Firstly, the manufacturing of a SynRM must be more accurate and becomes thus probably more expensive than the manufacturing of an IM. Secondly, the rotor surface losses caused by the air-gap spatial harmonics are increased at the same time as the air-gap is diminished. This, thirdly, requires the use of slot wedges or other methods to smoothen the air-gap flux density. This also suggests the use of better magnetic circuit material, which again increases the cost of the SynRM.

On the other hand, the prototype motor was constructed using exactly the same parts coming from a normal IM production line. The assembling of a small air-gap rotor progressed well and without any problems. It might thus be concluded that the manufacturing technology at least in ABB’s Vaasa factory is accurate enough for the producing of TLA-rotor SynRMs with low air-gaps.

The study shows that, with some special measures, the efficiency of the TLA SynRM is equivalent to that of a high efficiency IM. The power factor at nominal load will probably remain smaller and thus the stator of the SynRM should be, contrary to this study, designed particularly for the SynRM. It might be even assumed that using more copper and less stator core material a more suitable resistance and better inductance values for the machine may be possibly achieved – at least the q-axis inductance should become a little smaller, which unexpectedly could improve the power factor.

Compared to the TLA-rotor machine the ALA-rotor structure offers better possibilities to produce larger inductance ratios. A high inductance ratio might be achievable even with an air-gap slightly larger than that of the TLA-rotor machine. This may probably bring a better power...
factor than that of an IM. The ALA-rotor iron losses, however, are sensitive to the spatial air-gap harmonics. The ALA-type rotor must thus be manufactured of high-resistivity thin laminations to avoid excessive iron losses. Also the methods of manufacturing are totally differing from those applied for the manufacturing of a TLA-rotor. The latter may be manufactured using the same tools which are used for the producing of IMs. The ALA-type rotor might for this reason be out of discussion for the time being.

One benefit of the SynRM remains to be discussed. The big rotor inductance ratio is of a considerable use for the estimating of the rotor position of the motor. This again is shown to be very advantageous for the designing of the rotor position sensorless drive. For example the laboratory test results in this thesis have partly been reached using a sensorless SynRMDTC drive.

It is probably easier to achieve an accurate sensorless speed control with a SynRM drive than with an IM drive. The slip of an IM makes the accurate estimation of speed extremely difficult. Paper machine drives for example require a speed accuracy of 0.1%. This accuracy may not be achieved with present-day position sensorless IM drives as for a position sensorless SynRM drive this level may easily be achieved. Also the partial load good properties of the synchronous reluctance motor might be encouraging. Often industrial motors run at partial loads and, in such cases, the SynRM might be a better choice than an IM.

The adding permanent magnets to the SynRM rotor may be one alternative to improve the SynRM performance and especially its power factor and efficiency. Using PM material may improve the motor power factor and consequently reduce the motor current. The reduced currents also reduce the motor stator ohmic losses. On the other hand, the reluctance torque reduces the quantity of the expensive PM material needed and makes this solution cheaper than the corresponding permanent magnet motor (Haataja 2002b).
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APPENDIX A  Construction of the 30 kW induction motor

Table A1. Stator slot parameter.

<table>
<thead>
<tr>
<th>Constants</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot opening</td>
<td>$x_1$</td>
<td>3.8</td>
<td>mm</td>
</tr>
<tr>
<td>Slot width 2</td>
<td>$x_2$</td>
<td>7.7</td>
<td>mm</td>
</tr>
<tr>
<td>Slot width 3</td>
<td>$x_3$</td>
<td>9.9</td>
<td>mm</td>
</tr>
<tr>
<td>Slot height 1</td>
<td>$y_1$</td>
<td>1.0</td>
<td>mm</td>
</tr>
<tr>
<td>Slot height 2</td>
<td>$y_2$</td>
<td>1.0</td>
<td>mm</td>
</tr>
<tr>
<td>Slot height 3</td>
<td>$y_3$</td>
<td>16.85</td>
<td>mm</td>
</tr>
<tr>
<td>Slot height 4</td>
<td>$y_4$</td>
<td>23.8</td>
<td>mm</td>
</tr>
</tbody>
</table>

Figure A1. Stator slot

Table A2. Data of the initial IM.

<table>
<thead>
<tr>
<th>Constants</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal power</td>
<td>$P_N$</td>
<td>30</td>
<td>kW</td>
</tr>
<tr>
<td>Stator inner diameter</td>
<td>$D$</td>
<td>215</td>
<td>mm</td>
</tr>
<tr>
<td>Length</td>
<td>$l$</td>
<td>205</td>
<td>mm</td>
</tr>
<tr>
<td>Pole pair</td>
<td>$p$</td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>Air-gap diameter</td>
<td>$\delta_m$</td>
<td>0.8</td>
<td>mm</td>
</tr>
<tr>
<td>Number of stator slots</td>
<td>$Q_s$</td>
<td>48</td>
<td></td>
</tr>
<tr>
<td>Number of wires / slot</td>
<td>$N_{st}$</td>
<td>13</td>
<td></td>
</tr>
<tr>
<td>Number of rotor bars</td>
<td>$Q_r$</td>
<td>36</td>
<td></td>
</tr>
<tr>
<td>Steel quality</td>
<td></td>
<td>DK 70</td>
<td></td>
</tr>
<tr>
<td>Stator outer diameter</td>
<td></td>
<td>323</td>
<td>mm</td>
</tr>
<tr>
<td>Connection</td>
<td>$D$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Voltage</td>
<td>$U_N$</td>
<td>400</td>
<td>V</td>
</tr>
</tbody>
</table>

Figure A2. Rotor bar

Table A3. Rotor bar parameter.

<table>
<thead>
<tr>
<th>Constants</th>
<th>Symbol</th>
<th>Value</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Bar width 1</td>
<td>$x_1$</td>
<td>6.0</td>
<td>mm</td>
</tr>
<tr>
<td>Bar width 2</td>
<td>$x_2$</td>
<td>2.5</td>
<td>mm</td>
</tr>
<tr>
<td>Bar width 3</td>
<td>$x_3$</td>
<td>6.6</td>
<td>mm</td>
</tr>
<tr>
<td>Bar width 4</td>
<td>$x_4$</td>
<td>2.9</td>
<td>mm</td>
</tr>
<tr>
<td>Bridge height 1</td>
<td>$y_1$</td>
<td>0.7</td>
<td>mm</td>
</tr>
<tr>
<td>Bar height 2</td>
<td>$y_2$</td>
<td>5.7</td>
<td>mm</td>
</tr>
<tr>
<td>Bar height 3</td>
<td>$y_3$</td>
<td>8.0</td>
<td>mm</td>
</tr>
<tr>
<td>Bar height 4</td>
<td>$y_4$</td>
<td>3.05</td>
<td>mm</td>
</tr>
<tr>
<td>Bar height 5</td>
<td>$y_5$</td>
<td>21.2</td>
<td>mm</td>
</tr>
<tr>
<td>Bar height 6</td>
<td>$y_6$</td>
<td>1.45</td>
<td>mm</td>
</tr>
</tbody>
</table>
APPENDIX B  The designed rotor photo

In Fig. B.1. the photo is shown of the rotor constructions of the optimum designed TLA-type 30 kW SynRM rotor.

Figure B.1. Skewed rotor with rotor lamination.
### APPENDIX C  THE MEASURING TABLE OF MOTOR 6

#### IEEE Std 112-1996

**Method B**

<table>
<thead>
<tr>
<th>Design SynRM</th>
<th>Frame ST200M</th>
<th>Power 30 kW</th>
<th>Phase 3</th>
</tr>
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<tbody>
<tr>
<td>Frequency 50</td>
<td>Volts 690 / 400</td>
<td>Synchronous r/min 1500</td>
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</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Item</th>
<th>Description</th>
<th>Load, in % of rated</th>
<th>25</th>
<th>50</th>
<th>75</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>p</td>
<td></td>
<td>2</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>k1</td>
<td></td>
<td>234.5</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>k2</td>
<td></td>
<td>9.549</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>Average Cold Stator Winding Resistance Between Terminals</td>
<td>0.1614 Ohm</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>Ambient Temperature, in °C</td>
<td>21</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>Rated Load Heat Run Stator Winding Resistance Between Terminals</td>
<td>0.2063 Ohm</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>Decrees C Temperature Rise rating</td>
<td>89.43 °C</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>Ambient Temperature, in °C</td>
<td>22</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

#### IEC34-2 Summation of losses

<table>
<thead>
<tr>
<th>Item</th>
<th>Description</th>
<th>Load, in % of rated</th>
<th>25</th>
<th>50</th>
<th>75</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>n</td>
<td></td>
<td>4</td>
<td></td>
<td></td>
<td></td>
</tr>
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<td>X</td>
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<td>225</td>
<td></td>
<td></td>
<td></td>
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<td>3</td>
<td>k1</td>
<td></td>
<td>234.5</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>k2</td>
<td></td>
<td>9.549</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>Average Cold Stator Winding Resistance Between Terminals</td>
<td>0.1614 Ohm</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>Ambient Temperature, in °C</td>
<td>21</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>Rated Load Heat Run Stator Winding Resistance Between Terminals</td>
<td>0.2063 Ohm</td>
<td></td>
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</tr>
<tr>
<td>8</td>
<td>Decrees C Temperature Rise rating</td>
<td>89.43 °C</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>Ambient Temperature, in °C</td>
<td>22</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

#### Load, in % of rated

<table>
<thead>
<tr>
<th>Load, in % of rated</th>
<th>25</th>
<th>50</th>
<th>75</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power Factor, in %</td>
<td>67.7</td>
<td>75.8</td>
<td>74.6</td>
<td>72.1</td>
</tr>
<tr>
<td>Efficiency, in %</td>
<td>90.0</td>
<td>91.4</td>
<td>92.7</td>
<td>92.7</td>
</tr>
<tr>
<td>Speed, in r/min</td>
<td>1500</td>
<td>1500</td>
<td>1500</td>
<td>1500</td>
</tr>
<tr>
<td>Line Current, in A</td>
<td>19.77</td>
<td>29.31</td>
<td>48.23</td>
<td>58.65</td>
</tr>
</tbody>
</table>

---

*(Calculations for wye connection)*

---

\( t \) = temperature of stator winding as determined from stator resistance or temperature detector during test.
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