On Design and Analysis of Synchronous Permanent Magnet Machines for Field-weakening Operation in Hybrid Electric Vehicles

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Abstract

A regular vehicle of today is equipped with an internal combustion engine that runs on either gasoline or diesel, which are fossil fuels from oil reserves that are millions of years old. In all types of combustion processes carbon dioxide and several other emissions are produced. There are none known technologies of today that can reduce the emissions of carbon dioxide from combustion, but the amount that is produced is mainly dependent on the fuel that is used. Combustion of fossil fuels increases the contamination of carbon dioxide in the atmosphere and diminishes the oil resources. The results are global warming and empty oil reserves within a few decades with the current production tempo, in addition to many other pollution effects that are harmful to the environment. A transition towards a society based on sustainable transportation is therefore urgent. The hydrogen fuel cell powered car with an electric propulsion system has the potential to be the car of the future that possesses the required characteristics of no harmful tailpipe emissions. There are some obstacles in the way for an early commercialisation, including the expensive catalysts used today and the lack of an infrastructure based on hydrogen, though. The hybrid electric vehicle, with both a conventional as well as an electric drivetrain, is a natural candidate for making the transition from the conventional car towards the car of the future.

This thesis is focused on the design and analysis of permanent magnet machines for a novel hybrid electric vehicle drive system called the Four Quadrant Transducer. A number of electrical machine aspects are identified, including cores of soft magnetic composites, fractional pitch concentrated windings, core segmentation, novel machine topologies and cost effective production methods. The main objective is to analyse and judge the many unconventional machine aspects of which some may have the potential to improve the performance and reduce the cost of permanent magnet machines. Another objective is to study the effects of the use of fossil fuels and describe them with a new perspective and thereby make one small contribution to the debate about energy issues.

Much focus has been spent on the theory of concentrated windings for permanent magnet machines. The potential parasitic effects and methods to improve the torque performance have been described. Other topics that have been given a high priority are material and power loss studies. An important contribution to the understanding of iron losses during field-weakening operation has been presented. A comprehensive use of finite element modeling has been done in the analysis combined with measurements on several laboratory prototypes.

The Four Quadrant Transducer drivetrain and its two electrical machines intended for a mid-sized passenger car has been studied. The gearbox can be of a simple single stage type, which reduces the mechanical complexity and makes the traction performance of the vehicle smooth, without gear changes and drops in power. Simulations on a complete hybrid system show that fuel savings of more than 40% compared to a conventional vehicle can be achieved at city-traffic driving. The savings are modest at highway driving, since the engine is required to operate at high power during such conditions, and the support from the electrical system is negligible. The laboratory prototypes have shown that it is possible to manufacture high performance electrical machines with high material utilization and potential for automated production. The described concepts offer cost effective solutions for future drive systems in automotive and industrial applications. A number of weaknesses with the presented constructions have also been characterized, which should serve as guidelines for creating more optimized machines.

**Keywords:** Permanent magnet machines, Field-weakening, Hybrid electric vehicles, Soft magnetic composites, Fractional pitch concentrated windings, Power losses, Sliprings, Oil resources, Global warming, Renewable energy
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1 Introduction

1.1 Background

A regular vehicle of today is equipped with an internal combustion engine that runs on either gasoline or diesel, which are fossil fuels from oil reserves that are millions of years old. In all types of combustion processes carbon dioxide and several other emissions are produced. There are none known technologies of today that can reduce the emissions of carbon dioxide from combustion, but the amount that is produced is entirely dependent on the fuel that is used. Combustion of fossil fuels increases the contamination of carbon dioxide in the atmosphere and diminishes the oil resources. The results are global warming and empty oil reserves within a few decades with the current production tempo, in addition to many other pollution effects that are harmful to the environment. Is there a vehicle system without tailpipe emissions? Yes, there is. The electric vehicle is a zero emission vehicle. The electric vehicle is not a new invention. Already in the beginning of the past century, electric cars were competing with the engine driven cars and they were much appreciated for their quiet operation. They had a big drawback though, the poor battery capacity made the driving range short before recharging was needed. When the engine driven car later was equipped with an electric start motor, at the expense of the unappreciated manual start, the electric vehicles disappeared from the market. Battery electric vehicles of today also suffer for the lack of a satisfactory driving range for the same reason, but also because high-energy batteries are quite expensive. Fuel cells are another form of energy source that can power an electric propulsion system. The driving range is longer than for a battery, but the system is also expensive and an infrastructure of hydrogen tanking stations is required. The main obstacle of a faster introduction of more environmental friendly drivetrains though, is probably the investments already made in the production of conventional engines. The hybrid electric vehicle is a suitable compromise. It has two drivetrains; one conventional engine and one electric drive system. The electric system is supporting the engine so that the system efficiency is increased. The installed electric power is normally much lower than the engine power, except for series hybrids. The investment needed in new equipment is therefore relatively small and current production facilities can still be used. Hybrid electric vehicles are expected to rise rapidly in numbers in a couple of years mainly due to tougher emission regulations.

The permanent magnet electrical machine is also gaining popularity. The big interest in this machine type, which started in the mid-1980s by the introduction of the newly invented neodymium iron boron rare-earth permanent magnets, is due to its superiorly high efficiency, high torque density and accurate controllability. These advantages, and the fact that the prices of rare-earth permanent magnets are decreasing, make the permanent magnet machine especially suitable in applications like electric vehicles, robotics and high efficiency industrial drives.

This thesis concerns the design and analysis of permanent magnet machines for a novel hybrid electric drive system called the Four Quadrant Transducer. The new drive system is also analysed and compared to existing hybrid drivetrains.
1.2 Objectives

The automotive industry is very cost sensitive. Weight minimizations of the vehicle components are also big concerns, since higher weight translates into higher fuel consumption. Then there are the space limitations in the vehicle. The electric drive system must compete against the conventional drivetrain about the available space mainly in the engine compartment. All these factors result in a tough specification for the electric system. The research work described in this thesis is focused on electrical machine design with the tough specification in mind. A number of qualified machine aspects are identified, including cores of soft magnetic composites, fractional pitch concentrated windings, core segmentation, novel machine topologies and cost effective production methods. The main objective is to analyse and judge the many unconventional machine aspects of which some may have the potential to improve the performance and reduce the cost of permanent magnet machines.

Today, the automotive market is almost entirely based on conventional engine driven vehicles. The introduction of more environmental friendly vehicles is slow and the resistance from the establishment has been huge. One important objective is therefore to study the effects of the use of fossil fuels and describe them with a new perspective and thereby make one small contribution to the debate.

The novel Four Quadrant Transducer drive system is the platform of which the major work is based. However, the research achievements are not limited for this system alone, but the findings are also interesting in other drive applications with equally high demands on performance.

1.3 Outline of the thesis

This thesis is organised as an extended summary mainly based on published conference proceedings that are included. Chapter 2 and Chapter 3 are written in order to broaden the perspectives in the subject of electric propulsion for road vehicles, and can be read without any knowledge about machines, Chapter 2 even by non-engineers.

Chapter 2 defines the meaning of sustainable transportation. The objective of this chapter is partly to motivate the research on electric and hybrid electric vehicles, but also to give the reader, hopefully some decision-makers within government and industry, some valuable ideas around this important subject about energy issues. The chapter starts with a presentation of the exhaust gases from conventional vehicles and their impact on the environment. Detailed statistics on the oil business are introduced in order to show the contradiction in the tempo oil is burnt and what is left of it, which can seem like a big paradox. The perspective is broadened further by a presentation of humanly caused effects like global warming, pollution and their potentially resulting social-economical problems. The chapter is ended by a discussion about trends in vehicle technology and renewable energy sources.

Chapter 3 describes the fundamentals of road vehicles. The objective of this chapter is to give an introduction to the conventional and electric vehicle components. The running resistance of a vehicle in motion is defined. Factors that influence the fuel consumption the most are discussed. The basic operating principles of the gasoline and the diesel engines are presented. Short descriptions on the main components of conventional vehicles followed by the main components for electric propulsion, including an overview of electrical machine types, are made. The most typical hybrid electric system topologies and some commercially available vehicles are presented.
Chapter 4 focuses on the design of permanent magnet machines. The objective of this chapter is to present some strengths and weaknesses of some different permanent magnet machine topologies mainly suitable for servo and traction applications. This part of the thesis contributes to a better understanding of these types of electrical machines. General characteristics and important findings are explained in this chapter, while more detailed descriptions are given in the papers.

Chapter 5 deals with the novel hybrid electric drive system called the Four Quadrant Transducer. The objective of this chapter is to evaluate the 4QT concept. The characteristic features of the system are described and some simulation results are discussed. The two electrical machines that are specifically designed using this concept for a mid-sized passenger car are presented. The system and the two electrical machines are also analysed in several papers.

Chapter 6 presents the laboratory prototype of the Four Quadrant Transducer with its two electrical machines. The objective of this chapter is to evaluate the chosen electrical machine constructions. The manufacturing process is illustrated with the aid of photos and text. Some potential manufacturing difficulties are described. Measurement results from tests on the electrical machines are analysed.

Chapter 7 summarizes this thesis by giving some conclusive remarks and gives some suggestions of future work.

1.4 Publications

The following papers are listed in chronological order of publication:

Paper 1

*Electromagnetic Transducer for Hybrid Electric Vehicles*
Freddy Magnussen, Chandur Sadarangani
Published in the Proceedings of the Nordic Workshop on Power and Industrial Electronics (NORPIE), Stockholm, Sweden, 12-14 August 2002.

The paper presents the novel hybrid electric vehicle drive system called the Four Quadrant Transducer. The operating principle and the strengths and weaknesses of the system are discussed. Simulation results and estimated fuel consumption for a mid-sized passenger car are reported. Some design ideas on suitable electrical machine topologies are given.

Paper 2

*Winding Factors and Joule Losses of Permanent Magnet Machines with Concentrated Windings*
F. Magnussen, C. Sadarangani
Published in the Proceedings of the IEEE International Electric Machines and Drives Conference (IEMDC), Madison, Wisconsin, USA, 1-4 June 2003.

The paper presents the theory to calculate the winding factors of electrical machines, especially for fractional pitch concentrated winding machines, by means of electromotive force phasors. Some pole and slot combinations that give high fundamental winding factors and low cogging torques without the need of skewing are described. Furthermore, the magnetomotive force, the
joule losses and the axial build of machines with both distributed and concentrated machines are compared.

**Paper 3**

*Design of Compact Permanent Magnet Machines for a Novel HEV Propulsion System*

F. Magnussen, P. Thelin, C. Sadarangani


The paper presents the electromagnetic design of two novel electrical machines for the Four Quadrant Transducer. Both machines are synchronous permanent magnet machines. The first machine has a core of Soft Magnetic Composites and gramme-windings, while the second one has grain-oriented teeth and non-oriented yoke of silicon iron laminations and concentrated windings. The cooling aspects and the estimated cost of the active materials are described.

**Paper 4**

*Measurements on Slip Ring Units for Characterization of Performance*

F. Magnussen, E. Nordlund, S. Châtelet, C. Sadarangani


The paper deals with measurement results and analysis of slipring units for performance characterization. Some typical physical behaviour of brush operation for various parameters like speed, electrical frequency and current density are identified and explained.

**Paper 5**

*Performance Evaluation of Permanent Magnet Synchronous Machines with Concentrated and Distributed Windings Including the Effect of Field-weakening*

F. Magnussen, P. Thelin, C. Sadarangani

Published in the Proceedings of the IEE Second International Conference on Power Electronics, Machines and Drives (PMD), Edinburgh, United Kingdom, 31 March – 2 April 2004.

The paper compares the performances of three PM machines with equal outer rotors and active dimensions. They differ by their number of slots and winding topology. One machine has distributed windings, while the other two have concentrated windings with different numbers of slots per pole per phase. Their power capabilities considering inverter capacity are shown in normalized quantities.

**Paper 6**

*Analysis of a PM Machine with Concentrated Fractional Pitch Windings*

F. Magnussen, D. Svechkarenko, P. Thelin, C. Sadarangani

Published in the Proceedings of the Nordic Workshop on Power and Industrial Electronics (NORPIE), Trondheim, Norway, 14-16 June 2004.

Submitted for possible publication in the EPE Journal.

The paper focuses on some parasitic effects in the laminated machine with concentrated windings designed for the Four Quadrant Transducer. Alternating magnetic fields in the rotor and magnetic noise, which are
typical characteristics of the design, are analysed. The thermal performance
of the machine is characterized based on measurements and simulations.

**Paper 7**
*Testing of Silver-, Copper- and Electro-Graphite Brush Materials for Slip Ring Units*
E. Nordlund, F. Magnussen, G. Bassilious, P. Thelin
Published in the Proceedings of the Nordic Workshop on Power and Industrial Electronics (NORPIE), Trondheim, Norway, 14-16 June 2004.

The paper is, together with Paper 4, dedicated to the behaviour of slipring units under variable speed drive operation. The paper advances the study further by testing brushes of different materials - silver, copper and pure graphite.

**Paper 8**
*Iron Losses in Salient Permanent Magnet Machines at Field-weakening Operation*
F. Magnussen, Y.K. Chin, J. Soulard, A. Broddefalk, S. Eriksson, C. Sadarangani
Submitted for possible publication in the IEEE Transactions on Industry Applications.

The paper deals with iron losses in inset and interior PM machines under field-weakening conditions. The magnetic field in the stator core has a very high harmonic content during such operation. The analysis is performed with the use of finite element computations and measurements on several machines.

**Paper 9**
*Analysis of a PM Machine with Soft Magnetic Composites Core*
F. Magnussen, D. Svechkarenko, P. Thelin, C. Sadarangani

The paper focuses on the overall concept of the combined axial-radial flux machine for the Four Quadrant Transducer and the particularities linked to the use of soft magnetic composites for the core material. The magnetic and thermal performances are analysed using finite element computations. Some measurements on a laboratory prototype are presented.
2 Sustainable transportation

This chapter discusses passenger vehicles classified as low emission vehicles. The different classes are defined in measurable terms according to established standards. In order to give a broader perspective of factors that influence the design of motor vehicles, statistical information on oil resources in the world are presented. Emissions contributing to global warming and pollution are discussed. Furthermore, the subject is analysed from a more humanistic perspective to give a definition on "sustainable mobility". The phrase “sustainable development” was adopted at the Earth Summit Conference in Rio de Janeiro in 1992 (United Nations Conference on Environment and Development). It is a basic philosophy for the protection of the global environment. Governments from around 180 nations were represented and signed up an action plan, “Agenda 21”, which is based on this philosophy.

2.1 Global road transportation

2.1.1 Exhaust emissions from vehicles

2.1.1.1 Introduction

In all types of combustion processes carbon dioxide emissions are produced, irrespective of whether the fuels are based on fossil or renewable resources. Combustion of fossil fuels increases the contamination of carbon dioxide in the atmosphere. Fossil fuels are taken from oil-, gas- and coal-reserves that are million of years old and are obviously not naturally part in an environmental cycle with a short time span. Burning of fuels that come from renewable energy sources is not regarded to increase the amount of CO₂ in the atmosphere, since the biomass is bounding an equivalent amount of carbon dioxide during its growth. Today, there is no known technology that can reduce the emissions of carbon dioxide from combustion, but the amount that is produced is mainly dependent on the fuel that is used [1]. A normal vehicle of today is equipped with an internal combustion engine that runs on either gasoline or diesel. Diesel cars are very popular in Europe, with 39,4 percent market share in the first half of 2002 [2], because of more beneficial tax regulations for this engine type within the EU compared to the rest of the world. Diesel engines are more efficient than gasoline engines, but their exhaust emissions have higher contents of smog-forming and cancer-causing soot particles and nitrogen oxides. Diesel vehicles account for only 10 percent of passenger cars in Japan, but they generate more than 30 percent of all nitrogen oxide emissions. Japan and India are attempting to ban diesel vehicles altogether. Diesel-powered passenger cars are very rare in the USA [3].

2.1.1.2 The main contaminants of exhaust gas from engines

Fossil fuel is essentially a mixture of hydrocarbon compounds. When an internal combustion engine burns fuel, in the presence of oxygen, the ideal reaction result is carbon dioxide, water
vapour and energy. In reality, since automobile engines of today are not incredibly efficient, these products are produced along with three major contaminants of exhaust gas: hydrocarbons (HC), carbon monoxide (CO) and oxides of nitrogen (NOx). It is these three types of gases that catalytic converters are trying to remove by using precious metals such as platinum, palladium and rhodium as catalysts [4]. A catalytic converter has no effect on the emission of carbon dioxide, which is only dependent of the fuel consumption of the vehicle. Some short descriptions of contaminants in exhaust emissions are given below.

2.1.1.2.1 Carbon dioxide (CO2) and other greenhouse gases
Carbon dioxide is the number one gas that contributes most to the greenhouse effect and the global warming. This gas along with other greenhouse gases like methane (CH4), fluorocarbons (FC), nitrous oxide (N2O), carbon monoxide (CO), hydrofluorocarbons (HFC) and sulphur hexafluoride (SF6), increase in the atmosphere and amplify the natural greenhouse effect. The greenhouse gases increase the risk of a future change of the climate on Tellus (earth) [1].

2.1.1.2.2 Nitrogen oxides (NOx)
Nitrogen oxides include nitrogen oxide (NO) and nitrogen dioxide (NO2). They are both gases that are created through combustion and emitted to the air. These emissions can travel long distances and influence vegetation and animal life. They can also cause acid rain and be responsible for the creation of ozone close to the ground [1].

2.1.1.2.3 Carbon monoxide (CO)
Carbon monoxide is created by incomplete combustion of fuel. These kinds of emissions have mainly local effects and can for instance cause breathing problems for people with a heart weakness. Carbon monoxide is also one of the greenhouse gases. Incomplete combustion is especially usual when internal combustion engines run at no-load, at high speed or at cold-start [1].

2.1.1.2.4 Hydrocarbons (HC)
Volatile Organic Compounds (VOC) is a common name for several organic substances, mainly hydrocarbons, which have negative impact on the environment. The hydrocarbons are rest products caused by imperfect combustion of fossil fuels. The VOC contribute to the build up of ozone close to the ground, which for instance damages fields and forests. Hydrocarbons also cause health problems like eye and throat irritation, asthma and allergy, and some hydrocarbons even cause cancer [1].

2.1.1.2.5 Sulphur oxides (SOx)
Sulphur oxides, mainly sulphur dioxide (SO2), which in the air is transformed into sulphur acid, is the single most important cause of acid rain that damages fields and water. The sulphur rich air mixture can travel long distances before it drops down with the rain. The acid is responsible for damages on the vegetation, like forest death. The normal food chain in vulnerable areas can therefore be broken and sensitive species of animals and plants can die [1].

2.1.1.2.6 Particles and Polycyclic Aromatic Hydrocarbons (PAHs)
Fuel combustion also causes emission of particles, mainly soot, due to imperfect combustion. The particles contain hydrocarbons, some of which, for instance the Polycyclic Aromatic Hydrocarbons (PAHs), can cause cancer. The health risks corresponding to the emissions of particles are mainly of local concern and concentrated around big cities with many vehicles
and factories. Small particles can enter the lungs from the inhaled air and sensitive persons may get breathing difficulties. The smog that a number of cities around the world experience is due to the emissions of particles, nitrogen oxides and volatile organic compounds. The diesel vehicles are responsible for the major part of the particle emissions within transport, but new particle filters for diesel engines can drastically reduce the amount, in some cases to levels that even outperforms gasoline powered vehicles [1].

2.1.1.3 Alternative fuels

An increased use of alternative fuels is one measure to reduce the negative impacts on the environment from transport. The alternative fuels are mainly considered to be those from renewable resources like ethanol and methanol from crops or cellulose and biomass from rot of organic material. Alternative fuels like methanol from natural gas or gasol are of fossil origin, but have lower emissions than gasoline and diesel. Even hydrogen gas is an alternative fuel with a high potential in a longer perspective [1]. These examples of alternative fuels are briefly discussed below. Life Cycle Assessments (LCA) of total emissions from Otto and diesel engines for different fuels, including alternative fuels, are given in Table I.

2.1.1.3.1 Ethanol

All gasoline powered passenger cars can without any technical modifications be tanked with 10-15% of ethanol. It is also possible for diesel engines after small adjustments. The gain is less emission of harmful exhaust gases, mainly reduced carbon dioxide, nitrogen oxides and soot particles. Ethanol is produced through fermentation of sugar, which can be extracted from crops or cellulose [1].

2.1.1.3.2 Methanol

Methanol is just like ethanol an alcohol that can be used as fuel and can be produced both from renewable and fossil resources. The methanol that is produced today is based entirely on fossil fuels, mainly natural gas, but it is also possible to produce methanol from biomass. Emissions from the use of methanol are lower than for gasoline and diesel. The fuel is poisonous and corrosive [1].

2.1.1.3.3 Biogas

Biomass is today considered to be the biofuel with the least harmful impacts on the environment. The gas consists of methane gas, carbon dioxide and water. The methane gas can be used as fuel in vehicle engines specially designed for this fuel. Biomass is produced from organic rest products and from cultivated biomass. Methane gas is a greenhouse gas, but the total exhaust emissions from using this fuel are much lower than compared to gasoline and diesel [1].

2.1.1.3.4 Natural gas

Natural gas has the same main contaminant as biogas, namely methane, but it has a different origin. Natural gas is a fossil fuel, while biogas is a renewable fuel. The net exhaust emissions of carbon dioxide from the use of natural gas is 15-20% lower than by using gasoline, but it is equivalent with the diesel fuel, due to the higher efficiency of the diesel engine. Engines using natural gas have mainly low emissions of particles and hydrocarbons. Methane gas is a strong greenhouse gas [1].

2.1.1.3.5 Liquified Petroleum Gas (LPG)/Gasol

Liquified Petroleum Gas (LPG) is gasol in liquified form and is basically a fossil fuel. The gas is relatively cheap and gives low exhaust emissions compared with gasoline and diesel [1].
2.1.1.3.6 Hydrogen gas

Hydrogen gas used as fuel for fuel cells is considered to be an optimal energy carrier for the future. The exhaust emission from the fuel cell is pure water vapour. Hydrogen is basically also an unlimited substance on earth. However, there are two main problems with hydrogen gas that first need to be solved. Firstly, the hydrogen gas is not found naturally (except in very small quantities in the atmosphere) and must consequently be produced, either through electrolytic or biochemical processes. Secondly, hydrogen gas is very difficult to store. The gas can be stored pressurized or cooled to liquid form in storage tanks. Nevertheless, the energy that is needed for the production and storage is not negligible. Since the energy sources of today are dominated by the fossil fuels, even fuel cells that run on hydrogen have emissions of greenhouse gases, if the energy for the production of the fuel is included. The energy sources of the future are hopefully renewable with no or very small impact on the environment. Some existing examples are hydro, wind, wave, tidal and solar energy. The hydrogen gas can then be the energy carrier of the future and end the damaging emissions of greenhouse gases based on burning of fossil fuel. An increased amount of hydrogen in the atmosphere due to leakage from storage tanks may cause degradation of the ozone layer in the stratosphere, but this potential effect is not known in detail today [1].

2.1.1.4 Exhaust emission development

The total yearly emissions of carbon dioxide in the world have decreased by 15 percent in the period 1990-2000, according to the UN [5]. In the same period, the carbon dioxide emissions caused by transport have increased by 19 percent, while the emissions of carbon monoxide and nitrogen oxides have decreased with 18 percent and increased with 5 percent, respectively. The emissions of carbon dioxide from transport were less than 19 percent of the world total in 1990, but this relative figure was increased to more than 26 percent in 2000. The absolute figures are given in Figure 2.1. There are two ways of reducing the emissions of carbon dioxide from the tailpipes of a standard vehicle, either to make the engine more energy efficient or to use fuel types with less carbon content, e.g. alcohol fuels.

2.1.1.5 Life Cycle Assessment (LCA) of emissions from engines

There are several alternative fuels that can be used exclusively or combined in conventional Otto and diesel engines. In order to compare the alternative fuels with the traditional gasoline and diesel fuels, in an environmental impact perspective, it is important to analyse the total emissions during the whole life cycle. This means that the emissions from the production of the fuel, the emissions in the fuel use chain and the deposition after the final use of the fuel are calculated. A Life Cycle Assessment (LCA), which is the generic term, is a calculation of the total costs and environmental effects of the fuel. LCA-calculation results of total emissions from Otto and diesel engines for different traditional and alternative fuels are shown in Table I [1]. It is clearly illustrated that emissions can be greatly reduced by switching from traditional to alternative fuels or mixing them up.
Figure 2.1: (a) The total yearly emissions of carbon dioxide in the world respectively the emissions exclusively from the transport sector. (b) The emissions of carbon monoxide and nitrogen oxides from the transport sector.

Table I: Life Cycle Assessment (LCA) of total emissions from Otto and diesel engines for different fuels (mg/MJ fuel).

<table>
<thead>
<tr>
<th>Fuel</th>
<th>NO\textsubscript{x}</th>
<th>HC</th>
<th>Particles</th>
<th>CO</th>
<th>SO\textsubscript{x}</th>
<th>N\textsubscript{2}O</th>
<th>CH\textsubscript{4}</th>
<th>CO\textsubscript{2}</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ethanol from corn</td>
<td>98</td>
<td>30</td>
<td>52</td>
<td>320</td>
<td>11</td>
<td>28</td>
<td>52</td>
<td>18 000</td>
</tr>
<tr>
<td>Ethanol from cellulose</td>
<td>60</td>
<td>30</td>
<td>3</td>
<td>160</td>
<td>25</td>
<td>3</td>
<td>2</td>
<td>6 000</td>
</tr>
<tr>
<td>Methanol from biomass</td>
<td>~85</td>
<td>~60</td>
<td>~3.5</td>
<td>~630</td>
<td>~13</td>
<td>~50</td>
<td>~0.7</td>
<td>20 000</td>
</tr>
<tr>
<td>Biogas</td>
<td>31</td>
<td>19</td>
<td>1.9</td>
<td>36</td>
<td>1</td>
<td>0</td>
<td>600</td>
<td>900</td>
</tr>
<tr>
<td>Liquified Petroleum Gas</td>
<td>55</td>
<td>49</td>
<td>2.8</td>
<td>37</td>
<td>16</td>
<td>0</td>
<td>22</td>
<td>68 000</td>
</tr>
<tr>
<td>Gasoline</td>
<td>68</td>
<td>69</td>
<td>4.5</td>
<td>180</td>
<td>30</td>
<td>20</td>
<td>9</td>
<td>79 000</td>
</tr>
<tr>
<td>Ethanol from corn</td>
<td>~300</td>
<td>~20</td>
<td>~1.2</td>
<td>~15</td>
<td>~0.4</td>
<td>0</td>
<td>0</td>
<td>~6 000</td>
</tr>
<tr>
<td>Ethanol from cellulose</td>
<td>~200</td>
<td>~20</td>
<td>~1.3</td>
<td>~20</td>
<td>~2.5</td>
<td>0</td>
<td>0</td>
<td>~6 000</td>
</tr>
<tr>
<td>Methanol from biomass</td>
<td>~200</td>
<td>~20</td>
<td>~1.3</td>
<td>~20</td>
<td>~2.5</td>
<td>0</td>
<td>0</td>
<td>~6 000</td>
</tr>
<tr>
<td>Biogas</td>
<td>31</td>
<td>19</td>
<td>1.9</td>
<td>36</td>
<td>1</td>
<td>0</td>
<td>640</td>
<td>900</td>
</tr>
<tr>
<td>Liquified Petroleum Gas</td>
<td>200</td>
<td>73</td>
<td>2.7</td>
<td>37</td>
<td>16</td>
<td>0</td>
<td>22</td>
<td>68 000</td>
</tr>
<tr>
<td>Diesel</td>
<td>750</td>
<td>44</td>
<td>12</td>
<td>13</td>
<td>21</td>
<td>3</td>
<td>8</td>
<td>77 000</td>
</tr>
</tbody>
</table>

2.1.2 Vehicle classifications

California is one state in the world that like no other is pushing for cleaner exhaust emissions from vehicles. The California Air Resources Board (CARB) first adopted low emission vehicle (LEV) standards in 1990. These standards run from 1994 through 2003. At CARB’s November 1998 meeting new more stringent regulations, known as LEV II, were amended [6]. The LEV II regulations run from 2004 through 2010. It is not a coincidence that such strict regulations were first introduced in California. Due to the early and massive motorization of that state, automotive pollution such as photochemical smog was early faced. The emission regulations introduced by CARB serve as models for the rest of the world. CARB has defined five classes of passenger vehicles in the LEV II regulations: Transitional Low Emission Vehicle (TLEV), Low Emission Vehicle (LEV), Ultra Low Emission Vehicle (ULEV), Super Ultra Low Emission Vehicle (SULEV) and Zero Emission Vehicle (ZEV).
Their allowed emissions of tailpipe exhaust gases are defined in relation to the minimum federal requirements for passenger cars (the “Tier 1” standard), which are less than 0.25 grams per mile of hydrocarbons, less than 3.4 grams per mile of carbon monoxide and less than 0.4 grams per mile of oxides of nitrogen. There are no specific regulations on the fuel consumption or the carbon dioxide emissions. The definitions of the five low emission vehicles are given in Table II [7]. California, the state of New York and Massachusetts have the TLEV standard as the minimum requirement. The even cleaner vehicles will successively be forced into market by accreditation systems. The goal is to increase the market share for the zero emission vehicles to 11 percent by 2009, 12 percent by 2012, 14 percent by 2015 and 16 percent by 2018.

Table II: Passenger car emissions reductions for LEV II regulations.

<table>
<thead>
<tr>
<th></th>
<th>HC</th>
<th>CO</th>
<th>NOx</th>
</tr>
</thead>
<tbody>
<tr>
<td>TLEV</td>
<td>50%</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>LEV</td>
<td>70%</td>
<td>0</td>
<td>50%</td>
</tr>
<tr>
<td>ULEV</td>
<td>85%</td>
<td>50%</td>
<td>50%</td>
</tr>
<tr>
<td>SULEV</td>
<td>96%</td>
<td>70%</td>
<td>95%</td>
</tr>
<tr>
<td>ZEV</td>
<td>100%</td>
<td>100%</td>
<td>100%</td>
</tr>
</tbody>
</table>

2.1.3 Fuel consumption data on new passenger cars

The average fuel consumption of new passenger cars in Sweden in the time period 1978-1996 is shown in Figure 2.2 [1]. During the period 1978 to 1987 the average fuel consumption for new passenger vehicles decreased from 9.3 liters per 100 km to 8.2 liters per 100 km. However, the average fuel consumption in 1996 had increased to 8.3 liters per 100 km. The reason for this is that the market share of big cars had doubled and that new technology and improved efficiency mainly was used for creating more powerful engines rather than improving the fuel consumption [1]. The passenger cars of today come in a broad range of sizes and engine powers. In Table III below the fuel consumption data of some popular new passenger cars that run on gasoline are shown [8]-[9].

Figure 2.2: Average fuel consumption of new passenger cars in Sweden.
Table III: Fuel consumption data of some popular passenger cars of model year 2004.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Lincoln</td>
<td>Navigator SUV</td>
<td>8</td>
<td>5.4</td>
<td>2WD</td>
<td>Aut. (4)</td>
<td>18.1</td>
<td>13.1</td>
<td>1310 3182 7.2</td>
<td></td>
</tr>
<tr>
<td>Ford</td>
<td>Expedition SUV</td>
<td>8</td>
<td>4.6</td>
<td>4WD</td>
<td>Aut. (4)</td>
<td>16.8</td>
<td>13.1</td>
<td>1118 3050 6.8</td>
<td></td>
</tr>
<tr>
<td>Volvo</td>
<td>XC90</td>
<td>SUV</td>
<td>6</td>
<td>2.9</td>
<td>4WD</td>
<td>Aut. (4)</td>
<td>15.7</td>
<td>11.8</td>
<td>1156 2805 6.1</td>
</tr>
<tr>
<td>BMW</td>
<td>X5</td>
<td>SUV</td>
<td>8</td>
<td>4.4</td>
<td>4WD</td>
<td>Aut. (6)</td>
<td>14.7</td>
<td>10.7</td>
<td>1092 2591 5.8</td>
</tr>
<tr>
<td>Mercedes</td>
<td>S500</td>
<td>Large</td>
<td>8</td>
<td>5.0</td>
<td>2WD</td>
<td>Aut. (7)</td>
<td>14.7</td>
<td>9.8</td>
<td>1034 2499 5.6</td>
</tr>
<tr>
<td>Saab</td>
<td>9-3</td>
<td>Compact</td>
<td>4</td>
<td>2.0</td>
<td>2WD</td>
<td>Man. (6)</td>
<td>11.8</td>
<td>8.4</td>
<td>855 2060 4.6</td>
</tr>
<tr>
<td>Saab</td>
<td>9-5</td>
<td>Midsize</td>
<td>4</td>
<td>2.3</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>11.2</td>
<td>8.1</td>
<td>819 1969 4.4</td>
</tr>
<tr>
<td>Volvo</td>
<td>V70</td>
<td>Midsize</td>
<td>5</td>
<td>2.4</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>10.7</td>
<td>7.8</td>
<td>786 1887 4.3</td>
</tr>
<tr>
<td>VW</td>
<td>Golf</td>
<td>Compact</td>
<td>4</td>
<td>2.0</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>9.8</td>
<td>7.6</td>
<td>663 1775 4.0</td>
</tr>
<tr>
<td>Honda</td>
<td>Civic</td>
<td>Compact</td>
<td>4</td>
<td>1.7</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>7.4</td>
<td>6.2</td>
<td>526 1387 3.1</td>
</tr>
<tr>
<td>Toyota</td>
<td>Corolla</td>
<td>Compact</td>
<td>4</td>
<td>1.8</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>7.4</td>
<td>5.9</td>
<td>497 1357 3.0</td>
</tr>
<tr>
<td>Toyota</td>
<td>Prius</td>
<td>Midsize</td>
<td>4</td>
<td>1.5</td>
<td>2WD</td>
<td>Aut. (4)</td>
<td>3.9</td>
<td>4.6</td>
<td>325 867 2.0</td>
</tr>
<tr>
<td>Honda</td>
<td>Insight</td>
<td>2-seater</td>
<td>3</td>
<td>1.0</td>
<td>2WD</td>
<td>Man. (5)</td>
<td>3.9</td>
<td>3.6</td>
<td>284 765 1.8</td>
</tr>
</tbody>
</table>

1 Definition according to the Environmental Protection Agency (EPA)
2 Based on 15 000 km/year, 50% city and 50% highway driving, Gasoline prices per 5 July 2004, USA: $0.507/L (Regular) and $0.557/L (Premium), Sweden: $1.36/L (Premium)
3 Greenhouse Gas Emissions (Generic term that refers to emissions that cause global warming)

The increasingly popular sport utility vehicles (SUV) have typically very high fuel consumption, due to their high weight and large size (increases air and rolling friction). A normal SUV produces 2-3 times more greenhouse gas emissions per year than the average compact car. The worst SUV, concerning fuel consumption, produces 4 times more greenhouse gas emissions than the best small car, the Honda Insight. The two least consuming cars in Table III, the Toyota Prius and Honda Insight, are hybrid electric vehicles and are described more in section 3.6.3.

2.2 The world’s finite oil resources

The oil resources in the world are finite, which means that the dominating energy carrier of today needs to be replaced in a near future, or expressed more accurately, during the current century. In this section a historic overview of the consumption, production, import and price of oil is given. Data on the proven oil reserves in the different regions of the world is also given. Estimations of when the oil reserves would be empty, under the assumptions that the productions continue at the levels of today and no new oil reserves are explored, are illustrated. All the data in this section refer to the total amount of oil, not only the oil that is limited for use in the transport sector.

2.2.1 Oil consumption

The developed countries consume most of the oil resources in the world. Figure 2.3 shows the yearly oil consumptions in different regions of the world for the period 1965 to 2001 [10]. The domination of oil concerning global energy consumption is illustrated in Figure 2.4 showing the five globally largest energy sources oil, gas, coal, nuclear energy and hydro energy for the period 1965-2001 [10].
2.2.2 Oil production

The Organization of the Petroleum Exporting Countries (OPEC) dominates the world production of oil. The organization has eleven developing state members, which are Algeria, Indonesia, Iran, Iraq, Kuwait, Libya, Nigeria, Qatar, Saudi Arabia, the United Arab Emirates and Venezuela [11]. The countries with the highest production are located in the Middle East. North America is the second largest producer, while South and Central America, Europe, Former Soviet Union, Africa and Asia produce roughly the same amount of oil. The total productions of oil in the world for the period 1965 to 2001 are shown in Figure 2.5 [10].
Figure 2.5: Total yearly production of oil for the period 1965-2001.

In the late 1970s and early 1980s the OPEC countries tightened their oil taps, which immediately resulted in a reduction of the world production and increased oil prices.

2.2.3 Net oil import for largest consumers

The net daily oil imports for the three largest consumers USA, Europe and Japan are shown in Figure 2.6. Europe and Japan have had a relatively stable net oil import in recent years, while the net import for the United States has steadily been rising.

Figure 2.6: Net daily oil import for the three largest consumers by regions in the world.
(Net import = gross import – gross export.)
2.2.4 Oil prices

The averaged monthly spot prices of crude oil for the period January 1955 to July 2004 are shown in Figure 2.7 [12]. The most dramatic increase in oil price to date in relative figures happened during the oil crisis in 1973, with more than 250 percent increase from 1973 to 1974. It is clearly shown in Figure 2.5 and Figure 2.7 that reduced production speed, e.g. in 1973 and 1979, results in immediately increased oil prices. The most abrupt changes have been caused by different wars in the Middle East. The oil crisis in 1973 was due to the Yom Kippur War and the Arab oil embargo. The Iranian revolution was in 1979 and the Iran/Iraq War started in 1980. The Gulf war started in 1990. During the recent years the increase in oil price has been mainly due to the economic growth and the oil demand in China, because of their accelerating export industry. China has an own oil production of approximately 3,5 million barrels per day, roughly the same as for Norway, but the last four years this amount has not been enough to cover the demand. The result is that China stands for half of the increase of the total oil demand. In the last week of July 2004, China’s ministry of trade declared that the oil demand of the country is estimated to increase by 12 percent during 2004 [13]. The price of oil rose above $ 40 per barrel in May 2004.

![Figure 2.7: The spot prices of crude oil for the period January 1955 to July 2004 at the West Texas Intermediate oil market](1 barrel = 0.1364 metric tons = 159 litres).

2.2.5 Oil reserves

The proved oil reserves in the world are primarily located in the Middle East. The reserves controlled by the OPEC countries are relatively much greater than reflected by their production levels compared to the rest of the world. The proved reserves of oil in the world for the period 1981 to 2001 are shown in Figure 2.8 [10]. The oil reserves have historically risen from one year to another due to new findings and improved exploration techniques, but no revolutionary findings of new reserves are expected in the future, because geologists have already investigated the potentially most oil-rich areas [14]. The proved reserves of oil is generally taken to be those quantities that geological and engineering information indicate with reasonable certainty can be recovered in the future from known reserves under existing conditions [10].
Figure 2.8: Proved oil reserves for the period 1981-2001.

If the oil productions in the world would continue at constant levels as for the year 2001 and the oil reserves do not increase, the world’s total reserves would be empty in the 2040s. The oil reserves for the three regions with highest consumption, North America, Europe and Asia Pacific, would all be empty before 2020, Europe already in 2009. The estimations for different regions are shown in Figure 2.9.

Figure 2.9: Estimated future total oil reserves, based on known reserves under existing conditions, if the oil productions would continue at constant levels as for the year 2001.

The market share of oil production for the OPEC states will quite probably hit 50 percent by 2010. Thus radical increases in oil prices can be expected [14]. The share of the total proven
oil reserves controlled by the OPEC countries was 78% at the end of 2001 [10]. Within OPEC the states with the highest shares are Saudi Arabia (25%), Iraq (11%), United Arab Emirates (9%), Kuwait (9%), Iran (9%) and Venezuela (7%).

The world’s proved gas and coal reserves would for comparison last 62 and 216 years, respectively, if the productions would continue at constant levels calculated from and including the year 2001.

### 2.3 Environmental and humanistic issues

#### 2.3.1 Global warming

##### 2.3.1.1 Introduction

The main factor for global warming and climate change is the increasing amount of carbon dioxide in the atmosphere that contributes to the greenhouse effect and rising temperatures. The UN Intergovernmental Panel on Climate Change projects the following climate changes [15]:

- The globally averaged surface temperature is projected to increase by 1.4 to 5.8 °C in the 21st century.
- The projected rate of warming is much larger than the observed changes during the 20th century and is very likely to be without precedent during at least the last 10,000 years.
- Global mean sea level is projected to rise by 0.09 to 0.88 meters between 1990 and 2100.
- Global mean surface temperature increases and rising sea level from thermal expansion of the ocean are projected to continue for hundreds of years after stabilisation of greenhouse gas concentrations (even at present levels).

##### 2.3.1.2 The Kyoto Protocol

More than 160 nations met in Kyoto, Japan, from December 1 through 11 in 1997 to negotiate binding limitations on greenhouse gases (primarily carbon dioxide) for the developed nations [16]-[17]. The decision-making was in accordance with the objectives adopted at the Earth Summit Conference in Rio de Janeiro in 1992. The outcome of the meeting in Kyoto was the Kyoto Protocol, in which the developed nations agreed to limit their greenhouse gas emissions, relative to the levels emitted in 1990. Each nation must individually decide how to implement their reduction goal during a five-year period, between 2008 and 2012. The United States for instance, the largest emitter of greenhouse gases by nation, agreed to reduce emissions from 1990 levels by 7 percent during this period.

##### 2.3.1.3 Historical global temperature rises

Research on historical climate changes is carried out by studying temperature indicators embedded in ancient layers of Arctic ice. Such data show that a number of dramatic shifts in average temperature took place in the past with rapid speed, in just a few years in some cases [18]. Scientists are not sure what caused the warming that triggered such incredible changes in the remote past, but obviously it was not due to human activity. The data from the Arctic ice and other sources suggest the atmospheric changes that preceded earlier historical collapses
were very similar to the global warming of today. One example happened 13000 years ago as the ice age began drawing to a close. Then the temperatures in Greenland rose to levels near those of recent decades before they abruptly plunged, ushering in the “Younger Dryas” period, a 1300-year reversion to ice-age conditions. (A dryas is an Arctic flower that flourished in Europe at the time.) Icebergs appeared as far south as the coast of Portugal during this period. Another abrupt climate change happened 8200 years ago, which resulted in a century of cold, dry and windy weather across the Northern Hemisphere. An era called the “Little Ice Age” had hard winters, violent storms, and droughts between 1300 and 1850. This period was mild compared with the “Younger Dryas” era, but caused devastating famines [18].

The variations of the Earth’s surface temperature over the past 140 years are shown in Figure 2.10, as departures in temperature from the 1961 to 1990 average. The temperature data is collected by the use of thermometers and goes back as far as possible, concerning documented calibrated modern instruments. Over both the last 140 years and the last 100 years, the best estimate is that the global average surface temperature has increased by $0.6 \pm 0.2 ^\circ C$, according to the United Nations [19].

![Figure 2.10: Variations of the Earth’s surface temperature over the past 140 years. The temperature is shown year by year (red bars) and approximately decade by decade (black line, a filtered annual curve suppressing fluctuations below near decadal time-scales). There are uncertainties in the annual data (the black whisker bars represents a 95% confidence range).](image)

The variations of the average surface temperature of the Northern Hemisphere over the past millennium are shown in Figure 2.11, as departures in temperature from the 1961 to 1990 average. The temperature data is reconstructed from proxy data calibrated against thermometer data. The main proxy data are from tree rings, corals, ice cores and historical records. The uncertainties of the statistical temperatures increase in more distant times, due to the use of relatively sparse proxy data. The rate and duration of the warming in the 20th century has been far greater than in any of other nine centuries in the past millennium. The average surface temperature was actually slightly declining in the period 1000 to 1900 before it rapidly rose in the last century, as shown in Figure 2.11. It is likely that the 1990s have been the warmest decade and 1998 the warmest year of the millennium [19].
Figure 2.11: Variations of the average surface temperature of the Northern Hemisphere over the past millennium. The temperature is shown year by year (blue curve) and as a 50 years average (black curve). The 95% confidence range in the annual data is represented by the grey region. The red curve represents data collected by modern thermometers.

2.3.1.4 Global temperature rise of today

An international panel of climate experts concluded in 2001 that there is increasingly strong evidence that most of the global warming observed over the past 50 years is due to human activities [18]. The National Academy of Sciences issued a report in 2002 concluding that human activities could trigger abrupt climate change. The burning of fossil fuels such as oil and coal, which release heat-trapping carbon dioxide, is the main cause behind the temperature rise. Global warming indicators include shrinking Arctic ice, melting alpine glaciers, and remarkably earlier springs in northern regions. The eastern North America and northern Europe are warmed by a huge Atlantic Ocean current that flows north from the tropics, the Gulf Stream. The current pumps out warm moist air and gets cooler and denser on its way north. That causes the current to sink in the North Atlantic, where it heads south again in the depths of the ocean. The sinking process draws more water from the south, keeping the roughly circular current on the go, and functions like a heat pump. When the climate warms however, fresh water from melting Arctic glaciers flows into the North Atlantic, lowering the salinity of the current and thereby its density and tendency to sink, according to the theory of the climate scientists. A higher average temperature also increases rainfall and runoff into the current, which further is lowering its saltiness. As a result, the pump loses its main motive force and can rapidly collapse and alter the climate over much of the Northern Hemisphere. A total shutdown of the Gulf Stream apparently resulted in the "Younger Dryas", while it was temporarily slowed down in the “Little Ice Age”, according to theory. Historical information
about climate changes can therefore be used to foresee possible scenarios, but it is not possible to give exact predictions. It makes sense to focus on a midrange case of severity, like the era that came on 8 200 years ago, for planning purposes [18].

### 2.3.1.5 Migration problems

A lot of land area around the world populated by millions of people is just above sea level. Much of Bangladesh with 141 million people (summer 2004) is one example of such land area. Another European example is the Netherlands. If the sea level rises, this means that millions of people are left without a home. The then wealthier nations are likely to be invaded by refugees. The cause of migration is not only caused by sea level rise, though. During global warming the cold regions are getting colder while warm regions are getting warmer. This means increasing areas of ice glaciers in some parts and increasing areas of deserts in other, result in dramatic migration and movements or changes in wildlife.

### 2.3.1.6 National security issues

When millions of refugees are seeking new places to live, extreme social and economic pressure can be expected in the countries receiving the immigrants. Many countries already have big diversities within their nation and towards their neighbours. Imagine nuclear powers like Pakistan, India and China put under even more pressure when their neighbours in Bangladesh and their own inhabitants need to find new homes. The gap between the rich and the poor countries will probably increase and conflicts may arise in numbers never before seen. Terrorism is a big concern today, but the question is what will happen in future potential societies as described above? The Pentagon has become interested in abrupt climate change and its implications on national security issues. An unclassified report from the Pentagon was completed in 2003 and was shared with the Fortune Magazine, which in turn published an article in January 2004 based on this report [18].

### 2.3.1.7 Plausible future climate scenarios

Some plausible future climate scenarios are severe [18]. The average temperature may fall by up to 3 °C in some regions of North America and Asia and even more in parts of Europe. (The average temperature over the North Atlantic was by comparison 5-8 °C lower than it is today during the last ice age.) Massive droughts can start in key agricultural areas and cause famines. The average annual rainfall may drop by 30% in northern Europe and result in a climate like Siberia. (Siberia is actually on the same latitude as the north of Europe, but is not heated by the Gulf Stream.) Violent storms can appear more frequent. In the United States, especially in the south, heavy droughts along with winds that are 15% stronger on average than now can cause widespread dust storms and soil loss. China may be hit by unpredictable monsoon rains, which can cause massive floods and destruction. The same apply for other parts of Asia and eastern Africa. Bangladesh can become inhabitable, due to a rising sea level. Accumulated data over the past decade suggest that the plausibility of abrupt climate change is higher than most of the scientific community, and perhaps all of the political community, are prepared to accept.

### 2.3.1.8 Political action plan

The World Economic Forum held in 2003 in Davos, Switzerland, included a session at which the director of the Woods Hole Oceanographic Institution in Massachusetts, Robert Gagosian, urged policymakers to consider the implications of possible abrupt climate change within two decades [18]. The climate change issue should be elevated beyond a scientific debate. The
Kyoto Protocol is a good starting point for action. Unfortunately, the largest emitter of greenhouse gases, the United States, has withdrawn from the agreement. President George W. Bush and his administration has rejected the Kyoto Protocol, with the motivation that it will hurt U.S. economy. Actions on different measures now matters, because it may reduce the likelihood of the dramatic climate change to take place, and governments and people can be better prepared if it does. More resources should consequently be spent on climate research and strategies for different potential scenarios should be identified. Policymakers should be bold enough to legislate tougher and more stringent fuel-economy standards for new passenger vehicles.

2.3.2 Pollution and health aspects

In addition to global warming, emissions from vehicles also cause allergic problems and cancer. Mexico City and Bangkok among other big cities suffer from serious smog problems that are unhealthy for every living organism. Athens has for instance experienced erosion of their antique monuments in Acropolis due to smog-particles. Acid rain is responsible for forest death.

2.3.3 Population growth

The world population is estimated to grow by 50 percent in the next 50 years from 6.1 billion people in mid 2001 to 9.3 billion people by 2050 according to the United Nations Population Fund [20]. At the same time it is expected that the developing countries will increase their standard of living. This is especially evident in China, where multinational companies are already building up factories for high scale production of advanced consumer products, e.g. mobile phones. These production lines are often replacing lines that are shut down in Europe and USA, which are owned by the same companies. The reason for the export of jobs is of course the lower salaries in the developing countries. When people in the developing countries get higher skills in production of consumer goods, their income will probably increase over time. Their need for transportation to work and for their spare time must, as a result, also increase. The number of vehicles in the developing countries is therefore likely to rise rapidly in the 21st century.

2.3.4 GNP and its impact on energy consumption

The developments of the gross national products (GNP) and the energy consumptions within the transport sector in the period 1971-2000 for the OECD- (Organization for Economic Co-operation and Development) countries and the non-OECD-countries are shown in Figure 2.12 [1]. There has historically been a strong correlation between economical growth and energy consumption. This rule of thumb was slightly weakened in the 1970s and early 1980s when the rising oil prices contributed to more energy efficient systems in the developed countries. In the non-OECD countries on the contrary the energy demand has recently increased more than the GNP. Taking into account that the economical growth is expected to continue with accelerating speed in some countries, especially in Asia with large populations, the energy demand in the transport sector can be expected to rise rapidly.

In 1960 fewer than 4 percent of the world’s population possessed vehicles. By 1980 this share had reached 9 percent, and currently 12 percent are vehicle owners [21]. In 2020 as many as 15 percent of the people on the planet could have a vehicle, based on the present growth rates. The world’s population may climb from around 6 billion today to nearly 7.5 billion in two decades. The number of vehicles on the planet could, as a result of the combined economical
and population growth, rise from about 700 million to more than 1.1 billion [twenty]. The United States, Europe and Japan currently have a 75 percent share of all automobiles. However, it is expected that more than 60 percent of the increase in new vehicle sales during the next decade will occur in eight emerging markets consisting of China, Brazil, India, Korea, Russia, Mexico, Poland and Thailand [21].

Figure 2.12: Development of GNP and energy consumption within the transport sector for the (a) OECD-countries and (b) the non-OECD-countries.

2.3.5 Tax regulations

Tax regulations are powerful tools to change people’s preferences. Diesel engines in passenger vehicles are for instance very popular in Europe, due to favourable taxes compared to the rest of the world. The relatively high prices of gasoline and diesel in Europe contribute to the choice of fuel. Since the diesel engine is more efficient than the Otto-engine, the long-distance drivers often prefer the diesel engine because of reduced fuel cost. The reduction in fuel cost can then motivate the choice, despite of the slightly higher purchase price. In countries with relatively low fuel prices, like in the United States, the gain is almost negligible for normal driving, and the diesel engine is therefore not usual in passenger vehicles. The gasoline pricing in the United States and Sweden is shown in Figure 2.13 for comparison [9]. The taxes account for 21% of the gasoline price in the United States and 70% in Sweden. The high tax level on gasoline in Sweden is not only of environmental concern, but more of a political tax regularisation principle used to finance other parts of the welfare system. The transport need is generally insensitive to changes in the transport cost. Research has shown that a ten percent increase of fuel prices only results in a one to three percent decrease of traveling [1]. The explanation behind this behaviour can be that few people have the possibilities to change their traveling pattern, once they have chosen a home and a workplace. The fuel cost is also just a fraction of the total cost of owning a new vehicle. Increased fuel cost can give incentives to the vehicle manufacturers to produce more efficient vehicles though.
Figure 2.13: The gasoline prices in the United States and Sweden shown in absolute and relative values split into different cost posts based on the gasoline prices in May 2004 (USA: $/L 0.52, Sweden: $/L 1.36), assuming equal production costs.

2.3.6 People awareness and opinion change

People in the developed countries are in general aware of the environmental changes like global warming and pollution caused by human activity from the transport sector. Nevertheless, it has been proven hard to change habit or slightly reduce the standard of living for the individual, even if it could result in less harm to the environment. Sport utility vehicles (SUVs), pickup trucks and minivans account for nearly 50 percent of all new vehicles sold in California today. These cars have high price tags, an average weight of 2-3 tonnes and a gasoline consumption of around 15 litres per 100 km. The medium-duty vehicles in this group emit 50 to 150 percent more smog-forming pollution than the typical passenger car [6]. This market is very attractive for the business driven car industry and more and more players are entering with new models. Even the brands that base their image on safety and environment, like Volvo and Volkswagen, are now for the first time offering models in this segment. The low prices on gasoline in the USA and a relatively wealthy middle class that prefer to drive big cars are not changing the trend. Experience has shown that consumers are not willing to buy a vehicle with poorer performance (regarding comfort and motor power) than their previous. Market economy and individual responsibility does not seem to be a good combination towards cleaner vehicles. The auto manufacturers should ideally take a much stronger responsibility and produce more efficient vehicles, but at the same time vehicles that are good-looking and fun to drive. First then the opinion could change among many people, who now more or less equate environmental friendly vehicles with golf cars.

<table>
<thead>
<tr>
<th>Cost Post</th>
<th>USA</th>
<th>Sweden</th>
</tr>
</thead>
<tbody>
<tr>
<td>Taxes</td>
<td>0.11</td>
<td>0.95</td>
</tr>
<tr>
<td>Distribution and marketing</td>
<td>0.04</td>
<td>0.04</td>
</tr>
<tr>
<td>Refining</td>
<td>0.16</td>
<td>0.16</td>
</tr>
<tr>
<td>Crude oil</td>
<td>0.21</td>
<td>0.21</td>
</tr>
</tbody>
</table>
2.3.7 Trends in transport technology development

2.3.7.1 Alternative vehicle propulsion systems

Big challenges are facing the car manufacturers in the near future. Increasing oil prices and tougher regulations on permitted exhaust gases will force manufacturers to do a technology change. The emission reduction requirements for the TLEV, LEV and ULEV classes can be achieved with normal internal combustion engines, but they have to improve significantly. The reductions can be achieved with more optimised combustion, more efficient catalytic converters or blended with alcohol fuels from cellulosic crops. Alcohol can be blended with gasoline up to 15-20 percent by volume and used in current vehicles. These fuels may never replace petroleum fuels completely, because of the vast amounts of land required for growing the crops, but they could eventually replace up to 10 percent in some countries and reduce CO\textsubscript{2} emissions by up to nearly 10 percent [22]. The SULEV class has even tougher requirements that will be extremely difficult to achieve with conventional drivelines. All major automakers are therefore also developing alternative drivelines. The most promising driveline in a not too far away future is the hybrid electric driveline, which consists of two power sources, an internal combustion engine and an electrical machine. The electrical machine normally handles transient power variations and helps the engine to operate more constantly whereby higher efficiency and lower emissions can be achieved. The emission goals in the SULEV class can therefore be achieved with hybrid electric vehicles (HEVs). There are already some HEVs on the market; both Toyota and Honda are offering models. The best-selling hybrid, Toyota Prius, has been sold in close to 200 000 units, since its introduction in Japan in 1998.

Some vehicle engines are designed for compressed natural gas (CNG) and a few manufacturers offer models with CNG drivelines with methane fuel (e.g. Fiat and Opel), which reduce emissions but require new networks of distribution channels [2]. A third alternative driveline is the pure electric, where the energy source is either a battery or a fuel cell. In the case of the battery source, the vehicle will have no tailpipe emissions and is consequently classified as a zero emission vehicle. The electric power needs to be charged from the grid. Depending on how the electric energy is generated in the particular region, this will add to the power plant emissions if not generated by renewable energy sources. In California, where a lot of the electric energy is produced from fossil fuels, battery-powered electric vehicles reduce pollutants by more than 90 percent when compared to the cleanest conventional gasoline-powered vehicles, even when including the emissions from power plants generating the electricity to charge the electric vehicles [23]. The drawbacks of the battery EV are the relatively short driving range (around 150 km) and high cost (partly due to low production volumes). The fuel cell powered EV in its cleanest form produces no pollution when tanked with pure hydrogen and has ZEV status. A fuel cell converts chemical energy directly into electricity by combining oxygen from the air with hydrogen stored in the tank. Therefore it does not need to be charged from the grid as long as hydrogen is supplied. Some vehicles are equipped with a fuel reformer that converts hydrocarbon fuels, e.g. methanol, natural gas or gasoline, into a hydrogen-rich gas and will have some low emissions. Auto manufacturers such as DaimlerChrysler, General Motors and Toyota have announced plans to commercialize fuel cell-powered electric vehicles [24].

2.3.7.2 The car of the future

The hydrogen fuel cell car could be the car of the future, without compromising the environment or depleting the earth’s natural resources. This car could revolutionize the car industry in more than one way. Since it has some unique characteristics, it can be built
smarter, more flexible, with increased modularity, with better comfort, with dramatically new designs and even allow for increases in the profit margins within the automobile business [21]. The car of today with predominately mechanical systems for steering, braking, throttling and for transmission parts could be replaced by drive-by-wire technology, which instead consists of fully electronically controllable units. This replacement would dramatically free up space, since electronic systems tend to be less bulky than mechanical ones. Moreover, since electronic units are connected through cables and not mechanical transmissions the different units can be distributed freely in the vehicle. A fuel cell powered car, in opposite to the conventional internal combustion engine powered car, enable the use of a flat chassis with an ideal weight distribution and low center of gravity for improved road handling, which also frees up space for the body interior and creates design opportunities that are not available today.

The auto industry of today is very capital-intensive, with modest profit margins. The competition and global excess production capacity is driving down prices on vehicles. At the same time is the steadily tougher regulatory standards pushing up costs. The combination of lower prices and higher costs are threatening profit margins. A fuel cell car concept as described above, however, could significantly change the current business model. With the use of modules that can be produced independently it could considerably reduce the vehicle development costs. The chassis, which is the most costly part of the vehicle, could be used on several different body styles and thereby increase the production volume and reduce the cost.

The preferred performance and drivability of a vehicle could be limited to a pure software control issue using the same hardware as in neighbouring types of vehicles. General Motors has recognized these design opportunities and come up with a “skateboard chassis” concept called AUTOnomy, which was introduced in Detroit in the spring of 2002 [21]. The concept is illustrated in Figure 2.14 and gives a glance of a potential and beneficial outline of future propulsion equipment.

*Figure 2.14: The fuel cell powered concept car from General Motors called AUTOnomy.*
In September 2002 GM also presented a drivable prototype based on the same concept at the Paris Motor Show. The prototype was called Hy-wire (for hydrogen by-wire) [21].

### 2.3.7.3 Infrastructures of hydrogen societies

A project called HyNor initiated by Norsk Hydro is working on building an infrastructure that can become the world’s longest road equipped with Hydrogen tanking stations [25]. The six initial stations along the road are planned to be opened in 2008 and their locations are shown on the map of the south of Norway in Figure 2.15. The distance from Oslo to Stavanger along this road is close to 600 km. California is also working for a large network of tanking stations by 2010 [25]. This is also true for Vancouver in British Columbia, Canada, who will host the Olympic Games in 2010. A big fleet of fuel cell powered buses will contribute to the transport during the Olympics. The Canadian vision is a “Hydrogen Highway” that at first will cover the distance from Vancouver to Whistler. Not only the government is interested in this project, but also companies like Hydrogen British Columbia and Ballard Power Systems [25]. In a longer perspective, California and British Columbia can interlink their two regions with Hydrogen tanking stations and thereby establish a market with 50 million people. The country that currently has the most profound plans towards a hydrogen society is Iceland. A joint-venture company called Icelandic New Energy Ltd. has been founded by the owners Vistorka hf (EcoEnergy), DaimlerChrysler AG, Norsk Hydro ASA and Shell Hydrogen B.V. [26]. The purpose of the company is to investigate the potential for eventually replacing the use of fossil fuels in Iceland with hydrogen based alternatives and create the world’s first hydrogen economy. The reasons for choosing Iceland for this investigation are several. Iceland has similar standards and transportation systems as most other developed countries and the results can therefore easily be adapted elsewhere. Iceland also has an electricity production almost entirely based on renewable energy sources, which may be used for hydrogen production. The evaluation of the new technology must be carried out under severe weather conditions and Iceland is idealistic from this point of view. The government of Iceland has furthermore announced that it is aiming to transform Iceland into a hydrogen society in the near future. The transformation could possibly be completed in the years 2030-2040 [26].

![Figure 2.15: A long road with hydrogen tanking stations planned opened in 2008 in Norway.](image)

### 2.3.8 Future electrical energy generation

The annual electrical energy consumption has continuously been rising in the past and is also expected to do so in the next decades, as illustrated in Figure 2.16, which is estimated by Shell with a growth rate of 2 % per year.
The consumption of today is expected to rise by a factor of two by 2040 and by a factor of three by 2060. New types of electrical energy generation will drastically increase in importance and open up new markets for suppliers of electrical equipment. These suppliers should have an obligation in creating environmentally friendly products, which means high efficiency products and sustainable material use. Some examples of renewable electrical generation methods are shown in Figure 2.17. The pollution-free solar energy plant shown in Figure 2.17.b is an example of a vision proposed by Sanyo to cover 1/3rd of China’s energy demand in 2030 by these farms along the historic Silk Road [27]. The project is called the “Silk Road GENESIS Project” and seven companies are taking part in it. The first plan is to build solar power plants at 139 locations in Western China around 40 degrees north latitude forming a chain running east to west. The solar plants, each with a capacity of 100 megawatts, are planned to be built a few at a time between 2001 and 2020. One station would cost some ¥10 billion to build so total investment would be about ¥23 trillion by 2020. Still, the planners predict revenues from the generated electricity would make it possible to pay back the investment and earn roughly ¥140 billion in profits in 2020 [28]-[29].
2.4 Summary

Several factors that influence motor vehicle design have been discussed in this chapter, including oil resources, regulations, pollution, health aspects, global warming and more. It is proven and claimed by prestigious organizations like the United Nations that vehicles of today have negative impacts on the global environment. Since the total number of vehicles is expected to increase, due to population rise and economy rise in the developing countries, this can actually lead to increased global emissions despite the fact that the engines are getting cleaner. Advocates of the auto industry often claim that the zero emission vehicles produce emissions at the power plant instead. That is however not for the car manufacturers to solve. A lot of research is focused on renewable energy sources such as wind power, solar power and geothermal power. Iceland is for instance leading the way towards a hydrogen society. A lot more needs to be done in order to maintain the welfare in the developed countries, when the fossil resources slowly start to vanish. When taking into account the projected increase in average global temperature and climate change, health aspects and the rapidly diminishing oil reserves in the world, it is urgent that we now begin the transition towards a society based on sustainable transportation. The hydrogen fuel cell powered car could be the car of the future that possesses the required characteristics. There are some obstacles in the way for an early commercialization, including the expensive catalysts used today and the lack of infrastructure, though. The hybrid electric vehicle is a natural candidate for making the transition from the conventional car towards the car of the future.
3 Fundamentals of road vehicles

3.1 Physics of vehicle motion

The total running resistance (force) on a vehicle in motion is given by the expression [30]

\[ F_x = (m + m_j) \ddot{x} + mg (f_r + \sin \alpha) + \frac{1}{2} c_x A \rho \dot{x}^2, \]  

(3.1)

where \( m \) is the mass of the vehicle, \( g \) is the acceleration of gravity, \( f_r \) is the rolling friction coefficient (0.013 for pneumatic car tires on asphalt), \( \alpha \) is the angle of the slope, \( c_x \) is the air drag coefficient in the direction of movement (~0.35 for a modern passenger sedan car), \( A \) is the exposed cross-sectional area of the vehicle, \( \rho \) is the mass density of air (1.226 kg/m\(^3\) at 15°C and 1,013 bar), \( \dot{x} \) is the velocity and \( \ddot{x} \) is the acceleration of the vehicle, respectively, and \( m_j \) is the equivalent mass of the rotating transmission components as given by

\[ m_j = \frac{J_{\text{wheel}} + U_{\text{final drive}}^2 \cdot J_{\text{drivetrain}} + U_{\text{final drive}}^2 \cdot U_{\text{drivetrain}}^2 \cdot J_{\text{engine}}}{r_{\text{wheel}}^2}, \]  

(3.2)

where \( J_{\text{wheel}}, J_{\text{drivetrain}} \) and \( J_{\text{engine}} \) is the inertia of the wheels, the different rotating components in the drivetrain and the engine, respectively. \( U_{\text{final drive}} \) and \( U_{\text{drivetrain}} \) is the gear ratio of the final drive and the drivetrain, respectively, while \( r_{\text{wheel}} \) is the rolling radius of the wheel. When the vehicle is running at constant speed, the first term on the right-hand side in (3.1) is zero and the forces acting on it in this condition are illustrated in Figure 3.1.

\[ \frac{1}{2} c_x A \rho (dx/dt)^2 \]

\[ mg \sin \alpha \]

\[ \frac{1}{2} mg f_r \]

\[ \frac{1}{2} mg f_r \]

\( \alpha \)

\( \text{Centre of gravity} \)

\( mg \)

\( \text{Figure 3.1: The resistance forces acting on a vehicle running up a slope at constant speed.} \)
The force developed by the vehicle at the wheels can be related to the engine power by

\[ F_{\text{wheels}} = \eta_{\text{transmission}} \cdot \frac{60}{2\pi r_{\text{wheel}}} \cdot \frac{P_{\text{engine}}}{n_{\text{engine}}} = \eta_{\text{transmission}} \cdot \frac{P_{\text{engine}}}{\dot{x}}, \]

(3.3)

where \( \eta_{\text{transmission}} \) is the efficiency of the transmission, \( U \) is the gear ratio (the ratio between the engine speed and the wheel speed), \( P_{\text{engine}} \) is the output power and \( n_{\text{engine}} \) is the speed of the engine, respectively.

A discussion about the required and available power under different working conditions for a mid-sized typical European family car is given below. Its main data are given in Table IV and its engine power for different gear steps and required power for running at constant speed versus the vehicle speed are given in Figure 3.2. The total transmission efficiency is chosen to be 0.90. It is evident by studying Figure 3.2 that the difference between the required power when running at constant moderate speed and the available power from the engine is huge. This power gap can be used for accelerating the vehicle. The engine angular speed range is in the interval 750-6500 rpm, with its maximum power at 5500 rpm.

**Table IV: Main data of a mid-sized typical European family car.**

<table>
<thead>
<tr>
<th>Vehicle</th>
<th>Engine</th>
<th>Transmission</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mass (incl. driver)</td>
<td>1500 [kg]</td>
<td>Maximum power</td>
</tr>
<tr>
<td>Air drag coefficient</td>
<td>0,35 [*]</td>
<td>Maximum torque</td>
</tr>
<tr>
<td>Front area</td>
<td>1,86 [m²]</td>
<td>Number of cylinders</td>
</tr>
<tr>
<td>Rolling wheel radius</td>
<td>0,315 [m]</td>
<td>Cylinder volume</td>
</tr>
</tbody>
</table>

**Figure 3.2: Engine power versus vehicle speed for different gear steps and the running resistance force versus vehicle speed for different slopes at constant speed.**

The required powers for running at constant speed on a flat road versus vehicle speed for different types of cars are shown in Figure 3.3. At 100 km/h the required power is 29 kW for a typical SUV, while it is 14 kW for a typical midsized car. The required power increases almost cubically with the speed at higher speeds, because the air friction is then the dominating friction force component.
3.2 The ideal motor

A motor in a vehicle produces its maximum torque \( T_{\text{max}} \) at the angular speed \( \omega_{\text{r, max}} \) and its maximum power \( P_{\text{max}} \) at the angular speed \( \omega_{\text{p, max}} \). The vehicle is equipped with an ideal continuous variable transmission that is connected to the wheels with a rolling radius \( r_{\text{wheel}} \). The gear ratio of the transmission versus the vehicle speed is given by

\[
U(\dot{x}) = \frac{\omega r_{\text{wheel}}}{\dot{x}}, \tag{3.4}
\]

and the accelerating force against the road is given by

\[
F_{\text{wheel}} = F_x = \frac{T U(\dot{x})}{r_{\text{wheel}}} = \frac{\omega T}{\dot{x}}. \tag{3.5}
\]

The maximum value of \( F_{\text{wheel}} \) gives the maximum instantaneous acceleration. This condition occurs at an arbitrarily chosen vehicle speed when the following expression is true

\[
\omega T = P_{\text{max}}. \tag{3.6}
\]

The ideal gear ratio of the transmission that gives the best acceleration is found by combining (3.4), (3.5) and (3.6) and is expressed by

\[
U(\dot{x}) = \frac{\omega P_{\text{max}} r_{\text{wheel}}}{\dot{x}}, \tag{3.7}
\]

which means that the transmission should adjust the gear ratio so that the motor operates at the angular speed where it has its maximum power. Inversely, this means that the ideal motor,
from an acceleration point of view, is the one that delivers maximum power in the whole speed range. It is assumed that the vehicle described above is standing still on a flat road and starts to perform an acceleration test. At low speed the air friction is negligible and all the driving force is available for the acceleration and the road friction. However, there is a limiting factor on how high the driving force is allowed to be before the wheels start to spin, due to the available friction between the tires and the road. The friction coefficient is the ratio of the tangential to the normal force components acting on a tire at the connection point to the ground surface. The available friction coefficient for a normal tire is 0.6-0.9 on dry, 0.3-0.5 on moist and 0.2 on wet asphalt, respectively [30]. When the vehicle speed increases (> ~50 km/h), the air drag force is influential and rises quadratically with the speed, i.e. the equivalent power need due to air drag increases cubically with the speed. It is therefore advantageous that the motor delivers maximum power from this speed. The ideal motor for a passenger car, regarding acceleration, consequently has the performance characteristic as illustrated in Figure 3.4.

![Figure 3.4: The ideal motor characteristic for a passenger car regarding acceleration.](image)

When comparing Figure 3.2 and Figure 3.4 it is obvious that a transmission with a variable gear ratio is needed using an internal combustion engine for propulsion. However, an electrical machine with field-weakening capabilities can achieve such motor specifications with only a single-step gearbox.

### 3.3 Internal combustion engines

The internal combustion engine (ICE) is the most frequently used power source for motor vehicles [31]. Its torque-speed characteristic is relatively constant, which means that its output power is approximately proportional to the angular speed of the engine shaft and that the maximum power is achieved at high revolutions. A variable transmission is therefore needed between the engine output shaft and the wheels in order to achieve satisfactory acceleration power at low vehicle speed. Internal combustion engines produce power by converting chemical energy from the fuel into heat, and the heat is then converted into mechanical work. The chemical energy conversion into heat is accomplished through combustion, while the following thermal energy conversion into mechanical work is achieved by allowing the heat energy to increase the pressure within a medium, which then performs work as it expands.
Liquids or gases are used as working media. A liquid can supply an increase in working pressure via a change of phase or vaporization, while a gas can increase the working pressure through compression. The fuels are mainly hydrocarbons, which require oxygen in order to burn. The oxygen is usually supplied as a constituent of the intake air. The process is called internal combustion if it takes place in the cylinder itself and here the combustion gas itself is used as the working medium. The combustion process is called external combustion if it takes place outside the cylinder [31]. The former process is normally used in motors for vehicles, like in the spark-ignition or Otto engine and in the Diesel engine.

### 3.3.1 Otto engine

The German inventor Nikolaus August Otto first described the Otto engine that uses a four-stroke cycle, in 1876. The engine that he got patented uses gasoline as its fuel source. The patent was invalidated in 1886 when it was discovered that a French inventor, Alphonse Beau de Rochas, had already described the four-stroke cycle principle in a privately published pamphlet dated 1862 [32].

In a gasoline four-stroke engine the cycle begins with the supply of a gasoline and air mixture in the cylinder as the piston moves down on the first stroke. Moving up on the second stroke, the piston compresses the mixture in the top of the cylinder and an electric spark ignites the mixture. The gas explosion that is generated forces the piston down on its third power stroke and on the fourth stroke the piston expels the burned gases from the cylinder into the exhaust. The air-fuel mixture in the Otto engine is compressed to approximately 20-30 bars on the compression stroke, to produce a final compression temperature of 400-500°C. This must be below the automatic ignition level of the mixture and it has to be ignited by a spark shortly before the piston reaches the top dead centre (TDC) [31]. The compressed mixture must be reliably ignited by the ignition system at a precisely defined instant, even under dynamic operation. Otherwise the risk of combustion knock or pre-ignition increases, which occurs when compressed end gas reaches ignition temperature. The gas then burns instantaneously without regular flame propagation and releases very large amounts of heat. This can cause very high local temperatures that can create extreme loads on the engine components and damage them.

### 3.3.2 Diesel engine

The German inventor Rudolf Diesel published a paper in 1893 entitled “The Theory and Construction of a Rational Heat Engine”. The paper presented an engine in which air is compressed by a piston to a very high pressure, causing a high temperature, and fuel (diesel) is then injected into the cylinder as is self-ignited by the high compression temperature.

The diesel engine uses like the Otto engine a four-stroke cycle process, but it has no electric spark ignition system. On the first stroke only air is supplied through the intake valves and on the second stroke the air is powerfully compressed. Shortly before the third stroke diesel is directly injected into the compressed air and the fuel-air mixture ignites spontaneously, which results in a high force on the piston. The fourth stroke is like in the Otto engine, i.e. the piston expels the burned gases from the cylinder into the exhaust.

During the compression stroke intake air is compressed to 30-55 bars in naturally aspirated engines or 80-110 bars in supercharged engines, so that the compression temperature increases to 700-900°C. This temperature is sufficiently high to start the automatic ignition in the fuel injected into the cylinder just before the end of the compression stroke, as the piston reaches TDC [31]. Since the diesel engine relies on auto-ignition, highly volatile fuel is required. The engine must furthermore be designed for operation with high peak pressures,
which means that its materials and dimensions must be selected accordingly, and it therefore tends to be more robust and heavier than the Otto engine. The diesel engine must also operate with excess air (lean) at full throttle, so it generally has lower specific power outputs than their spark-ignition counterparts [31]. Soot is an inevitable byproduct of the combustion process that takes place in the diesel engine. It has recently proven possible to reduce the particulate emissions from modern diesel engines to below the threshold of visibility, e.g. by introducing particulate filters, but it is still a big concern, especially in cities with high population and vehicle concentrations. The diesel engine is more efficient than the Otto engine and emits therefore less carbon dioxide.

### 3.3.3 Engine configurations

The working chamber or cylinder of a reciprocating-single-piston engine is formed by a cylinder head, a cylinder liner, and a piston [31]. The reciprocating-piston engine is completely dominating in vehicles. There also exist a rotary-piston engine, called a Wankel engine named after its inventor, but this type is very unconventional and is at the moment only commercially explored in a Mazda sport car. Some common engine types are the in-line engine, where the cylinders are arranged consecutively in a single plane, the V-engine, where the cylinders are arranged in two planes in a V configuration, and the opposed cylinder or boxer engine, where the cylinders are horizontally opposed. These engine types are illustrated in Figure 3.5.

![Figure 3.5: Some illustrations of common engine types used in commercial passenger vehicles. (a) In-line engine. (b) V-engine. (c) Opposed cylinder or boxer engine.](image)

The in-line engine is the most frequently used engine type in passenger cars and the major models in the small and medium sized vehicles use this topology with four cylinders in a row. The V-engine configuration is used in engines with a high number of cylinders in order to limit the axial length of the engine. Up to 12 cylinders are normally used in luxury sedans. The boxer engine is more rare, but is used when a low centre of gravity is required and is for instance found in sport cars like Porsche and the four-wheel driven Subaru.

### 3.4 Transmissions

The transmission has several functions. In conventional vehicles with internal combustion engines, it must close the speed gap between wheels at standstill and the minimum angular speed of the engine, i.e. it must enable start. The transmission must adapt the available machine power to the operating condition of the vehicle. At maximum acceleration the
machine should deliver maximum power and the transmission should have a gear ratio that fulfills this requirement for any specific vehicle speed. The transmission must also optimize the average working point of the machine according to criteria as fuel consumption, exhaust emissions and traffic noise. Some experts predict that there will be more changes in the field of transmissions over the next 15 years than there have been throughout the past 50 years, due to driving forces like fuel-saving potentials and comfort [33]. The manual transmission (MT) has traditionally been preferred in Europe while the automatic transmission (AT) has so far been preferred in the USA, where 90% of drivers use automatics. The automated shift gearbox (ASG) offers a 15% improvement in fuel economy compared to conventional automatic transmissions. It is constructed like a manual transmission, but with additional actuators, a controller and software, which enables automatic shifting. This system is successfully used in Europe, where the population is familiar with manual transmissions. The torque interruption from shifting is unacceptable in the USA, however, due to the preference of conventional automatics, which are constructed with planetary gears, brakes and clutches. The parallel shift gearbox (PSG) is being developed to improve the comfort while still performing ASG economy. This system uses a dual-clutch system that operates without interruption of the tractive force, and the subjective impression of a pause in gear shifting is eliminated. The PSG can also be adapted as a parallel hybrid transmission system like an integrated starter generator (ISG/ESG), which eliminates the starter motor and generator, with the result that the system functionality is dramatically increased with no additional components. The continuously variable transmission (CVT), that can change the gear ratio arbitrarily, also continuous to evolve with greater torque capability and improved efficiency [33]. Some different transmission components are shown in Figure 3.6.

![Figure 3.6: Different transmission components: (MT) Manual Transmission, (PSG) Parallel Shift Gearbox, (AT) Automatic Transmission and (CVT) Continuously Variable Transmission.](image)

### 3.5 Components for electric propulsion

#### 3.5.1 Batteries

There are three types of batteries, which are usually used in vehicles for electric traction. These are the lead-acid battery, the nickel-metal hydride battery (NiMH) and the lithium-ion battery (Li-ion). The traditional type is the lead-acid battery, which is also found in almost all conventional vehicles of today as a standard battery and is the most cost effective alternative. A typical pure electric vehicle equipped with lead-acid batteries has a driving range of 50-70 km [31]. The NiMH-battery is more costly, due to the use of relatively expensive materials and the complex manufacturing process, but is partially offset by a much longer service life than that of lead-acid batteries. A battery service life of up to 10 years or 2000 cycles has been demonstrated [31]. The Li-ion-battery allows energy densities of over 100 Wh/kg and power
densities of over 300 W/kg and has become very successful in the electrical-appliance market (for products like laptops, mobile phones and cameras). A disadvantage with the lithium battery is that it requires a relatively complex battery protection system, because the individual battery cells are not proof against overcharging and must also be protected against short circuits [31]. Some properties of the described battery types are given in Table V.

<table>
<thead>
<tr>
<th>Properties</th>
<th>Lead-acid</th>
<th>NiMH</th>
<th>Li-ion</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cell voltage, [V]</td>
<td>2</td>
<td>1.2</td>
<td>3-4</td>
</tr>
<tr>
<td>Energy density, [Wh/kg]</td>
<td>25-30</td>
<td>35-100</td>
<td>60-150</td>
</tr>
<tr>
<td>Power density, [W/kg]</td>
<td>100-200</td>
<td>100-500</td>
<td>300-1500</td>
</tr>
<tr>
<td>Energy efficiency¹</td>
<td>75-85</td>
<td>60-85</td>
<td>85-90</td>
</tr>
<tr>
<td>Operating temperature [°C]</td>
<td>10-55</td>
<td>-10-55</td>
<td>-10-60</td>
</tr>
<tr>
<td>Service life (cycles)</td>
<td>600-900</td>
<td>&gt;1000</td>
<td>&gt;1000</td>
</tr>
<tr>
<td>Maintenance-free</td>
<td>Yes/No²</td>
<td>Yes/No²</td>
<td>Yes</td>
</tr>
</tbody>
</table>

¹ Without heating/cooling
² Depending on design

The energy density of gasoline and diesel is for comparison 12,0 kWh/kg (43.5 MJ/kg) and 11.8 kWh/kg (42.5 MJ/kg), respectively. These fuels have energy densities that are close to one hundred times higher than for the best batteries.

### 3.5.2 Fuel cells

Fuel cells convert chemical energy of a suitable fuel and atmospheric oxygen into electrical energy. The most common fuels are hydrogen (H₂), methanol (CH₃OH) and, to a limited extent, also methane (CH₄) (at very high temperatures) [31]. Conventional fuels cannot be used directly, but must be converted into hydrogen in a chemical gas-reforming reaction. The fuel cells are modular and can be used over a wide power range from a few watts to several megawatts. They are also very efficient and produce low levels of harmful emissions. In the case of hydrogen as fuel, the only exhaust emission is water vapor. Many automobile manufacturers therefore see the fuel cell drive as a serious alternative to the internal combustion engine and the major vehicle manufacturers are therefore working intensively with development of fuel cells suitable for automobiles. The fuel cells have mainly been used in spacecrafts and submarines up to date, as means of generating electrical energy. Fuel cells have an operating temperature, depending on the design, in the range of room-temperature up to approximately 1000 °C. There are some drawbacks with using fuel cells that need to be solved, in order to be explored in the automotive applications. Hydrogen as fuel has only satisfactory energy density if it is stored in pressurized or liquefied form, in gaseous form in cylinders at pressures of up to 300 bars or in liquid form in cryotanks at −253 °C. Precious metals like Platinum and Palladium is today needed as catalysts in low temperature fuel cells, while Nickel can be used in high temperature fuel cells for industrial applications. A new infrastructure, different from the existing gasoline and diesel infrastructure, is needed in a future hydrogen society. Hydrogen is like gasoline an explosive fuel, which requires to be handled with safety. The response time of fuel cells are poor, which means that they need to be supported by batteries or capacitors in transport applications. When taking into account the entire energy chain from the primary energy source to the powered wheels of the vehicle, and given the technology available at present, the overall efficiency of modern diesel and fuel-cell-powered vehicles of equal power-to-weight ratios is comparable [31]. However, in
comparison to the oil reserves, which could be empty within the time frame of a few
generations, the amount of hydrogen is almost unlimited (not as a primary source, though).
There are approximately 100 (one hundred) fuel cell powered vehicles in the world today.

3.5.3 Inverters

Inverters are devices that convert a direct current (DC) from a battery or a similar source into
an alternating current (AC), which can be used for driving electrical AC machines. An
inverter consists of a power module, DC-link capacitors, sensors, a filter and a control system.
The power module for hybrid electric vehicles is usually semiconductor components of the
types MOSFET (Metal-Oxide Semiconductor Field Effect Transistor) in low voltage
applications (<300 V) and IGBT (Insulated Gate Bipolar Transistor) in applications at higher
voltage (200-2000 V).

3.5.4 Electrical machines

3.5.4.1 DC machines with commutators

Direct Current (DC) motors normally have windings in the rotor and permanent magnets in
the stator. The permanent magnets are sometimes replaced by windings in the stator. The
rotor windings of a DC motor are wound from the outside, which is an advantage for a simple
and cost effective winding automation. In order to transfer current to the rotor windings, the
DC motor is equipped with brushes and a commutator. The brushes are the static part and the
commutator is the rotating part of this electric contact. A dc rotor with brushes is shown in
Figure 3.7 together with a cross-sectional drawing of a typical DC servomotor [34].

![Figure 3.7: (a) A DC-rotor. (b) A commutator (rotational contact). (c) A principle cross-
sectional sketch of a DC-machine with permanent magnet field excitation in the stator.]

Due to the fact that a dc servomotor has the armature winding (of copper) in the rotor, the
rotor inertia is relatively high. A higher inertia results in a slower acceleration of the rotor for
a given value of torque. A DC motor requires a drive with at least four power transistors in an
H-bridge configuration, which are less than in traditional three-phase machines that require
six transistors in variable speed drives. The DC motor also has simpler control. DC machines
are used for traction in electric vehicles in for instance the Renault Clio, and also in several
models of fork lifts. Every new conventional car has also 50 to 100 small DC motors for
various apparatus like door-lock openers, screen wipers, CD-players, fans, pumps, electric
mirrors, sun-roof and many more.
3.5.4.2 Induction machines

The induction machine is the workhorse of the industry applications. The main advantage of the induction machine, or asynchronous machine which it is also called, is that it can run without any inverter or control electronics directly from the line supply. The rotor winding is often manufactured with short-circuited aluminium bars, called a squirrel-cage, and the stator windings with conventional copper coils. The induction machine is robust and very cost effective for its size. One disadvantage is that it has rotor losses due to the currents in the rotor bars. Since the rotor is relatively difficult to cool, the rotor gets hot and the bearings are also warmed up. These effects affect the rated power or the lifetime of the machine. A squirrel-cage induction machine is shown in Figure 3.8 [35]-[36]. The induction machine can be used in electric vehicles and is then controlled by a frequency-inverter, which enables variable speed operation of the drive system. For instance the General Motors EV-1 is equipped with an induction machine driving the wheels.

![Figure 3.8: (a) X-ray drawing of a squirrel-cage industrial induction machine (ABB). (b) Cross-sectional view of a quarter of a four-pole induction machine.](image)

3.5.4.3 Reluctance machines

3.5.4.3.1 Synchronous reluctance machines

The synchronous reluctance machine (SynRM) can have similar stator design as the induction machine. The difference is mainly in the rotor design. The SynRM rotor can be manufactured with punched iron laminations, making it a very cost effective motor alternative. It is even slightly more cost effective than the induction machine of the same size, since it does not have any rotor windings or aluminium bars. Increasing the d-inductance and the d-inductance to q-inductance ratio maximizes the torque production of the synchronous reluctance machine. This can be achieved by designing the rotor with a certain pattern of iron and air along the flux path as illustrated in Figure 3.9 [37]. The SynRM has poor damping and bad starting characteristics, which makes it a poor alternative as a line-connected device. However, in drive systems it is a very interesting candidate. The SynRM should preferably be designed with a low pole number, otherwise the leakage inductance would be high and the torque production would be poor. The usual pole number is four or six. The synchronous reluctance machine is reportedly capable to perform an average of 20% more rated torque compared to an induction machine with similar stator [38]. The main reason for this is that it lacks rotor losses. There are no winding losses in the rotor and almost no core losses either, because the rotor experiences only a magnetic DC field since it rotates in synchronism with the rotating magnetic field of the stator.
3.5.4.3.2 Switched reluctance machines

The switched reluctance machine (SRM) is a doubly-salient electrical machine, i.e. that the poles are salient in both stator and rotor. It is also referred to as a singly-excited machine, since the excitation is provided solely by the phase windings. SRMs exist in many variants as illustrated in Figure 3.10 [39].

![Figure 3.10: Six types of switched reluctance machines. (a) 6/4-pole 3-phase, (b) 8/6-pole 4-phase, (c) short flux-path 10/8-pole 5-phase machine, (d) full-pitch 6/4-pole 3-phase, (e) stepped-gap 4/2-pole 2-phase and (f) bifurcated teeth and 12/10-pole 3-phase machine.](image)

The torque generating principle in a SRM is based on exciting the phases with respect to the actual rotor position. The rotor tends to minimize the reluctance in the magnetic circuit comprising of the rotor and the stator. By switching the active phases in a certain pattern, the rotor can get a controlled rotation. The phases are excited with rectangular or part-time direct current. The machine is robust and it is also cost effective to manufacture, since the active part only consists of two materials, iron and copper. Due to the current commutations and the machine’s “digital” character it produces high torque ripple. Adding more phases can reduce the torque ripple, but this also increases the complexity of the system. The SRM need special control algorithms. The base algorithms used for conventional asynchronous- and synchronous AC machines are not applicable. The switched reluctance machine has also been developed for electric vehicles. Volkswagen in cooperation with the University of Aachen in Germany has manufactured prototype machines [40]. The SRM can also be equipped with permanent magnets in the stator or rotor to increase its torque capability [41]-[42].

3.5.4.4 Permanent magnet machines

Permanent magnet machines are often categorized after the orientation of the magnetic flux they are carrying, e.g. radial flux, axial flux and transversal flux machines. Short descriptions of these three topologies are given below.
3.5.4.4.1 Radial flux machines

The radial flux machine is the most common variant. It is very flexible in terms of scaling in production. The radial flux machine can be increased in torque and power capability simply by stacking more silicon iron laminations in the axial direction. The rotor is normally placed on the inside of the stator core, but an opposite placement is also possible. In the former case the permanent magnets can be mounted in several different ways. Three main categories are normally defining them, surface-mounted, inset and interior permanent magnet rotor designs. The radial flux permanent magnet machines will be further described in Chapter 4.

3.5.4.4.2 Axial flux machines

The axial flux machine has recently got a lot of attention in university- and industry research programs [43]-[55]. The main characteristic of the axial flux machine is the high torque density. It is therefore often used in applications with low space- and weight requirements. Most axial flux machines are permanent magnet machines, but also induction machines and reluctance machines can use this configuration. A number of topologies are possible with the axial flux concept. Among them are [43]-[44]:

- Single air gap with one stator and one rotor,
- dual air gaps with one stator and two rotors,
- dual air gaps with two stators and one rotor and
- multiple air gaps with theoretically unlimited stacked stators and rotors.

By stacking stators and rotors, the concept is easily scaled using building blocks or modules. Doubling the rated output power of the axial flux machine can be achieved by simply stacking two machine modules together, leading to manufacturing flexibilities. Axial flux machines typically have very short axial length as illustrated by the in-wheel-motor in Figure 3.11 [54].

\[ T = k_{w1} \pi K_{11} B_1 R_o^3 k_r (1 - k_r^2), \]  

(3.8)

where \( K_{11} \) is the fundamental electric loading at the inner radius, \( B_1 \) is the air gap flux density, \( R_o \) is the outer radius of the permanent magnets, \( k_r \) is the ratio between inner and outer radii of the permanent magnets and \( k_{w1} \) is the fundamental winding factor. The constant \( k_r \) is an important design parameter of axial flux machines. Given the air gap flux density and the electric loading at the inner radius, the torque generated by the machine can be maximized by choosing \( k_r \) equal to the inverse of the square-root of three \((\approx 0.57)\). This is understood by taking the derivative of (3.8) and setting it equal to zero. Due to practical reasons, the value of \( k_r \) is sometimes chosen in the interval 0.6-0.7. Because of the high torque density characteristic of the axial flux machine, it is for example used in low speed applications like elevator systems without gearbox and in traction applications [53]-[54].
3.5.4.4.3 Transversal flux machines

Transversal flux machines (TFM) have a toroidal armature coil, carrying current parallel to the direction of rotation. The stator core is salient and the rotor is equipped with permanent magnets, often in a doubly salient arrangement. The stator core often carries flux in three directions, which means that iron powder is a preferred core material. The transverse flux machine is recognized of having an extremely high torque density [56]-[57]. In order to understand the increased torque capability of TFMs, the torque limitations of conventional machines are discussed below [58]. The torque delivered by any winding is given by the rate of change of co-energy of that winding with respect to position and may be expressed by:

$$T = l \frac{\partial \psi}{\partial \theta} = N l \frac{\partial \phi}{\partial \theta}$$

(3.9)

The total MMF of the winding is related to the electric loading and thermally limited. The maximum flux linking each turn of the winding is generally the flux per pole. This is inversely proportional to the number of poles and is limited by the magnetic loading. The spatial frequency of the magnetic flux is directly proportional to the number of poles. This means that the torque of conventional machines is mainly independent of the pole number. Transversal flux machines overcome these limitations by ensuring that the winding links the flux of every pole in the machine. Simple analysis suggests that the torque capability of the TFM will rise linearly with pole number. In practice this does not happen, due to magnetic leakage, but the optimum pole number is generally very high [58]. Figure 3.12 illustrates an example of a one-phase arrangement of a TFM. The magnets are magnetized circumferentially in alternating directions. In between the magnets are iron poles whose axial surfaces form the active faces of the rotor. Flux passes circumferentially through the magnets and then axially into the air gap. It passes axially through teeth on the stator before turning radially into the stator core back. It then passes axially through the core back to return in the same way through the other axial half of the machine thereby enclosing the stator phase coil [58]. The machine described above uses separate magnetic circuits for each phase. Consequently, each phase of the machine can operate completely independently of all others. It is possible to stack the machine with more phases to produce a multi-phase machine. In fact, if a transversal flux machine consists of only one phase, then its torque output becomes zero for every zero-crossing of the phase current, i.e. twice per electric period. Using the analogy of a rectifier, the torque of a one-phased TFM has the same waveform as the output voltage from a rectifier fed with a sinusoidal input voltage. However, the transversal flux machine is usually stacked with more phases in a manner so that the other phases produce torque when the first does not and vice versa. The ripple torque and the cogging torque are generally very high in this machine type. Because the pole number is often very high they are preferably used in low speed applications. The leakage inductance is normally greater than the main inductance, which results in a poor power factor of typically 0.4.

![Figure 3.12](image)

**Figure 3.12** (a) Transversal flux machine, showing the arrangement of a single phase. (b) Ditto in longitudinal cross-sectional cut.
3.6 Hybrid electric vehicles

3.6.1 The market of hybrid electric vehicles

The first commercial Hybrid Electric Vehicle (HEV), the Prius, was introduced in Japan by Toyota in December 1997. The hybrid vehicle business is currently a $2.5 billion market, and is expected to triple by 2006 [59]. The Japanese automobile manufacturers Toyota and Honda are the pioneers within the field. Toyota sells about seven models, including trucks and buses, and additionally plans to release two SUV variants this year; the Lexus RX 400h and the Highlander. The accumulated sales of Toyota hybrids from 1997 reached 202,735 units at the end of February 2004 [59]. Honda sells hybrid versions of its Civic compact car and Insight two-seater, and will also introduce the Accord midsized car in the United States later this year. Honda has sold 58,835 hybrid vehicles worldwide between November 1999 and 19 March 2004. Other car manufacturers, including General Motors, Ford and Nissan, are also planning hybrid releases [59]. Hybrid cars generally cost about $3,000 more than a conventional gasoline car, due to its additional components [59]. This difference can be reduced with increased production volumes, but there is also a need for evolutionary concepts in order to cut the extra cost down to zero or beyond.

3.6.2 Hybrid electric system topologies

The drivetrains of hybrid electric vehicles are normally divided into three topologies, namely the series hybrid, the parallel hybrid and the power-split hybrid. They are briefly described below.

3.6.2.1 Series hybrid

The internal combustion engine in a series hybrid is not mechanically connected to the wheels. The speed of the engine can therefore be chosen independently from the speed of the wheels. The engine is instead mechanically connected to a generator. The generator produces electrical energy controlled by an inverter to feed an electric motor controlled by yet another inverter and to charge the electric energy storage device, which can be a battery or supercapacitors. The electric motor is mechanically connected to the wheels, usually with a single- or dual-stage gearbox in between, but a direct drive motor solution is also theoretically possible. Motors integrated in the wheels can for instance directly drive the wheels. A principle sketch of the series hybrid drive system is illustrated in Figure 3.13. The torque of the engine is preferably chosen at a high average level, where the efficiency is high and the speed is chosen so that the engine delivers sufficiently high average power. The battery or supercapacitors function as an energy source for the transient load conditions. This means that the ICE can be operated at very efficient working points, especially at low loads, so that the fuel consumption and the emissions are minimized. The series hybrid vehicle normally has a smooth and easily controlled performance and an excellent comfort for the passengers, due to the pure electric traction force. It can also be run a considerable distance in zero emission mode, depending on the installed battery capacity. One disadvantage with the series hybrid is that the power ratings of all the components need to be of the same order of magnitude, since they are connected in series, which results in a relatively heavy and expensive system. Another disadvantage is that the power is converted several times from the ICE to the wheels, leading to additional losses and higher fuel consumption at highway driving, when high power is required [60]-[61].
3.6.2.2 Parallel hybrid

Quite oppositely compared to the series hybrid the internal combustion engine in a parallel hybrid is mechanically connected to the wheels. An electric machine is also connected on the same drive shaft. The electric machine controlled by an inverter can either brake the torque from the engine or add torque to the shaft and handles the worst torque transients in the propulsion system. The electric energy is stored in batteries or supercapacitors. A principle sketch of the parallel hybrid system is illustrated in Figure 3.14. The main advantage of the parallel hybrid is that the electrical components are few and can have lower power ratings than the engine, because the power is not converted as in the series system, but added in parallel. Furthermore, the engine can operate on its own at highway driving, when high power is required, for better system efficiency. One disadvantage is, since the engine is mechanically connected to the wheels, the angular speed of the engine cannot be chosen freely. However, the torque of the engine can be chosen independently by support from the electrical machine. This sometimes means higher fuel consumption and definitely higher emissions compared to the series hybrid at city traffic driving. In some parallel systems it is possible to disconnect the engine by using a clutch. The vehicle can thus be run entirely electric, but this requires a high-energy battery to acquire a reasonable driving range, which adds to the cost and weight of the vehicle. The gearbox is often conventional with several steps, due to the mechanical connection with the engine [61]-[62].

3.6.2.3 Power-split hybrid

A power-split hybrid is a combination of the series and parallel hybrids, with the intention to maximize the benefits of both systems. It has two electrical machines and in normal operation one is used as a generator which is feeding the other that runs as a motor with the required power. At first impression it can look inefficient to introduce this loop of electric power, but
this is deliberately done in order to improve the efficiency of the internal combustion engine by supporting it and allowing it to work at its best working points concerning fuel consumption. Secondly, the installed battery capacity can be chosen to be low in order to reduce cost and weight, but relatively high electrical machine powers that allow flexible operation is still possible, because of the electric power loop. A principle sketch of the power-split hybrid system is illustrated in Figure 3.15.

Figure 3.15: Principle sketch illustrating the power-split hybrid system.

3.6.3 Commercially available hybrid electric vehicles

3.6.3.1 Toyota Prius

The Prius is a so-called power-split hybrid, which combines some of the advantages from the series and the parallel hybrid systems, respectively. It means that the vehicle is able to run as a zero emission vehicle, but with an installed electric power that is less than the rated power of the internal combustion engine. Toyota Prius is currently also available in Europe and in the United States. The second version sold abroad has a modified drivetrain with higher output power than the original version and is illustrated in Figure 3.16. The Prius has a maximum speed of 170 km/h and an acceleration performance of 10,9 seconds (0-100 km/h) [63].

Figure 3.16: (a) The Toyota Prius. (b) The drive system consisting of a NiMH-battery (1), an internal combustion engine (2), inverters (3) and electrical machines (4).

A vital component in the drivetrain of the Prius is a planetary gearbox. The Otto-engine is connected to the planetary carrier, a generator is connected to the sun gear and motor is connected to the ring gear. The planetary gearbox is the power-splitting device in the vehicle.
The drivetrain and a schematic view of the planetary gearbox are illustrated in Figure 3.17. The electric machines are called generator and motor due to their normal mode of operation, but they do also switch behaviour at certain working conditions. The engine delivers a maximum power of 57 kW (78 hp) and a maximum torque of 115 Nm. The engine has a displacement of 1.5 liters and uses a high expansion ratio cycle called Atkinson after the British inventor. The thermal efficiency of this engine is high, but it has virtually no practical application unless used with a supercharger, because it does not easily provide high output torque. In the case of the Toyota Prius, the engine is supported by two electrical machines that gives the vehicle a performance level that exceeds that of a conventional vehicle with a 2.0 liters gasoline engine [63]. The electrical machines have a maximum power of 50 kW (68 hp) and a maximum torque of 400 Nm. They are synchronous permanent magnet machines with distributed windings. At least the machine called motor has buried permanent magnets in a V-shape for flux concentration. A 25 kW NiMH (Nickel-Metal hydride) battery from Panasonic stores the electrical energy in the vehicle. It has been speculations about Toyota’s choice of battery chemistry. The Li-ion battery has a better potential concerning power density and cost. One possible explanation is that the real expertise on this field may lie outside of Japan, in the pioneering companies SAFT and LG [64]. The motivation of the choice of NiMH instead of Li-ion batteries is normally a safety concern, but the existing production facilities for HEV batteries are also dominated by the NiMH technology. A Japanese battery specialist, Hideo Takeshito, estimates that a switch of chemistry could happen in 2006-2007 and that 25% of the hybrids could use Li-ion batteries by 2010 if the cost drops down to $0.30/Wh [64]. The 2004-model of Toyota Prius has a retail price in the United States of $20 810 [65]. The drive system in Toyota Prius was awarded the trophy “International Engine of the Year 2004” in competition with all the commercial engines of the world in Stuttgart on May 26, 2004 [66]. The production plan for the vehicle has recently increased from 76 000 to 130 000 a year, and the technology will also be licensed out to Ford for use in its own vehicles [66].

![Figure 3.17 The drivetrain of the Toyota Prius shown in (a) a X-ray view and (b) a schematic view.](image)

### 3.6.3.2 Honda Insight

The second hybrid vehicle to be commercially available was the Honda Insight, which first was introduced in Japan and the USA in November of 1999. It is equipped with an electric driveline called a power assist and is a parallel HEV. The Honda Insight is shown in Figure 3.18 and its drivetrain is shown in Figure 3.19. The vehicle is powered by a 3-cylinders 1,0
liters lean-burn engine that delivers a maximum of 49 kW (67 hp). The electric machine is permanent magnet excited and delivers a maximum of 10 kW (14 hp). It is a brushless permanent magnet machine with 12 poles and 18 slots. The permanent magnets are surface-mounted with a negligible saliency, according to Figure 3.19. The stator core is segmented and the coils are pre-fabricated that results in a high copper packing factor. The windings are concentrated with double layers per slot. The total length of the machine is 60 mm, 40% less than would be possible with a conventional motor (with full-pitch windings) [62]. The electrical energy for the electric dive system is stored in a NiMH battery from Panasonic. In 2005, Panasonic and (to a lesser extent) Sanyo will be the only players in the market of NiMH batteries for hybrids [64]. This market is estimated to be worth $300 million [64].

Figure 3.18: The Honda Insight hybrid electric vehicle.

The skeleton body of the Honda Insight made of aluminium is approximately 40% lighter than a steel body [4]. It also has a very low air drag coefficient of 0.25. Together with the automatic idling stop function, the system works like an ISG (Integrated Starter Generator). The Insight is the most fuel efficient car in the world, including diesel powered cars, but of course with the exception of zero emission vehicles. The 2004-model of the Honda Insight has a retail price in the United States of $19,180 [67]. Honda also offers a hybrid variant of its Civic model that uses the same electric system as the Insight.

Figure 3.19: (a) The drivetrain of the Honda Insight shown in X-ray view. (b) The electric machine that consists of 12 poles and 18 slots.
4 Permanent magnet machines

4.1 Introduction

This chapter will describe in general terms the different topics that are important for the characterization of permanent magnet machines. Among the presented topics are magnetic materials, power losses, field-weakening, windings and the lifetime of electrical machines. In order to give an intuitive understanding of the machine type, several various topologies have been designed and analysed, as described below.

4.1.1 Overview of analysed permanent magnet machines

An overview of the in the thesis analysed permanent magnet machines with laminated cores is shown in Figure 4.1 and Table VI. Their design variations incorporate among others inner and outer rotors, distributed and concentrated windings, and surface-mounted, inset and interior permanent magnets. They are used to find characteristic performances for the different machine topologies in terms of torque capability, ripple torque, inductance, losses, magnetic noise and field-weakening capability. The main differences and the location of the analysis of the various machines are summarized in Table VI.

Table VI: Overview of the analysed permanent magnet machines in the thesis.

<table>
<thead>
<tr>
<th>Machine number</th>
<th>Permanent magnet rotor Mounting</th>
<th>Position</th>
<th>Winding configuration ¹</th>
<th>Design data ²</th>
<th>Location of analysis</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>Surface</td>
<td>Outer</td>
<td>Distributed</td>
<td>14 42 1</td>
<td>5 -</td>
</tr>
<tr>
<td>II</td>
<td>Surface</td>
<td>Outer</td>
<td>Concentrated</td>
<td>14 21 0,5</td>
<td>5 -</td>
</tr>
<tr>
<td>III</td>
<td>Surface</td>
<td>Outer</td>
<td>Concentrated (Modular)</td>
<td>14 15 0,36</td>
<td>3, 5, 6 5</td>
</tr>
<tr>
<td>IV</td>
<td>Surface</td>
<td>Inner</td>
<td>Concentrated</td>
<td>8 12 0,5</td>
<td>- 4</td>
</tr>
<tr>
<td>V</td>
<td>Surface</td>
<td>Inner</td>
<td>Concentrated</td>
<td>8 12 0,5</td>
<td>- 4</td>
</tr>
<tr>
<td>VI</td>
<td>Surface</td>
<td>Inner</td>
<td>Concentrated (Modular)</td>
<td>10 12 0,4</td>
<td>- 4</td>
</tr>
<tr>
<td>VII</td>
<td>Surface</td>
<td>Inner</td>
<td>Concentrated (Modular)</td>
<td>10 12 0,4</td>
<td>- 4</td>
</tr>
<tr>
<td>VIII</td>
<td>Inset</td>
<td>Inner</td>
<td>Distributed</td>
<td>8 48 2</td>
<td>8 -</td>
</tr>
<tr>
<td>IX</td>
<td>Interior</td>
<td>Inner</td>
<td>Distributed</td>
<td>8 48 2</td>
<td>8 -</td>
</tr>
<tr>
<td>X³</td>
<td>Surface</td>
<td>Outer</td>
<td>Distributed (Gramme)</td>
<td>16 48 1</td>
<td>3, 9 5</td>
</tr>
</tbody>
</table>

¹Layers. 1: single layer per slot, 2: double layers per slot
²Design data. \( p \): number of poles, \( Q \): number of slots, \( q \): number of slots per pole per phase
³Machine with Soft Magnetic Composites core (X), (Machines I-IX have silicon iron laminated cores)
Figure 4.1: Overview of the analysed permanent magnet machines with laminated cores, illustrated with their open-circuit flux distributions.

4.2 Magnetic materials

4.2.1 Introduction

The induction $B$ theoretically consists of two components in matter, where the first comes from the applied field and the second from the material, as expressed by

$$B = \mu_0 \cdot H + J$$

(4.1)

where $\mu_0$ is the permeability in free space, $H$ is the magnetic field strength and $J$ is the magnetic polarization. In physical terms, $J$ is the magnetic dipole moment per unit volume, and is generally a function of the magnetic field strength. For many materials, the second term is normally much larger than the second term in (4.1), and the magnetic polarization is proportional to the magnetic field strength, so that
\[ B = \mu_r \cdot \mu_0 \cdot H \]  

(4.2)

where \( \mu_r \) is the relative permeability of the material. Materials can be divided into three groups, diamagnetic, paramagnetic and ferromagnetic, according to their relative permeability values as classified in Table VII [31]. It is the ferromagnetic materials that are treated in this section. The magnetic polarization of these materials is very high, changes non-linearly with the magnetic field strength, and is also dependent upon hysteresis.

**Table VII: Classification of materials into three groups according to the relative permeability.**

<table>
<thead>
<tr>
<th>Diamagnetic materials (( \mu_r &lt; 1 ))</th>
<th>Paramagnetic materials (( \mu_r &gt; 1 ))</th>
<th>Ferromagnetic mat. (( \mu_r \gg 1 ))</th>
</tr>
</thead>
</table>
| \( \mu_r \) is independent of magnetic field strength and smaller than 1.  
(1-10^{-5} < \mu_r < 1-10^{-11}) 
(e.g. Ag, Au, Cu, Zn, H2O) | \( \mu_r \) is independent of magnetic field strength and greater than 1.  
(1+10^{-8} < \mu_r < 1+4 \cdot 10^{+4}) 
(e.g. Al, Pt, Ti, O2) | \( \mu_r \) is a function of magnetic field strength and much greater than 1.  
(10^{2} < \mu_r < 5 \cdot 10^{5}) 
(e.g. Fe, Co, Ni) |

In electrical engineering \( \mu_r \) is chosen as a function of \( H \), according to (4.2). The hysteresis loop, which illustrates the relationship between \( B \) and \( H \) as well as \( J \) and \( H \), is shown in Figure 4.2. The letters in the figure are defined below:

- \( J_s \): Saturation polarization,
- \( H_s \): Magnetic field strength where saturation polarization occurs,
- \( B_r \): Remanence, which is the intersection of the hysteresis loop and the ordinate,
- \( H_{cB} \): Coercivity of \( B \), which is the intersection of the \( B(H) \) loop with the abscissa,
- \( H_{cJ} \): Coercivity of \( J \), which is the intersection of the \( J(H) \) loop with the abscissa.

Let the material exemplified by Figure 4.2 be in an unmagnetized state. When a magnetic field \( H \) is applied, the magnetization of the material follows the virgin curves. When the saturation polarization is reached, which is material-dependent, the polarization can no longer be increased and all magnetic dipoles are aligned, oriented parallel to the external magnetic field. The flux density however, continues to increase with the field strength. If the magnetic field strength is now reduced, the flux density and the polarization changes in accordance with the \( B(H) \) and the \( J(H) \) loop, respectively. When \( H \) becomes zero, the flux density attains residual flux density or remanence (\( B_r \)). The flux density and the polarization have the same values in this point. The second quadrant of the hysteresis loops describes the demagnetization behaviour of the material. The flux density and polarization drop to zero only upon application of an opposing field whose field strength is \( H_{cB} \) and \( H_{cJ} \), respectively. If the opposing field is further increased, saturation polarization in the opposite direction is reached. The curves are symmetrical around the axes [31],[68].

Ferromagnetic materials are divided into hard and soft magnetic materials. A comparison of their typical magnetic characteristics is shown in Table VIII [31]. Short introductions of hard and soft magnetic materials are given below.

**Table VIII: Range of magnetic characteristics of typical hard and soft magnetic materials.**

<table>
<thead>
<tr>
<th>Material</th>
<th>( H_c ) [A/m]</th>
<th>( J_s ) [T]</th>
<th>( B_r ) [T]</th>
<th>( \mu_r ) [-]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hard magnetic</td>
<td>( 5 \times 10^4 ) ... ( 2 \times 10^6 )</td>
<td>0,45 ... 1,4</td>
<td>0,4 ... 1,4</td>
<td>1,1 ... 5</td>
</tr>
<tr>
<td>Soft magnetic</td>
<td>0,3 ... 400</td>
<td>0,9 ... 2,4</td>
<td>Depend on applic.</td>
<td>&lt; 500 000</td>
</tr>
</tbody>
</table>
4.2.1.1 Hard magnetic materials

Hard magnetic materials are also called permanent magnet materials. They have high coercive field strengths, which means that high demagnetizing fields can occur without loss in the magnetic polarization of the material. The magnetic operating range of a permanent magnet is mainly in the second quadrant of the hysteresis loop, on the so-called demagnetization curve. Some permanent magnets with extremely high coercive field strengths can also operate into the third quadrant without loss, if the temperature is moderate. In practice, the working point of a permanent magnet does not coincide with the remanence point, because a demagnetizing field is always present due to the intrinsic self-demagnetization of the magnet [31].

4.2.1.2 Soft magnetic materials

Soft magnetic materials have, in contradiction to hard magnetic materials, a low coercive field strength, which translates into a narrow hysteresis loop. The flux density reaches high values (large $\mu_r$ values) already for low magnetic field strengths, such that $J \gg \mu_0 H$ and no practical distinction needs to be made between the $B(H)$ and the $J(H)$ curves. Soft magnetic materials are used as conductors of magnetic flux due to their high induction at low field strengths. Their typically narrow hysteresis loop results in low hysteresis losses (the hysteresis losses are proportional to the area of the hysteresis loop). These materials are consequently very suited for alternating magnetizing field applications [31].

4.2.2 Permanent magnets

Some common pole configurations of permanent magnets are illustrated in Figure 4.3 [68].
Permanent magnets are often characterized according to their energy product, which is a measure for the maximum attainable air gap energy and is the product of the flux density and the magnetic field strength. The point on the demagnetization curve at which the maximum energy product \((B\cdot H)_{\text{max}}\) is located is illustrated in Figure 4.4.a, and is in the point \((H_a,B_a)\). The magnetic field from the poles of a permanent magnet has a demagnetizing effect as it is in the opposing direction to the flux density. The operational state of a permanent magnet is therefore always in the range of the demagnetization curve. The working point on the demagnetization curve is denoted \(P\) in Figure 4.4.b, and corresponds to a pair of values that refers to the field strength and the flux density, e.g. \((H_b,B_b)\). The position of \(P\) depends on the geometry of the magnetic circuit and on externally applied opposing magnetic fields. \(P\) is located on the intersection of the working or sharing lines with the \(B(H)\) curve [68].

In dynamic magnetic systems where the working points change (e.g. electrical machines), the shearing should be selected so that the working point always remains on the straight region of the demagnetization curve. Otherwise, stability to interference magnetic fields and temperature influences cannot be ensured. Consider point \(P_1\) in Figure 4.4.b, which is on the straight region. An even greater opposing field is applied, so that the new working point is located at \(P_2\). This point is below the characteristic “knee” in the \(B(H)\) curve of the permanent magnet material, where the losses are irreversible. If the applied field is now reduced, the working point shifts along an inner return curve to the point \(P_3\). The slope of this return curve is referred to as \textit{permanent permeability}. The permanent magnet has lost in strength, which is characterized by the lower value of remanence \(B_p\) [68].

The demagnetization curve of permanent magnets are temperature dependent. This dependence is characterized by the temperature coefficients of the remanent flux density and the coercivity, as given by
\[
TC(B_r) = \frac{100}{B_r} \partial B_r/\partial T,  \tag{4.3}
\]

and

\[
TC(H_{cj}) = \frac{100}{H_{cj}} \partial H_{cj}/\partial T. \tag{4.4}
\]

The temperature dependence of permanent magnets is further explained in the section about the lifetime of electrical machines in the end of this chapter. AlNiCo (Aluminium-Nickel-Cobalt), ferrite (mainly iron), SmCo (Samarium-Cobalt) and NdFeB (Neodymium-Iron-Boron) are currently the most important types of permanent magnets in terms of technical applications. Some typical magnetic data of these different permanent magnet materials are shown in Table IX [31],[68].

<table>
<thead>
<tr>
<th>Material</th>
<th>(\rho) [kg/m(^3)]</th>
<th>((B\times H)_{\text{max}})</th>
<th>(B_r)</th>
<th>(H_{cj})</th>
<th>(H_{cl})</th>
<th>(\mu_r)</th>
<th>(TC(B_r)) [%/K]</th>
<th>(TC(H_{cj})) [%/K]</th>
</tr>
</thead>
<tbody>
<tr>
<td>AlNiCo</td>
<td>5500 ... 7200</td>
<td>7 ... 60</td>
<td>0,55 ... 1,25</td>
<td>44 ... 136</td>
<td>47 ... 144</td>
<td>1,5 ... 5,0</td>
<td>-0,02</td>
<td>-0,07 ... +0,03</td>
</tr>
<tr>
<td>Ferrite</td>
<td>3500 ... 4900</td>
<td>3,2 ... 32</td>
<td>0,13 ... 0,41</td>
<td>85 ... 260</td>
<td>165 ... 350</td>
<td>1,1 ... 1,2</td>
<td>-0,2</td>
<td>0,2 ... 0,5</td>
</tr>
<tr>
<td>SmCo</td>
<td>8400</td>
<td>140 ... 240</td>
<td>0,85 ... 1,12</td>
<td>580 ... 820</td>
<td>640 ... 2400</td>
<td>1,05 ... 1,09</td>
<td>-0,04 ... -0,03</td>
<td>-0,25 ... -0,15</td>
</tr>
<tr>
<td>NdFeB</td>
<td>7500 ... 7800</td>
<td>200 ... 415</td>
<td>1,03 ... 1,47</td>
<td>765 ... 1115</td>
<td>795 ... 2865</td>
<td>1,03 ... 1,08</td>
<td>-0,13 ... -0,08</td>
<td>-0,80 ... -0,46</td>
</tr>
</tbody>
</table>

### 4.2.3 Silicon iron

Silicon iron is a soft magnetic material that is predominately used in electromagnetic power apparatus like transformers and electrical machines. The material has a high induction at low field strengths and is therefore an excellent conductor of magnetic flux. Another advantage is the typical low coercive field strength that results in a narrow hysteresis loop, which makes it well-suited for alternating magnetic field applications. The characteristics of soft magnetic materials depend to a large extent on the pretreatment. The effect of machining increases the coercive field strength and the hysteresis loop becomes broader. The coercive field strength can be reduced back towards its initial value through annealing of the material at high temperatures [31]. Silicon is added (~3%) to the iron in order to increase the resistivity and decrease the losses in the material.

The manufacturing process of silicon iron sheets includes cold rolling, annealing, coating and stress relief annealing. The coating is applied in order to improve the surface insulation resistance and the thickness is normally ranging from 1 to 3 micrometers. The coating shall withstand normal punching operations, provides some corrosion resistance and has only a marginal effect on the stacking factor [69]. Silicon iron sheets are produced in a grain-oriented as well as a non-oriented variant. The former has much better magnetic performance in the direction of the grain-orientation than in the perpendicular direction, and is often used in transformers. The latter has almost the same performance in all directions, and is often used in electrical machines. The polarization versus the magnetic field strength and core losses versus frequency for typical grain- and non-oriented silicon iron sheets are shown in Figure 4.5 and Figure 4.6, respectively.
4.2.4 Soft Magnetic Composites

Soft magnetic composites (SMC) or iron powder has recently been introduced in magnetic applications like electrical machines and has consequently a very small market share in electromagnetic apparatus compared with silicon iron sheets.
The manufacturing process of SMC includes crushing, separation, charging, reduction, discharging, mixing, coating and annealing [70]. Iron powder has lower permeability and higher hysteresis losses than silicon steel sheets, but the powder has other unique advantages. The best advantage is the 3D-isotropically magnetic properties. Iron powder can also be pressed into complex parts. The BH-curve and the loss characteristic of the iron powder SOMALOY™500 with 0.5% of the lubricant Kenolube from the manufacturer Höganäs AB are shown in Figure 4.7 and Figure 4.8, respectively.

**Figure 4.7:** Magnetic flux density versus magnetic field strength for the iron powder core material. The BH-curve for a silicon iron laminated core material (Cogent M235-35A) is shown for comparison (broken line).

**Figure 4.8:** Core loss versus frequency for the iron powder material at the sinusoidal peak flux densities of 0.5 T, 1.0 T and 1.5 T, respectively (from bottom to top).
4.2.5 Thermal material properties

Some typical thermal material properties of magnetic materials are shown in Table X. Data on some other materials frequently used as electrical machine parts are also shown for comparison. The materials are listed in accordance to their respective thermal conductivity.

Table X: Thermal material properties of electrical machine parts.

(\(\rho\): mass density, \(c\): thermal capacity, \(k\): thermal conductivity)

<table>
<thead>
<tr>
<th>Machine part</th>
<th>Material</th>
<th>(\rho), [kg/m(^3)]</th>
<th>(c), [J/(kg K)]</th>
<th>(k), [W/(m K)]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductors</td>
<td>Copper</td>
<td>8900</td>
<td>390</td>
<td>401</td>
</tr>
<tr>
<td>Cooling cylinder</td>
<td>Aluminium</td>
<td>2702</td>
<td>896</td>
<td>255</td>
</tr>
<tr>
<td>Core</td>
<td>Iron/steel</td>
<td>7600</td>
<td>745</td>
<td>40</td>
</tr>
<tr>
<td>Core</td>
<td>SMC</td>
<td>7400</td>
<td>450</td>
<td>20</td>
</tr>
<tr>
<td>Permanent magnet</td>
<td>NdFeB</td>
<td>7800</td>
<td>440</td>
<td>9</td>
</tr>
<tr>
<td>Cooling ducts</td>
<td>Water (20°C)</td>
<td>998</td>
<td>4190</td>
<td>0,58</td>
</tr>
<tr>
<td>Conductor insulation</td>
<td>Varnish (Isonol 51)</td>
<td>1100</td>
<td>1000</td>
<td>0,44</td>
</tr>
<tr>
<td>Slot insulation</td>
<td>Nomex</td>
<td>950</td>
<td>1170</td>
<td>0,143</td>
</tr>
<tr>
<td>Airgap</td>
<td>Air (40°C)</td>
<td>1,09</td>
<td>1014</td>
<td>0,0265</td>
</tr>
</tbody>
</table>

Air has very low thermal conductivity, as seen in Table X. It is therefore advantageous to impregnate the windings in the slots after assembly in the production line. A much better cooling performance and a good thermal conductivity between the core and the winding will be achieved.

4.3 Power losses

4.3.1 Iron losses

The most comprehensively analysed and frequently used iron loss model of today includes the hysteresis, eddy current and excess loss components [71]. For a sinusoidal magnetic field in sheet geometry, the iron losses per mass unit can be expressed as

\[
p_{iron} = k_h B_m^2 f + \frac{\sigma \pi^2 d^3}{6 \rho} B_m^2 f^2 + k_e B_m^{3/2} f^{3/2}
\]

(4.5)

where \(k_h\) is the hysteresis material constant, \(B_m\) is the maximum flux density, \(f\) is the electrical frequency, \(\sigma\) is the conductivity of the iron material, \(\rho\) is the mass density of the iron material, \(d\) is the lamination thickness or gauge and \(k_e\) is the excess material constant. If the magnetic field is not sinusoidal, Equation (4.1) must be refined. Several authors have made modified models [71]-[75]. The analytical models in (4.1) and in the references assume that skin effect in the iron sheet is negligible. This is true for ordinary frequencies and gauges, but this assumption is incorrect for higher frequencies [76]. The skin effect causes the flux density to change across the thickness of the sheet. The iron losses will therefore have a frequency dependent behaviour, which is not according to the theoretical models. The frequency limit for which (4.1) becomes inaccurate depends on the gauge and the permeability, which varies with the induction level. The frequency limit is extended for decreasing gauge. The models are valid when \(\gamma << 1\), which is defined by [77]
**Chapter 4 Permanent magnet machines**

\[ \gamma = \sqrt{\pi \sigma \mu} f \cdot d = \frac{d}{\delta_p}, \]  

(4.6)

where \( \mu \) is the permeability of the iron material and \( \delta_p \) is the penetration depth. When the penetration depth becomes smaller than the gauge, the eddy current loss model must be refined. The following expression is proposed by Bertotti for the eddy current loss component in (4.1) when \( \gamma \gg 1 \), [77]

\[ p_{\text{eddy}} = \frac{\pi^{3/2} \eta d}{2 \rho} \sqrt{\frac{\sigma}{\mu}} B_m^2 f^{3/2}. \]  

(4.7)

The model is only valid under linear magnetic conditions. The frequency dependence in the two models differs. The eddy current loss component in (4.7) is larger than in (4.5) at first, but at increasing frequencies a breaking point is reached and it becomes vice versa.

A novel mathematical model for the core losses per mass unit that includes the skin effect is given by [78]

\[ P_{\text{iron}} = k_h B_m^{2+p_1+2} f^{2+p_2} + \frac{\sigma \pi^2 d^2}{6 \rho} B_m^2 f^{2-p_4} p_3 + k_e B_m^{(3/2)+p_5} f^{(3/2)+p_6}, \]  

(4.8)

where \( p_1, p_2, p_3, p_4, p_5 \) and \( p_6 \) are constants. \( p_4 \) has the value of 0.9. \( p_1, p_3 \) and \( p_5 \) are coefficients that depend on the material properties and are not dependent on the flux density nor the frequency. \( p_2 \) and \( p_6 \) are formed from the peak value of the magnetic induction as expressed by

\[ p_2 = \kappa_1 B_m^2 + \kappa_2 B_m + \kappa_3, \]  

(4.9)

and

\[ p_6 = \kappa_4 B_m^2 + \kappa_5 B_m + \kappa_6 B_m + \kappa_7, \]  

(4.10)

where \( \kappa_1, \kappa_2, \kappa_3, \kappa_4, \kappa_5, \kappa_6 \) and \( \kappa_7 \) are empirical constants found to be −0.34, 0.47, 0.53, 1.8, −4.3 and −0.3, respectively [78]. The described method is verified by tests on toroidal wound cores in different sizes in the frequency range 50-1000 Hz [78].

As can be understood by the discussion above, the modelling of iron losses is quite complex and therefore the most reliable analytical method of today is based on empirical equations from measurements performed on similar machine types.

**4.3.2 Winding losses**

**4.3.2.1 DC-resistance losses**

The winding losses in a three-phase machine fed with direct current is given by

\[ P_{\text{winding,DC}} = 3 R_{DC} I^2 \]  

(4.11)
where $R_{DC}$ is the DC-resistance for one phase winding and $I$ is the phase current (RMS). The DC-resistance is given by

$$R_{DC} = \rho \frac{L_{winding}}{A_{winding}}, \quad (4.12)$$

where $\rho$ is the resistivity, and $L_{winding}$ and $A_{winding}$ is the length and the cross-sectional area of the phase winding, respectively.

### 4.3.2.2 Eddy current losses

Essentially two phenomena related to eddy currents are present in electrical machine windings [79]. In the first case the eddy currents are caused by the stator current, i.e. skin effect and internal proximity effects, and in the second case the eddy currents are caused by external alternating magnetic fields originating from for instance permanent magnets in motion. The spatial distributions of the eddy currents are quite different in the two cases. They are shown qualitatively in Figure 4.9.

![Figure 4.9: The current distribution in a circular conductor due to a) skin effect and b) an external alternating magnetic field (external proximity effect).](image)

Eddy currents are induced in conductors carrying alternating current. In the cross-section of such a conductor, the centre has the maximum inductance. The inductance decreases in the radial direction meeting an inductance minimum at the periphery of the conductor. The result is that the current density is not uniform over the cross-section. The current density has a minimum in the centre and maximum at the periphery of the conductor. This current distribution increases the effective resistance of the conductor. The effect, called the skin effect, must be considered when large conductor cross-sections are used and/or high frequencies are applied. The DC resistance of the phase winding must be multiplied with a resistance factor, which takes the eddy currents into account. The total eddy currents due to the skin effect and the alternating magnetic field can be superposed. The two described phenomena are given in detail below.

#### 4.3.2.2.1 Losses due to skin effect and internal proximity effects

The AC-resistance factor of the $p$:th conductor layer in a slot for a conductor with rectangular cross-section, as illustrated in Figure 4.10.a, is given by [80]

$$k_{zp} = \varphi(\xi) + \frac{I_z}{I_p} \theta(\xi), \quad (4.13)$$

where $I_p$ is the total current of the regarded ($p$:th) conductor layer and $I_u$ the total phase-equal current between the $p$:th conductor layer and the slot bottom and
\[ \varphi(\xi) = \xi \frac{\sinh 2\xi + \sin 2\xi}{\cosh 2\xi - \cos 2\xi}, \] (4.14)

\[ \psi(\xi) = 2\xi \frac{\sinh \xi - \sin \xi}{\cosh \xi + \cos \xi}, \] (4.15)

where the dimensionless, so-called “reduced conductor height” is

\[ \xi = \alpha \cdot h, \] (4.16)

with

\[ \alpha = \sqrt{\frac{nb}{a}} \frac{\pi \mu_0 f}{\rho_{Cu}} = \sqrt{\frac{nb}{a} \frac{1}{\delta_p}} \] (4.17)

where \(n\) is the number of conductors in the \(p\)th layer, \(b\) is the conductor width, \(h\) is the conductor height and \(a\) is the width of the slot. The resulting average AC-resistance factor for a conductor layer in a slot with rectangular conductor cross-section is given by [80]

\[ k_r = \varphi(\xi) + \frac{m^2 - 1}{3} \psi(\xi), \] (4.18)

where \(m\) is the number of conductor layers in the slot. Figure 4.10 illustrates the relationships of the geometries for rectangular conductors in a slot and for round conductors in air.

\[ \begin{align*}
  & h \quad b \\
  & a
\end{align*} \] (a)

\[ \begin{align*}
  & p = m \\
  & p = 3 \\
  & p = 2 \\
  & p = 1 \\
  & n = 18
\end{align*} \] (b)

\[ \begin{align*}
  & D \\
  & p = 4 \\
  & 3 \\
  & 2 \\
  & 1
\end{align*} \]

Figure 4.10: Definition of slot and conductor dimensions for (a) rectangular wires and (b) round wires.

The first term in (4.14) describes the skin effect, which is caused by alternating current in the studied conductor. The second term describes the internal proximity effect, which is caused by AC-currents in the proximity and in this particular case in other conductors in the same slot. The external proximity effect, which is caused by external alternating magnetic fields,
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e.g. from AC-currents in neighbouring slots or permanent magnets in motion, is described in the next section. The height of a rectangular conductor or the diameter of a round conductor should always be smaller (preferably significantly smaller) than the penetration depth in order to limit the eddy current losses in the conductor. If so, the value of $\xi$ will be lower than one and (4.18) can be simplified into

$$k_T = 1 + \frac{m^2 - 0.2}{9} \cdot \xi^4 \quad (0 \leq \xi \leq 1). \quad (4.19)$$

For the end-windings of an electrical machine, where the conductors are located in air, the eddy current losses are lower and the resistance factor is given by

$$k_{T,\text{air}} = 1 + \frac{m^2 - 0.8}{36} \cdot \xi^4 \quad (0 \leq \xi \leq 1). \quad (4.20)$$

The winding losses of a three-phase machine with rectangular conductors including the skin and internal proximity effects are then given by

$$P_{\text{winding}} = 3 \cdot \left[ k_T \frac{L_{\text{active}}}{L_c} + k_{T,\text{air}} \left( 1 - \frac{L_{\text{active}}}{L_c} \right) \right] \cdot R_{\text{DC}} I^2, \quad (4.21)$$

where $L_{\text{active}}$ and $L_c$ is the active length of the machine and the conductor length, respectively. In a case where the conductor consists of a bunch of round wires concentrically placed around an inner wire, (4.9) has to be modified into the following expression

$$k_o = 1 + 0.59 \frac{I_u (I_u + I_p)}{I_p^2} \psi(\xi) \quad (4.22)$$

where $I_u$ here is the total current enclosed by the regarded layer and 0.59 is a constant taking account for the fact that the stray losses of a round conductor is only 59% of a rectangular conductor. The approximate equation (4.18) is valid if the wire diameter is small and the wires are located in air. The parameters in (4.12) and (4.13) can be redefined for a round conductor according to

$$h = d \quad (4.23)$$

and

$$\frac{nb}{a} = \frac{nd}{nd} = 0.96 \frac{d}{d'}, \quad (4.24)$$

where $d$ is the wire diameter excluding the insulation, $d'$ is the radial distance between two adjacent layers and $d''$ is the distance between two adjacent wires in the same layer. The number of copper wires in the $p$:th layer, for $p \geq 2$, is given by

$$n = 6(p - 1) \quad (4.25)$$

and the number of wires enclosed by this layer is

$$z = 3(p - 1)(p - 2) + 1 \quad (4.26)$$
The geometry for the concentrically placed wires is shown in Figure 4.10. The AC-resistance factor for the first layer is equal to one and for the $p$:th layer it is given by

$$ k_{op} = 1 + 0.59 \left( \frac{1+3(p-1)(p-2)}{6(p-1)} \right) \frac{1+3p(p-1)}{6(p-1)} \varphi(\xi), $$

or when $\xi \leq 1$ the expression can be simplified into

$$ k_{op} \approx 1 + \frac{7+9p(p-2)}{183} \xi^4. $$

The resulting mean value of the AC-resistance factor for all the wires in a round conductor with $m$ layers is

$$ k_O \approx 1 + \frac{3m(m-1)-2}{122} \xi^4 = 1 + \frac{z-3}{122} \xi^4 $$

where the total number of wires in the conductor is

$$ z_{total} = 3m(m-1)+1. $$

$z_{total}$ is often much greater than 3, hence

$$ k_O \approx 1 + \frac{z}{122} \xi^4, $$

where $\xi$ from (4.14), (4.15), (4.23) and (4.24) for a round conductor is given by

$$ \xi = d \cdot \sqrt{0.96 \frac{d}{\delta} \frac{\pi \mu_0 f}{\rho_{Cu}}} = \frac{d}{\delta_p} \cdot \sqrt{0.96 \frac{d}{\delta}}. $$

The current distribution in a round conductor that consists of 19 wires and carrying an alternating current is illustrated in Figure 4.11.

![Current distribution in a conductor of 19 wires carrying an alternating current.](image)

*Figure 4.11: Current distribution in a conductor of 19 wires carrying an alternating current.*
4.3.2.2 Losses due to external alternating magnetic fields

The current distribution in a round conductor that consists of 19 wires that is exposed to an external alternating magnetic field is shown in Figure 4.12. It is important to notify that eddy current losses are induced in this conductor even if no current is applied to the conductor itself, which for instance is the case in a slotless permanent magnet machine at no-load operation. The induced eddy currents in the conductor create a field that opposes the applied external magnetic field. The distribution is equal in all the 19 wires, if the skin effect is neglected. The wire can be divided into two symmetrical sections with equal current distributions but with opposite signs of the currents in them. The resultant induced current density for the wire is therefore zero. Consequently, the eddy currents due to the alternating magnetic field only increase the losses in the conductor and do not contribute to the applied current.

\[ P_w = \frac{\pi \sigma L \omega^2 B^2 d^4}{64}, \]  

(4.33)

where \( \omega \) is the electrical angular frequency and \( B \) is the flux density (RMS) from the external source. The eddy currents caused by the alternating magnetic field is, as can be seen in (4.33), completely independent of the armature current.

Figure 4.12: Current distribution in a conductor of 19 wires exposed to an external alternating magnetic field. One of the wires is enlarged and shows the direction of the induced eddy current opposing the applied alternating field.
4.3.3 Mechanical losses

4.3.3.1 Windage losses

The windage losses of a rotating cylinder is given by [82]

\[ P_{\text{windage}} = C_f \rho \omega_m \omega^3 \rho R^4 L \]  

(4.34)

where \( \rho \) is the mass density of the gas (normally air with \( \rho_{\text{air}} = 1.092 \text{ kg/m}^3 \) at 40°C), \( \omega_m \) is the mechanical angular speed, \( R \) is the radius and \( L \) is the length of the cylinder, respectively. The friction coefficient \( C_f \) is an empirical constant that depends on geometrical dimensions and the state of the gaseous flow, i.e. if it is laminar, turbulent or in a transition state. The coefficient is given by [82]

\[
C_f = \begin{cases} 
0.515 \left( \frac{l_g}{R} \right)^{0.3} \frac{RE_g^{0.3}}{RE_{0.3}} & \text{if } 500 < RE_g < 10^4 \\
0.0325 \left( \frac{l_g}{R} \right)^{0.3} \frac{RE_g^{0.3}}{RE_{0.2}} & \text{if } RE_g > 10^4 
\end{cases}
\]  

(4.35)

where \( l_g \) is the air gap length and \( RE_g \) is the \textit{Couette Reynolds number} as given by

\[
RE_g = \frac{l_g R \omega_m}{\nu}
\]  

(4.36)

where \( \nu \) is the kinematic viscosity. The windage losses of a rotating disc having outer and inner radii \( R_o \) and \( R_i \), respectively, can be written as [82]

\[ P_{\text{windage}} = \frac{1}{2} C_f \rho \omega_m \left( R_o^3 \omega - R_i^3 \omega \right). \]  

(4.37)

The friction coefficient is in this case given by [82]

\[
C_f = \begin{cases} 
\frac{2 \pi}{(l_g / R) \omega R E_r} & \text{regime I} \\
3.7 (l_g / R)^{0.1} \frac{RE_r^{0.5}}{RE_{0.2}} & \text{regime II} \\
0.08 (l_g / R)^{0.10} \frac{RE_r^{0.25}}{RE_{0.2}} & \text{regime III} \\
0.0102 (l_g / R)^{0.1} \frac{RE_r^{0.2}}{RE_{0.2}} & \text{regime IV} 
\end{cases}
\]  

(4.38)

where \( RE_r \) is the \textit{tip Reynolds number} as given by

\[
RE_r = \frac{R_o^2 \omega_m}{\nu}.
\]  

(4.39)
The different regimes I-IV in (4.38) are functions of the geometry of the enclosure of the disc and the condition of the gaseous flow. Figure 4.13 defines the different approximate flow regimes [82].

![Figure 4.13](image)

(a) Rotating disc in an enclosure. (b) The approximate flow regimes for an enclosed rotating disc.

### 4.3.3.2 Bearing losses

The bearing losses of a ball bearing are calculated as [82]

\[
P_{\text{bearing}} = C_b D_m^3 \omega_m, \tag{4.40}
\]

where \(C_b\) is the bearing coefficient, which is dependent on the bearing type and the load condition and often given by the manufacturer, and \(D_m\) is the average diameter of the bearing.

### 4.3.4 Losses in permanent magnets

Especially in high-speed or high-frequency applications, eddy currents can be induced in the permanent magnets. There are three major causes of these eddy current losses; one no-load and two load related sources [82]:

1. The losses induced at no-load are due to permeance variations, because of the slotting.
2. A major cause at load conditions is the space harmonics of the MMF due to the winding distribution, which are not in synchronism with the rotor and consequently create alternating magnetic fields in the permanent magnets and the rotor core.
3. Another source of eddy currents when loaded is induced by time harmonics in the current, which is mainly determined by the inverter.

Increasing the air gap or applying magnetic slot wedges can reduce or eliminate losses from the first source. The second source is the most influential and it is therefore important with a good winding layout if the operating frequency is high. This is the reason why machines with concentrated windings are not suitable for high-speed applications, which will be understood by reading Paper 6, Section 4.5 and Section 5.3.
Losses induced in the permanent magnets or in the rotor core are difficult to cool away. Moreover, the heat that is produced may cause demagnetization of the permanent magnets. It is therefore crucial to consider these effects in the design stage of the machine. The induced power losses in a single permanent magnet piece is given by

\[ P_{PM} = \frac{\sigma}{24} l_m \lambda_m \tau_m B^2 \omega^2 \]  

where \( \sigma \) is the conductivity, \( B \) is the flux density produced by the applied MMF or permeance variation, and \( \omega \) is the angular frequency of the eddy current. The permanent magnet dimensions \( l_m, \lambda_m \) and \( \tau_m \) are defined in Figure 4.14. The penetration depth is assumed to be larger than the length of the magnet \( l_m \), the flux is assumed to penetrate the magnet uniformly, and end effects are neglected. The derivation of (4.41) is done in [82]. A comprehensive analysis of eddy current losses in permanent magnets can be found in [83].

4.4 Field-weakening

If an electrical machine is going to be operated above base speed, it needs to be field-weakened. This is because the induced voltage almost equals the maximum source voltage at base speed and in order to increase the speed further the induced voltage needs to be kept constant. In a permanent magnet machine field-weakening is achieved by applying a negative d-current, which produces a flux in opposite direction to the permanent magnet flux. Phasor diagrams for a machine operating below the base speed, at the base speed and at twice the base speed illustrate the principle of field weakening in Figure 4.15.

Phasor diagrams for non-salient PM machine in both motoring and generating operation are shown in Figure 4.16. Different angles and parameters are defined. Permanent magnet machines can have a theoretical infinite constant power region, which was presented by [84]. Non-salient PM machines have finite speed if \( \Psi_m > L_s I_{lim} \) and theoretically infinite speed if \( \Psi_m \leq L_s I_{lim} \). Optimum current control strategies for maximum torque-per-ampere for the two machine types are illustrated by the synchronous flux linkage vector trajectories in Figure 4.17. The operating point of the non-salient PM machine must always be within the voltage and current limiting boundaries in Figure 4.17. Maximum synchronous flux linkage condition occurs at maximum torque operation, which is current limited. The constant torque region ends at base speed, because of voltage limitation, and is represented by point A. The synchronous flux linkage needs to be weakened above base speed in inverse proportion to
the speed in order to keep the fundamental voltage constant, which is represented by region A-B. Finite speed machines have reached their maximum speed at point B, where the q-current and consequently the torque is zero. For infinite speed machines, point B represents an operating point where the direct axis armature flux linkage precisely counteracts the magnet flux linkage. The d-current can therefore be kept constant in region B-C for the purpose of current minimization and reducing the q-current further increases the speed. Region B-C is a true constant power region. (The parameters used in Figure 4.15 to Figure 4.17 are defined in the List of symbols.)
Chapter 4

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4.5 Windings

4.5.1 Introduction

The most common winding type in electrical machines for industrial applications is the distributed winding. This winding type achieves an almost ideally sinusoidal magnetomotive force when the number of slots per pole per phase is high. The winding can also link all the flux from the rotor, if leakage is neglected, and is then called a full pitch winding. The many qualities of the winding make it the preferred choice in high efficiency machinery. However, it also has some drawbacks. Each phase coil has to cross other phase coils in the end regions, which increases the length of the end-windings and the axial build of the machine. This of course adds to the material need and to the winding losses. It is also quite complicated to manufacture the winding, especially to get it into the slots. The distributed winding is consequently relatively costly to manufacture. Another winding type is called concentrated windings. Concentrated windings are windings that consist of individual coils that are concentrated around individual teeth. The two main advantages of this winding type are short end-windings and the simple winding structure that is very suitable for cost effective automated manufacturing. The academic interest in permanent magnet machines with concentrated windings has increased dramatically in the last couple of years [85]-[100]. Principle sketches of distributed and concentrated windings are shown in Figure 4.18.

The theory of all different types of windings was thoroughly studied and documented in the time period 1890-1955. The main contributors were professor E. Arnold and his successor professor R. Richter from the Technischen Hochschule in Karlsruhe, followed by a student of Richter, H. Sequenz, who later became professor at the Technischen Hochschule in Wien. The electrical machine designers of today are still leaning on the winding theory established by these authors [101]-[102]. However, some special types of concentrated windings and their associated parasitic effects have up to now not been well documented. An analysis of these new types of concentrated winding machines is given below and in Papers 2, 3, 5 and 6.
4.5.2 Concentrated windings

4.5.2.1 Traditional brushless PM machines

The traditional winding configuration of a PM machine with concentrated windings has a number of slots per pole per phase of 0.5 and normally double layers per slot, i.e. one phase coil around each tooth. Some small spindle motors use a single layer per slot, i.e. one phase coil around every second tooth, because of simpler manufacturing. The relation of the slot and pole combinations for these three-phase machines can be written

\[ Q = \frac{3p}{2}, \] (4.42)

where \( Q \) and \( p \) is the number of slots and poles, respectively. In these types of machines the slot pitch is \( 2/3^{rd} \) the length of the pole pitch and consequently the stator windings are also only linking \( 2/3^{rd} \) of the flux produced by the permanent magnets in the rotor. The result is a fundamental winding factor of 0.87. The electromotive force and the torque production capability are proportional to this factor. Compared to the conventional distributed winding, which is linking almost all the flux from the rotor and therefore has a fundamental winding factor equal or close to one, this is an inherit drawback with the traditional concentrated windings. The strengths are cost effective manufacturing, short end-windings, high copper fill factor in slots, high thermal conductivity between winding and core and short axial build. These machines are therefore gaining popularity in applications that strongly emphasizes the described qualities like servomotors for robots and traction motors in electric or hybrid electric vehicles. The analysis of concentrated windings as described below deals with machines with constant slot widths. Machines can also be designed with variable slot widths [94].

4.5.2.2 Modular PM machines

There are means to increase the fundamental winding factor in machines with concentrated windings. In Paper 2 a number of interesting candidates are presented, for instance a machine with 14 poles and 15 slots with a fundamental winding factor of 0.95 using concentrated

\[ (a) \quad (b) \]

Figure 4.18: Winding layouts of permanent magnet machines with (a) distributed full pitch windings \((q = 1)\) and (b) concentrated fractional pitch windings \( (q = 0.5) \).
windings. This machine was designed, manufactured as a laboratory prototype and comprehensively analysed and is described in Paper 3, Paper 5, Paper 6 and in Chapter 5. Unlike traditional concentrated windings this machine has coils joining the same phase around consecutive teeth. Brushless PM machines with such a distinction are sometimes called “modular” [95]-[96]. Besides the high fundamental winding factor and excellent torque capability, the modular PM machines in general also have a manufacturing advantage, thanks to their smaller number of slots for a given number of poles, and low cogging and ripple torques, due to a high least common multiple (LCM) between the number of poles and slots. Many feasible slot and pole combinations for three-phase modular PM machines can be found, especially for increasing pole numbers. The number of slots must be divisible by three and the phase windings must fulfill the basic requirement of a three-phase winding, i.e. the resulting phase windings are displaced an electrical angle of 120 degrees from each other. When the number of slots per phase is odd, the feasible slot and pole combinations are

\[ Q = p \pm 1, \]

while the combinations when the number of slots per phase is even are given by

\[ Q = p \pm 2. \]

The coils forming one phase from combinations given by (4.43) must all be connected in series, since the EMF which is induced in each phase coil is not exactly in phase [96]. If they hypothetically were connected in parallel, this would result in circulating currents within the phase windings. The combinations from (4.44) however can be connected in series or series-parallel groups if the number of pole pairs is even or in series-antiparallel groups if the number of pole pairs is odd [96]. The combinations from (4.44) therefore offer greater flexibility in production.

Some disadvantages with the modular machine topology have been identified as described in Paper 6. Especially the 14 poles and 15 slots combination and furthermore odd slots per phase combinations have some parasitic effects like unbalanced radial magnetic forces and alternating magnetic fields in the rotor. In fact there are not that many feasible combinations for machines with few pole numbers that even have high fundamental winding factors, as shown in Table 1 and Table 2 in Paper 2. The most promising candidate with the lowest pole number is the 10 poles and 12 slots combination. In medium- to high-speed applications like traction, this seems to be the most attractive alternative in order to avoid high frequency problems in the machine core and with the control. This combination is therefore studied and compared with a traditional brushless PM machine with the same number of slots, but with 8 poles, in the next section. Both double layers and single layer windings are considered. The idea is to give an intuitive understanding of the typical performance of modular versus traditional brushless PM machines.

4.5.2.3 Comparative study of traditional and modular topologies

In this section a comparative study of traditional and modular brushless PM machines intended for sinusoidal excitation is made. The traditional winding configuration of a PM machine with concentrated windings has a number of slots per pole per phase of 0.5 and double layers per slot, i.e. one phase coil around each tooth. This topology, exemplified by machine IV in Figure 4.1, is the reference in the study. The other machines in the study are V, VI and VII as defined by Figure 4.1 and Table VI. They all have the same stator cores with 12 slots. The outer stator diameter is 105 mm, while the rotor diameter is 60 mm. The active length is 100 mm. The machines differ in their number of poles (8 or 10) and winding
configurations. The radii of the permanent magnets are slightly smaller than the air gap radius and the permanent magnets are diametrically magnetized in order to achieve more sinusoidal open-circuit voltages. The studied parameters are electromotive force (EMF), magnetomotive force (MMF), inductance, cogging and ripple torque, torque capability, radial magnetic forces and alternating flux densities in the rotor core. The inductance and torque computations are done with a current loading of 350 A/cm (RMS). This value of the current loading is theoretically regarded as being the rated working point of the machines, which in practice also is dependent of the cooling system.

4.5.2.3.1 Electromotive force (EMF)

The analytically calculated fundamental winding factors of machines IV-VII are 0.87, 0.87, 0.93 and 0.97, respectively. The finite element computed open-circuit voltages of the machines at an angular speed of 1000 rpm are shown in Figure 4.19. It is shown that the modular machines have higher fundamental EMFs than the traditional machines and their waveforms are also more sinusoidal, resulting in lower ripple torques.

![Figure 4.19: (a) FEM-computed open-circuit voltages at an angular speed of 1000 rpm of the four PM machines. (The curves of the 8 pole machines are on each other.) (b) The fundamental and harmonic components of the open-circuit voltages.](image)

4.5.2.3.2 Magnetomotive force (MMF)

The winding layouts and the FEM-computed magnetic flux distribution caused by the magnetomotive force only (the permanent magnets are inactive) for the four machines are shown in Figure 4.20. The analytically calculated total and fundamental three-phase MMF, with peak current in one phase, versus the pole pitch and harmonic content for the four machines are shown in Figure 4.21. The effect of slotting is disregarded, in order make the influence of the winding distribution more transparent. The MMFs are normalised in the sense that all four machines have the same current loading, i.e. the windings with double layers have one coil turn around every tooth with unity current amplitude, while the windings with single layer have two coils around every second tooth with unity current amplitude. Some interesting characteristics of machines with concentrated windings are illustrated.
Figure 4.20: The winding layout and magnetic flux distribution caused by the magnetomotive force only (the permanent magnets are inactive) for the machines with (a) 8 poles and double layers, (b) 8 poles and single layer, (c) 10 poles and double layers and (d) 10 poles and single layer.

All machines with concentrated windings have even MMF harmonics, because the space angle between the coil-sides in each phase is not constant. The machines with single layer windings have subharmonics that are greater than their fundamental MMF harmonic and they have overall higher harmonic contents than machines with double layer windings. The modular machines with double layer windings have subharmonics, while the traditional machines with double layer windings do not. The only desired MMF component is the fundamental, which is required to create torque. The rest of the components are undesired, but are evidently present with varying magnitudes in these types of machines. Their presence cause higher inductance, earlier saturation of the core, higher core losses, rotor losses and unbalanced radial magnetic forces or noise.
4.5.2.3.3 Torque capability and ripple torque

The finite element computed ripple torques at a current loading of 350 A/cm (RMS) and the cogging torques are shown in Figure 4.22. The windings are fed with purely sinusoidal currents. The cogging torque of the traditional machines is higher than for the modular machines, due to the lower value of the least common multiple between the pole and slot numbers, which is 24 compared to 60. The peak-to-peak cogging torque is approximately 4% of the rated torque for the traditional machines, while the cogging torque for the modular machines is practically zero. As expected from the study of the electromotive forces, the traditional machines also have higher ripple torques, because of their more influential 5th and 7th harmonic that interacts with the fundamental current component and produce a 6th torque harmonic component (6 times the fundamental electrical frequency). The average torque of both the modular machines is higher though, due to their higher fundamental open-circuit voltages. Traditional machines sometimes have their permanent magnets stepwise skewed on the rotor shaft to reduce the cogging and ripple torques. This operation influences the fundamental winding factor negatively and is also increasing the complexity in the
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manufacturing process. Skewing of modular machines seems superfluous, since their ripple torques are very low by nature.

Figure 4.22: The finite element computed ripple torques at a current loading of 350 A/cm (RMS) and the cogging torques.

4.5.2.3.4 Inductance
The FEM-computed unsaturated inductances at a current loading of 350 A/cm (RMS) and 100 turns per phase are shown in Figure 4.23. The machines are consequently operated at the same phase current. The end-winding inductance is neglected. The permanent magnets are inactive in the study, so the magnetic energy in the model is caused by current sources only. The synchronous phase inductance is found by calculating the total magnetic energy in the complete active machine volume, and given by

\[ L_s = \frac{2W}{3M_{ph}^2} \cdot N_{ph}^2, \]  \hspace{1cm} (4.45)

where \( W \) is the total magnetic energy, \( N_{ph} \) is the number of phase turns and is expressed

\[ N_{ph} = \frac{Pq}{2c_s} \cdot N_i \cdot N_r, \]  \hspace{1cm} (4.46)

where \( q \) is the number of slots per pole per phase, \( c_s \) is the connection factor, \( N_i \) is the number of winding layers per slot (1 or 2) and \( N_r \) is the number of coil turns around a tooth. \( M_{ph} \) is the total magnetomotive force per phase (RMS), as given by

\[ M_{ph} = N_{ph}I, \]  \hspace{1cm} (4.47)

where \( I \) is the phase current (RMS). It is more usual for conventional machines to use the term magnetomotive force per slot, but since some modular machines do not have the same peak MMF in all the slots it is more logical to introduce the term magnetomotive force per phase instead. The inductances vary less than 0.7% as a function of the rotor position for all four machines. The inductances of the machines with single layer windings are 51% higher, while the inductance for the 10 poles machine with double layer windings is 8% higher, than for the 8 poles machine with double layer windings. The differences are due to the differences in the
harmonic contents of the total three-phase magnetomotive forces. A higher leakage inductance is equivalent to a lower overload capability, but also equivalent to a wider constant power operation region.

![Image](image_url)

**Figure 4.23:** The finite element computed unsaturated phase inductances with 100 turns per phase at equal current loadings.

### 4.5.2.3.5 Parasitic effects

The undesired alternating magnetic fields in the rotor due to the non-symmetrical MMF in the 14 poles 15 slots machine topology are described in Paper 5 and Paper 6. It is therefore important to investigate if the same problem is prominent in concentrated winding machines in general or if it is only limited to certain types of modular PM machines. The FEM-computed tangential flux density component at a fixed point on the rotor core at a radius of 22 mm in between two permanent magnets, when the rotor is rotated two pole pitches at a current loading of 350 A/cm (RMS) is shown for each of the four studied machines in Figure 4.24.

![Image](image_url)

**Figure 4.24:** The FEM-computed tangential flux density component at a fixed point on the rotor core at a radius of 22 mm in between two permanent magnets, when the rotor is rotated two pole pitches at a current loading of 350 A/cm (RMS).
The traditional machine with double layer windings is not especially sensitive to the alternating fields in the rotor, as seen in Figure 4.24. However the traditional machine with single layer windings has a considerable flux density variation in the rotor. The modular machine with double layer windings has also got a moderate alternating field. The modular machine with single layer windings has a variation that is quite severe. The reason for the different behaviours can be intuitively understood by considering the different MMF waveforms in Figure 4.21. The air gap flux density caused by the armature reaction has the same waveform as the total three-phase MMF. The armature reaction interacts with the permanent magnet produced flux density to form the resulting air gap flux density. Since the MMF of the traditional machine with double layer windings is pretty regular, the air gap flux and the rotor flux is also regular. But for the machines with single layer windings in particular, the MMF waveforms are highly irregular, and consequently cause the alternating fields in the rotor. In order to avoid considerable rotor losses in the modular machine with double layer windings it is perhaps sufficient to laminate the rotor core and manufacture the permanent magnets in many pieces to reduce the otherwise induced eddy current losses. For the machines with single layer windings this remedy is probably not even enough to avoid reduced machine efficiencies. Furthermore, the alternating magnetic fields in the rotor also influences the dynamical torque performance negatively, since fast and large current steps will generate high rotor power losses and thereby reduce the output torque.

Irregular MMF waveforms can also cause unbalanced radial magnetic forces and magnetic noise. Consider the layouts of the phase coils and the flux distributions caused by the magnetomotive forces in Figure 4.20. The machine in (a) has one quarter of the circumference between the coils in the same phase, while the phase coil displacement is one half of the circumference for the other three machines. This means that the radial magnetic force is in balance for every quarter of the air gap circumference for the machine in (a), while it is in balance only for every half of the circumference for the other machines. The traditional machine with double layer windings will therefore have the lowest magnetic noise of the four machines. The radial magnetic stresses in the air gap at rated current are shown in Figure 4.25. The effect of unbalanced radial forces can be exemplified by considering the stator or rotor to be a metal ring with different radial vibration mode shapes as illustrated in Figure 4.26. A unidirectional deflection of the stator and rotor for instance represents a radial vibration shape of mode one [103]. Some other modes are shown in the same figure. The radial forces usually cause undesired magnetic noise. The sound level depends on several factors like rigidity of the machine parts, machine dimensions and location of the natural frequencies of the machine structures. Small machines are in general not so sensitive to higher vibration modes as big machines, but especially mode number one can be influential even in this case, because relatively low vibration frequencies can easily coincide with the natural frequencies of the machine structures and result in resonance.

The radial magnetic force densities on the rotor core for the current loading 350 A/cm (RMS) at four different steps in time, while the rotors are rotating, for the four machines are shown in Figure 4.27. It is shown that the force density is symmetric for each quadrant of the machine with 8 poles and double layer windings. The force density for the machine with 10 poles and double layer windings is balanced at the opposite side of the air gap, but has a small mode 2 vibration mode. The machines with single layer windings have more alternating magnetic stress, as illustrated in Figure 4.27.
Figure 4.25: Radial magnetic stress in the air gap for the current loading 350 A/cm (RMS) for the machines with (a) 8 poles and double layers, (b) 8 poles and single layer, (c) 10 poles and double layers and (d) 10 poles and single layer.

Figure 4.26: Some radial vibration mode shapes for the stator and rotor.
Figure 4.27: Radial magnetic stress on the rotor core for the current loading 350 A/cm (RMS) at four different steps in time, while the rotor is rotating, for the machines with (a) 8 poles and double layers, (b) 8 poles and single layer, (c) 10 poles and double layers and (d) 10 poles and single layer.

4.5.2.3.6 Summary
A comparative analysis of the traditional brushless and modular PM machines with both double and single layer windings has been performed. Since the stator and rotor cores of the studied machines have been identical, several parameters that influence the machine performance have been characterized in relative merits. The results are summarized in Table XI. The traditional machine with double layer windings, which here is the reference machine, has lower torque capability and higher torque ripple, but less influential parasitic effects like rotor losses and noise radiation than modular machines. Machines with single layer windings have been found inferior in state-of-the-art applications due to their lower efficiency and higher inductance compared to the double layer winding machines. Furthermore, their end-windings and axial build are also longer. They are nevertheless sometimes used in fractional power applications, because of their fewer numbers of phase coils that leads to simpler manufacturing. The modular machine with double layer windings shows promising characteristics concerning high torque capability and low ripple torque. However, this
machine configuration may have some undesired parasitic effects, which at least must be taken into account in an early design stage. Some helpful manufacturing techniques are to also laminate the rotor core, divide the permanent magnets in many pieces per pole and make the cores very rigid. By introducing these measures, the modular machines with double layer windings may actually outperform the traditional brushless permanent magnet machines.

Table XI: Comparative results from the study between the traditional and modular permanent magnet brushless machines.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>8 poles double layers</th>
<th>8 poles single layer</th>
<th>10 poles double layers</th>
<th>10 poles single layer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Average torque</td>
<td>100%</td>
<td>98%</td>
<td>106%</td>
<td>108%</td>
</tr>
<tr>
<td>Peak-to-peak cogging torque to average torque ratio</td>
<td>4%</td>
<td>4%</td>
<td>~0%</td>
<td>~0%</td>
</tr>
<tr>
<td>Peak-to-peak ripple torque to average torque ratio</td>
<td>10%</td>
<td>15%</td>
<td>3%</td>
<td>4%</td>
</tr>
<tr>
<td>Unsaturated phase inductance</td>
<td>100%</td>
<td>151%</td>
<td>108%</td>
<td>151%</td>
</tr>
<tr>
<td>Lowest (most important) radial vibration mode</td>
<td>Mode 4</td>
<td>Mode 2</td>
<td>Mode 2</td>
<td>Mode 2</td>
</tr>
<tr>
<td>Sound radiation (magnetic noise)</td>
<td>Moderate</td>
<td>Considerable(^1)</td>
<td>Considerable(^1)</td>
<td>Considerable(^1)</td>
</tr>
<tr>
<td>Alternating magnetic fields in rotor</td>
<td>Negligible</td>
<td>Considerable</td>
<td>Moderate</td>
<td>Severe</td>
</tr>
</tbody>
</table>

\(^1\)Considerable means that there is a risk for magnetic noise, but the magnitude depends on the strength of the stator and rotor cores.

### 4.6 Lifetime of electrical machines

There are basically three different components that limit the lifetime of permanent magnet machines, and these are the winding insulation, the bearings and the permanent magnets. Insulation systems and bearings are going through ageing processes, which are strongly related to the operating temperature. The lifetime of the permanent magnets is also dependent on temperature in combination with magnetic conditions. The three main lifetime-limiting sources are described below.

#### 4.6.1 Winding insulation

The winding insulation is exposed to a thermal ageing process, even when the machine is not running. At one extreme it is ageing at a rate caused by the ambient temperature, while at the other extreme it is ageing at the rate caused by the highest operating temperature. The ageing is due to processes like oxidation, loss of volatile product, molecular polymerization, reaction to moisture, chemical breakdown, and dielectric, mechanical and environmental stresses [104]. After some time these effects make the insulation break down until the winding is short-circuited between turns or to ground. At this time the insulation system has failed by definition and the machine is not longer functioning. Figure 4.28 shows the expected lifetime of a winding system versus the operating winding temperature for the insulation classes A, B, F and H. It is assumed that the insulation life doubles for a 10 °C decrease in temperature [104].

The expected lifetime given in Figure 4.28 is based on sinusoidal power supplies for standard motors. When the power supply is an inverter, the lifetime is reduced by a factor that is dependent on switching frequency, DC-link voltage, intermittent operation and other parameters that makes it very complex to estimate the degree of derating. The maximum allowed temperature rises for the different insulation classes at a maximum ambient temperature of 40 °C, according to the standard NEMA MG-1-12.44, are given in Table XII [104]. The newer standard IEEE Std. 841-2001 says that the average temperature rise of any phase of the stator winding shall not exceed 80 °C, measured by winding resistance [104]. The idea behind this tougher limitation is to increase the expected lifetime of the insulation system of motors in industrial applications. Machines intended for electric vehicle propulsion
have normally a much shorter required lifetime in comparison, since vehicles (at least passenger cars) are only used a couple of hours every day, while industrial machines run almost continuously.

![Graph showing expected lifetime of winding system versus winding temperature for different temperature classes: (from left) A, B, F, and H.]

**Figure 4.28:** Expected lifetime of winding system versus winding temperature for the temperature classes: (from left) A, B, F, and H.

**Table XII:** The allowed temperature rise according to NEMA MG-1-12.44.

<table>
<thead>
<tr>
<th>Class of insulation system</th>
<th>A</th>
<th>B</th>
<th>F</th>
<th>H</th>
</tr>
</thead>
<tbody>
<tr>
<td>Temperature rise, [K]</td>
<td>60</td>
<td>80</td>
<td>105</td>
<td>125</td>
</tr>
</tbody>
</table>

### 4.6.2 Bearings

The lifetime of bearings is based on a number of factors including winding temperature, lubricant temperature, motor thermal circuit, cooling method, oil and grease viscosity, bearing seals and shields, lubricant type, amount of grease, radial internal clearance, ambient conditions and contamination, loading and speed, bearing type and size [104]. A general guide with temperature limits for safe operation of both standard and synthetic bearings is given in Table XIII [104]. The actual temperatures of bearings will usually be lower than those shown in the table for most applications.

**Table XIII:** Recommended temperature ranges for safe operation of bearings.

<table>
<thead>
<tr>
<th>Operating range</th>
<th>Normal</th>
<th>Alarm</th>
<th>Shutdown</th>
</tr>
</thead>
<tbody>
<tr>
<td>Standard lubricants</td>
<td>80-90 °C</td>
<td>90-95 °C</td>
<td>100-105 °C</td>
</tr>
<tr>
<td>Synthetic lubricants</td>
<td>110-115 °C</td>
<td>120-125 °C</td>
<td>130-135 °C</td>
</tr>
</tbody>
</table>

The lifetime of bearings is also reduced if the bearings are exposed to bearing currents. The awareness of bearing current flow has risen with the increased use of variable speed drives. The switching characteristic of the inverter power source has introduced a new source of
capacitively coupled bearing current that can flow through paths that would normally be considered to be insulators, like wire insulation, stator slot liners, motor air gap, bearing grease, and stator slot top sticks [105]. However, the phenomenon of bearing currents is not new, even sine-wave-driven electrical machines have experienced bearing currents since they were invented. The main cause behind the bearing currents in across-the-line powered machines is magnetic asymmetry. The asymmetry generates alternating flux that is linking the shaft, when the machine is operated, that can drive a current down the shaft, to the bearing, through the frame, and back again through the other bearing [105]. Some solutions are available to reduce or eliminate the current flows, and each remedy is dependent upon the source and type of bearing current. Three main sources of the bearing current flow are [105]:

A. Internally circulating currents due to magnetic asymmetry (normally low-frequency),
B. Inverter generated common mode (ground) current induced by fast voltage PWM-switching (dV/dt) taking a return path via the motor shaft extension and coupled equipment (high-frequency),
C. Inverter generated discharge through machine or load bearing of the shaft voltage, which is capacitively coupled to the common mode voltage (high-frequency).

The main remediation methods for bearing currents that will reduce or eliminate them are [105]:

a. Well terminated cable ground connections between the drive and the machine (B),
b. Bonding strap between the machine and the frame of the load equipment (B),
c. One insulated bearing on opposite drive end (A),
d. Two insulated bearings (A, B),
e. Shaft grounding brush across one bearing (C),
f. Faraday shield on the stator air gap surface (C),
g. Insulated coupling between machine and load shaft (B).

The best remedy to use of a-g is dependent on the type of source of A-C. The sources that each remedy has a positive impact on are listed in the parentheses above [105]. In some cases several remedies must be applied in order to eliminate the bearing currents. The problem is complex, since one remedy may actually increase the currents in another part of the system. Another strategy in variable speed drives is to work on the source of the problem directly, by making the inverter less harmful for the bearings. Some comprehensive analyses of bearing currents caused by inverters are given in [106]-[108].

4.6.3 Permanent magnets

Permanent magnets are sensitive to high temperatures in combination with large opposing magnetic fields and they can be irreversibly demagnetised if loaded improperly. This can and should normally be avoided already in the design stage by ensuring that the magnets are operated above the “knee” in the BH-curve for all working conditions. Figure 4.29 illustrates the influence of temperature on the demagnetisation curves of permanent magnets (here exemplified with NdFeB rare-earth permanent magnets). When the temperature is increasing the BH-curve is moving to the left in the second quadrant and the remanence is decreasing. Simultaneously, the “knee” of the BH-curve is moving upwards, which makes the material more sensitive to demagnetising fields. Let for instance a permanent magnet machine be operated along the working line as exemplified with the points $P_1$, $P_2$, $P_3$ and $P_4$ in Figure 4.29, when the temperature is increasing. The three first working points are on the straight
region of the BH-curve and the original remanence will be experienced again when the temperature drops to the original temperature. However, $P_d$ is a working point that is not longer on the straight region, and the remanence will never go back to its original value (unless the permanent magnets are remagnetised). The machine has consequently lost in performance. Dependent on how far down on the knee the working point occurred, the machine may still be functioning satisfactory or become completely useless. In the latter case this accident has defined the lifetime of the machine.

Figure 4.29: The influence of temperature on the demagnetisation curves of permanent magnets.

Permanent magnets are also susceptible to corrosion, but they can be made resistive against this process in the production by for instance surface coating of the magnets, so that no reaction with oxygen occurs.

A basic understanding of the temperatures and their impacts on the different parts of a machine can help achieve satisfactory lifetime and performance. Lower operating temperatures normally translate into longer life and higher efficiencies, but also into higher purchase cost, since a larger machine and more material are needed. A compromise must therefore be chosen that takes both the required lifetime and the cost sensitivity of the costumer into account.
5 Four Quadrant Transducer drivetrain

5.1 The system

The Four Quadrant Transducer drivetrain is presented in Paper 1. This section summarizes the characteristics of the system based on that paper and some additional reflections.

5.1.1 Introduction

The Four Quadrant Transducer is a power-split hybrid electric drive system and its main components are shown in Figure 5.1. The main components are two electric machines, two inverters and one slipring unit. The basic characteristics of the system are based on some of the advantages from the series and parallel hybrid systems. The internal combustion engine (ICE) is for instance not mechanically connected to the wheels, which allows for an independent choice of the engine speed in relation to the required vehicle speed.

![Figure 5.1: A simplified sketch of the Four Quadrant Transducer HEV propulsion system.](image)

In the 4QT hybrid system the torque from the ICE is supported from the electrical machine unit, which means that the ICE can operate at torque values fairly independent of the torque requirements at the wheels. The power ratings of the electrical machine unit and inverters are typically lower than the power of the ICE. The rated battery power can furthermore be chosen very low, since the electrical machine unit can operate as motor and generator simultaneously. The gearbox can be of a simple single-stage type, since only the electrical machine is mechanically connected to the wheels.

The electrical machine unit of the 4QT consists of two parts. One part is called the speed-changing machine or historically Double Rotor Machine (DRM), which is equipped with windings connected via slip rings. It is mechanically coupled to the shaft of the ICE. The
second part is called the torque-changing machine or historically the Stator (ST) machine, which is also equipped with windings. The DRM and the ST have one rotor in common, which is mechanically connected to the wheels via a gearbox. This rotor is preferably of permanent magnet or reluctance type.

The mission of the speed-changing machine is to make the speed of the ICE independent from the speed requirement at the wheels. This is possible by applying an AC current to its windings. The machine must be designed for the maximum torque of the ICE, since the torque of this electrical machine equals the torque of the ICE at steady-state. The mission of the torque-changing machine is to add or subtract the torque delivered by the ICE. The principle of operation is illustrated by the power flow diagram in Figure 5.2.

![Power flow diagram of the Four Quadrant Transducer hybrid system.](image)

**Figure 5.2: Power flow diagram of the Four Quadrant Transducer hybrid system.**

The power delivered to the wheels of the vehicle is given by (the component losses are neglected for convenience)

\[
P_{\text{wheels}} = P_{\text{load}} = P_{\text{ICE}} + P_{\text{Battery}} = P_{\text{ICE}} + P_{\text{DRM}} + P_{\text{ST}}. \tag{5.1}
\]

The torque delivered to the wheels is given by

\[
T_{\text{wheels}} = G \cdot T_{\text{load}} = G \cdot [T_{\text{ICE}} + T_{\text{ST}}] \tag{5.2}
\]

where \( G \) is the gear ratio. The angular speed of the wheels is given by

\[
\omega_{\text{wheels}} = \frac{1}{G} \omega_{\text{load}} = \frac{1}{G} \omega_{\text{ST}} = \frac{1}{G} [\omega_{\text{ICE}} + \omega_{\text{DRM}}]. \tag{5.3}
\]

The wheel power can also be expressed

\[
P_{\text{wheels}} = \omega_{\text{wheels}} T_{\text{wheels}} = [\omega_{\text{ICE}} + \omega_{\text{DRM}}] [T_{\text{ICE}} + T_{\text{ST}}]. \tag{5.4}
\]

Equation (5.4) shows that it is possible to operate the ICE at a desired speed and torque that differs from the required working point at the wheels, simply by adjusting the speed difference between the two shafts of the 4QT with the DRM and the torque with the ST. Every ICE has an optimal operation line (o.o.l.), torque versus engine speed, regarding fuel consumption. Figure 5 illustrates how the 4QT enables the ICE to operate on its optimal
operation line for two different load conditions with equal power requirements. From the working point of the ICE it is possible to move into any of the four quadrants with the assistance of the hybrid system, independent of load requirements. This function has also given the name to the Four Quadrant Transducer.

Figure 5.3: A diagram illustrating the flexibility of the 4QT in keeping the speed and the torque of the ICE on its optimal operation line independently of the changing load requirements.

The 4QT can be categorized as a power-split hybrid system, due to its characteristic simultaneous motoring and generating capabilities. The previously presented power split systems [109]-[111] are strongly dependent on mechanical transmissions like a planetary gearbox or a Continuously Variable Transmission (CVT). The 4QT and similar systems (see below) is unique in the sense that no complex mechanical transmission is needed, but still the advantages of the power split system are exploited. A very similar system called the Electrical Variable Transmission (EVT), based on two concentric induction machines, is studied in [112]. The idea of using two concentrically arranged electrical machines with two rotors is not new. There is a patent from 1935 that explains this configuration using DC machines [113].

5.1.2 Simulations

A complete hybrid electric vehicle system is simulated in Matlab/Simulink. The basic data are shown in Table XIV. The simulated vehicle is a passenger car and the gearbox is of the simple single stage type.

Table XIV: The basic data of the hybrid electric vehicle used in the simulations.

<table>
<thead>
<tr>
<th>Feature</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Engine power</td>
<td>80 kW @ 5750 rpm</td>
</tr>
<tr>
<td>Engine volume</td>
<td>1600 cm³ (4 cyl., 16 valves)</td>
</tr>
<tr>
<td>Mass of vehicle</td>
<td>1380 kg (incl. 2 persons)</td>
</tr>
<tr>
<td>Gearbox</td>
<td>1 speed (4,0:1)</td>
</tr>
<tr>
<td>Acceleration</td>
<td>13,0 s (0-100 km/h)</td>
</tr>
<tr>
<td>Top speed</td>
<td>180 km/h</td>
</tr>
</tbody>
</table>
A gear ratio of 4:0:1 is chosen to achieve low electrical machine power at maximum vehicle speed. A mechanical lock-up device between the ICE shaft and the outgoing 4QT shaft is possible. This would remove most of the losses from the electrical components at highway driving. At highway driving, high continuous power from the ICE is needed and the 4QT system will therefore not be able to reduce the consumption. A lock-up device is not used in the simulations, however. A more advanced multi speed gearbox can also be used to reduce the size of the 4QT machines. Only a single speed gearbox is used in the simulations to make it as simple and cost effective as possible. Three driving cycles according to Table XV are simulated.

**Table XV: Basic data of the simulated driving cycles.**

<table>
<thead>
<tr>
<th>Driving cycle</th>
<th>EC</th>
<th>Japan 10-15</th>
<th>FTP 75</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nationality</td>
<td>European</td>
<td>Japanese</td>
<td>US American</td>
</tr>
<tr>
<td>Speed curve</td>
<td>Created</td>
<td>Created</td>
<td>Measured in Los Ang.</td>
</tr>
<tr>
<td>Type</td>
<td>Urban/city</td>
<td>Urban/city</td>
<td>Urban/city</td>
</tr>
<tr>
<td>Cycle time</td>
<td>1220 s</td>
<td>635 s</td>
<td>1877 s + 600 s pause</td>
</tr>
<tr>
<td>Distance</td>
<td>11.0 km</td>
<td>4.2 km</td>
<td>17.9 km</td>
</tr>
<tr>
<td>Aver. speed</td>
<td>32.5 km/h</td>
<td>22.7 km/h</td>
<td>34.1 km/h</td>
</tr>
<tr>
<td>Max. speed</td>
<td>120.0 km/h</td>
<td>70.0 km/h</td>
<td>91.2 km/h</td>
</tr>
</tbody>
</table>

The simulated vehicle speed for the driving cycles EC, Japan 10-15 and FTP 75 are shown in Figure 5.4.

![Figure 5.4: The simulated vehicle speed (in km/h) versus time for the driving cycles (a) EC, (b) Japan 10-15 and (c) FTP 75.](image)

As can be seen in Figure 5.4, the three driving cycles have all a relatively low average speed and the speed is transient. These are the best conditions for the 4QT system to be able to reduce the fuel consumption. At highway driving the potential is completely different. This is because high continuous power is needed from the ICE, as mentioned earlier, and the vehicle speed is relatively constant. The working point of the ICE can no longer be chosen freely. The fuel consumption can therefore not be reduced and the power flow in the electrical components should be low in order to have low losses in the system. The greatest fuel savings are achieved at city traffic driving as for most hybrid systems. The simulation results of the fuel consumption for the 4QT system are shown in Figure 5.5.

The 4QT system is compared with a similar conventional vehicle (ICE only) with the same weight. The fuel saving is 40%, 44% and 43% for the driving cycles EC, Japan 10-15 and FTP 75, respectively. The 4QT drive system, the traction battery and the single speed gearbox in the novel hybrid replace the five-speed gearbox, the engine flywheel, the start motor and the generator in the conventional vehicle. The weight difference can be neglected if a small traction battery is used. A detailed analysis of the system behaviour is shown in Table XVI.
Figure 5.5: Calculated fuel consumption of the 4QT system compared to the conventional vehicle.

Table XVI: Analysis of the system behaviour for three driving cycles.

<table>
<thead>
<tr>
<th>Driving cycle</th>
<th>EC</th>
<th>Japan 10-15</th>
<th>FTP 75</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fuel consumption, corrected for SOC [l/100 km]</td>
<td>4.1</td>
<td>3.6</td>
<td>4.0</td>
</tr>
<tr>
<td>Fuel saving compared to conv. vehicle [%]</td>
<td>40</td>
<td>44</td>
<td>43</td>
</tr>
<tr>
<td>Duration in electric mode [s]</td>
<td>946.8</td>
<td>559.9</td>
<td>2064.6</td>
</tr>
<tr>
<td>Driven distance in electric mode [km]</td>
<td>5.4</td>
<td>3.0</td>
<td>11.3</td>
</tr>
<tr>
<td>Average speed in electric mode [km/h]</td>
<td>20.5</td>
<td>19.3</td>
<td>19.7</td>
</tr>
<tr>
<td>Duration in hybrid mode [s]</td>
<td>273.2</td>
<td>75.1</td>
<td>412.4</td>
</tr>
<tr>
<td>Driven distance in hybrid mode [km]</td>
<td>5.6</td>
<td>1.2</td>
<td>6.4</td>
</tr>
<tr>
<td>Average speed in hybrid mode [km/h]</td>
<td>74.1</td>
<td>55.9</td>
<td>55.5</td>
</tr>
<tr>
<td>Maximum battery power during cycle [kW]</td>
<td>13.9</td>
<td>13.6</td>
<td>15.8</td>
</tr>
<tr>
<td>Minimum speed of DRM [rpm]</td>
<td>-937</td>
<td>-793</td>
<td>-1285</td>
</tr>
<tr>
<td>Maximum speed of DRM [rpm]</td>
<td>4020</td>
<td>2341</td>
<td>3113</td>
</tr>
<tr>
<td>Maximum torque of DRM [Nm]</td>
<td>135.2</td>
<td>97.9</td>
<td>135.1</td>
</tr>
<tr>
<td>Minimum power of DRM [kW]</td>
<td>-8.5</td>
<td>-7.1</td>
<td>-16.8</td>
</tr>
<tr>
<td>Maximum power of DRM [kW]</td>
<td>22.9</td>
<td>7.5</td>
<td>12.7</td>
</tr>
<tr>
<td>Maximum speed of ST [rpm]</td>
<td>4232</td>
<td>2484</td>
<td>3221</td>
</tr>
<tr>
<td>Minimum torque of ST [Nm]</td>
<td>-87.9</td>
<td>-83.5</td>
<td>-111.3</td>
</tr>
<tr>
<td>Maximum torque of ST [Nm]</td>
<td>128.8</td>
<td>105.2</td>
<td>157.6</td>
</tr>
<tr>
<td>Minimum power of ST [kW]</td>
<td>-33.0</td>
<td>-21.0</td>
<td>-27.0</td>
</tr>
<tr>
<td>Maximum power of ST [kW]</td>
<td>11.7</td>
<td>11.6</td>
<td>13.6</td>
</tr>
</tbody>
</table>

The duration in electric mode in Table XVI is defined as the total cycle time minus the duration in hybrid mode. I.e. it includes even the time when the vehicle is at stand still. Hybrid mode means that the ICE is on. It can be seen in Table XVI that the ICE is frequently turned off. The strategy is to turn off the engine at low speed and load, when braking and at stand still. For the simulated driving cycles, which have relatively low power demands, the vehicle is run in electric mode for the majority of time. Furthermore, the required battery power is lower than 16 kW. One machine generates 33 kW at the most, but since the other machine simultaneously acts as motor, the battery only needs to handle the power difference and can consequently be dimensioned for lower power than the electrical machines, respectively.

If the working points of the electrical machines are mapped for a lot of driving cycles and maximum acceleration conditions, overall operating windows can be found, as shown in Figure 5.6. The simulated mapping diagrams for the three driving cycles plus an additional European highway driving cycle are shown in Appendix A.
The speed of the DRM is a slip speed and equal to the speed difference between the out- and ingoing shafts of the 4QT. The maximum torque and power are transient values. No field-weakening is needed for the DRM, but the ST needs to be field-weakened above 2000 rpm, according to Figure 5.6. The average required performance of the electrical machines of the 4QT is calculated in two minutes intervals for different driving cycles. The information is used to estimate the required continuous performance of the machines as shown in Table XVII.

**Table XVII: Estimated continuous performance of the 4QT.**

<table>
<thead>
<tr>
<th>Machine</th>
<th>DRM</th>
<th>ST</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>Torque</td>
<td>120</td>
<td>60</td>
</tr>
<tr>
<td>Speed</td>
<td>2000</td>
<td>4000</td>
</tr>
</tbody>
</table>

It is evident when comparing Table XVII with the overall operation windows, that the required continuous performance is much lower than the maximum performance. This is especially the case for the torque of the ST, which has a maximum of 170 Nm, but a continuous value of only 60 Nm. Such data can be used to cost optimise the electrical machines. It is well known that electrical machines are magnetically designed for the required transient performance, respectively thermally designed for the required continuous performance.

### 5.1.3 Different machine topologies

The available space for the electrical machines in a hybrid electric system for passenger cars is very limited. Front wheel drive is most common today. This makes the axial build of the machines limited by the width of the engine compartment, which also shall accommodate the engine and the gearbox. The distance to the transmission shafts for the wheels and the chosen ground clearance normally limits the outer diameters of the machines. Furthermore, there is of course always a requirement in a hybrid vehicle that all components shall have as low weight as possible. These requirements lead to the conclusion that the electrical machine unit needs to be compact. The available overall space envelope is defined as 260 mm in diameter and 220 mm in length for the total Four Quadrant Transducer machine unit, after discussions with representatives from a manufacturer of mid-sized passenger vehicles.
The electrical machines can have many different topologies and three examples are illustrated in Figure 5.7. The preferred topology depends on the maximum allowable machine dimensions and the required machine performance. The radial-radial flux topology in Figure 5.7 seems like the obvious choice if the dimensions can be defined arbitrarily, due to the concentric layout of the machines that results in a high utilization of the available volume. The axial-axial topology may be advantageous if the available volume requires very short axial length, but where the diameter can be chosen more freely. The axial/radial-radial flux topology is a combination of the other two. The magnetic circuits of the two machines in this alternative are not coupled, which means that the machines can be treated independently in opposite to the other two alternatives. The pole numbers of the machines can therefore be different, which could be an advantage if the normal operating speeds of the machines are far apart. The axial/radial-radial flux topology is chosen in this project. The motivation for this is not only due to space requirements, but also of academic reasons. The topology requires studies on different academically interesting electrical machine aspects like axial and radial flux machines and material characteristics (Soft Magnetic Composites and silicon iron laminated iron).

![Figure 5.7: Various electrical machine topologies for the 4QT propulsion system.](image)

### 5.2 Speed-changing machine

The speed-changing machine has been analysed in Paper 3 and Paper 9. This section discusses and summarizes the evaluation of the machine concept.

#### 5.2.1 Design philosophy

The speed-changing machine in the Four Quadrant Transducer must deliver the same torque as the internal combustion engine at stationary conditions. The equivalent speed the machine operates at is relatively low (500-2000 rpm), since it is just the angular speed difference between the output shaft and the input shaft. The machine needs to be compact, especially in the axial direction, because the available space in the engine compartment is very limited. Effort has therefore been made to design a machine with a large air gap surface in relation to the active volume. The specification can be achieved by introducing both axial and radial air
gaps. The core material must therefore carry flux in three dimensions. This kind of topology was first reported in [114], where a slotless prototype for an electric bicycle was built. The machine for the Four Quadrant Transducer drive system increases the concept’s torque density capability by adopting a slotted structure instead, which increases the magnetic loading. In order to increase the torque density even further, a rectangular conductor is toroidally oriented around the back in each slot, called Gramme-winding, making a coil with short end-windings and high slot fill factor, which increases the electric loading as well. This winding is a full-pitch winding that theoretically links all the induced flux from the rotor, in opposite to a concentrated winding that is inferior in this respect.

5.2.2 Electromagnetic analysis

5.2.2.1 Geometrical parameters

The rare-earth permanent magnets of Neodymium-Iron-Boron (NdFeB) in the rotor with sixteen poles produce air gap flux in both the axial and radial directions. The topology is not suitable for low pole numbers (<10), since the yoke area is getting larger with decreasing pole numbers, i.e. smaller inner diameter and longer end-windings. Forty-eight segments build up the core that consists of teeth and yoke. Due to the segmentation, which allows for the rectangular conductor shape, the slot fill factor can be extremely high (0.78) and the overall machine size will not be limited by the capability of the tool for pressing the iron powder. The iron powder used is the SOMALOY™ 500 with 0.5% of the lubricant Kenolube from the manufacturer Höganäs AB [115]. The machine is illustrated by the CAD-drawings in Figure 5.8. Since the yoke is shorter than the active axial length and even the axial surfaces are active in the torque production, this design has short end-windings. The length of the end-windings is approximately 20% of the active length in the presented pancake design. The radial air gap radius, the active length, the outer and inner radii of the axial air gaps of the presented machine is 105, 66, 104 and 73 mm, respectively. The air gap length is 1.5 mm and the length of the permanent magnets are 3.5 mm. The thickness of the rotor back is 10 mm. Each phase winding is connected in 4 parallel circuits, each with 4 coils connected in series, and the phase winding therefore consists of 16 coils. The phase windings are star-connected. (The phase distribution has the following pattern: +A, -C, +B, -A, +C, -B, +A, etc., with one slot per pole per phase.)

![Figure 5.8: The speed-changing electrical machine shown in (a) a radial cut, (b) a longitudinal cut of one pole segment and (c) a perspective of the total 48 assembled core and coil segments without the permanent magnet rotor.](image_url)
5.2.2.2 Flux densities

The flux densities in the different machine regions are shown in Figure 5.9 for the working point 128 Nm. The bottom right picture in Figure 5.9 shows that the intersection between the teeth and the core back is saturated. This region of the machine is a weak part of the magnetic design. It is possible to increase the area of this intersection by decreasing the slot depth or the slot width, but at the same time the available space for the copper winding is also decreased. If the number of poles is increased, the required areas for the intersection and the yoke itself are reduced (to the cost of a higher electrical frequency for the same speed) and the inner radius of the winding is increased, which means that the average conductor length is shortened. As a result, only machines with relatively high number of poles are practically feasible with the presented concept. The intersection area between the teeth and the yoke is slightly modified in the manufactured laboratory prototype compared with the FEM-geometry, as illustrated by comparing the CAD-drawing in Figure 5.8.a with Figure 5.9. The outer corners of the yoke have a smaller radius in the manufactured machine than in the FEM-model in order to increase this area. The average length of the copper coils is consequently slightly increased. Each tooth surface that faces the radial air gap is for simplification reasons drawn in a straight plane in the FEM-model, while the real machine has a curved surface. This influences the waveforms of the computed radial air gap flux densities by flattening the tops.

![Figure 5.9: The FEM-computed flux densities (in teslas) in one pole segment, which is axially divided in half, shown in different perspectives at the electromagnetic torque 128 Nm. Only 1/32-part of the machine is modelled due to symmetry.](image)

The FEM-computed no-load flux densities in the radial air gap and in the tooth that is most exposed to the permanent magnet flux are shown in Figure 5.10. It is clearly illustrated that the tooth is heavily saturated even at no-load. This condition does not change much when the machine is loaded though. What happens is that the tooth next to the middle tooth, which is unsaturated at no-load, gets saturated at higher load and takes over the hardest burden from the middle tooth. The saturation level in the machine is almost independent of the load, since
the armature reaction is so small in comparison to the permanent magnet excited flux density, mainly due to the high pole number. The flux densities in the radial air gap and in three teeth at 128 Nm are shown in Figure 5.11.

![Figure 5.10](image1.png)

**Figure 5.10:** The FEM-computed no-load flux density in (a) the tooth that is most exposed to the permanent magnet flux and (b) the middle of the radial air gap showing the radial component (x-direction for the global coordinate system).

![Figure 5.11](image2.png)

**Figure 5.11:** The FEM-computed flux density at 128 Nm in (a) one pole segment and (b) the middle of the radial air gap showing the radial component (x-direction for the global coordinate system).

The FEM-computed flux densities in the axial air gaps at 128 Nm are shown in Figure 5.12 at the radii 100, 92, 84 and 76 mm. The waveform of the flux density gets thinner with decreasing radius, due to the V-shape of the axially directed permanent magnets. The dips in the waveform are results of the slotting effect. Slot bridges of SMC can reduce this effect and at the same time reduce the cogging torque and the iron losses in the tooth tips.

Figure 5.13 defines the global coordinate system in polar cylindrical coordinates for the FEM-model. The computed flux densities in the teeth at no-load and at 128 Nm are shown in Figure 5.14 and Figure 5.15, respectively. Similarly, the flux densities in the yoke are shown in Figure 5.16 and Figure 5.17, respectively. The stepwise waveforms in the yoke is due to a relatively coarse mesh in this region. The radial and the axial components of the flux density in the teeth have the same order of magnitudes on this computational line. This region is among the most saturated parts of the machine, since fluxes from both the radially and axially directed permanent magnets are combined here before entering the yoke.
Figure 5.12: The FEM-computed axial component (z-direction for the global coordinate system in Figure 5.9) of the flux density in the axial air gaps at the working point 128 Nm at the radius (a) 100 mm, (b) 92 mm, (c) 84 mm and (d) 76 mm.

Figure 5.13: The definition of the global coordinate system in polar cylindrical coordinates.
Figure 5.14: FEM-computed flux density in the teeth at no-load (R=91 mm, $0^\circ \leq \theta \leq 45^\circ$, $L=20$ mm).

Figure 5.15: FEM-computed flux density in the teeth at the working point 127.5 Nm (R=91 mm, $0^\circ \leq \theta \leq 45^\circ$, $L=20$ mm).
Figure 5.16: FEM-computed no-load flux density in the yoke ($R=85.5\,\text{mm}$, $0^\circ \leq \theta \leq 45^\circ$, $L=0\,\text{mm}$).

Figure 5.17: FEM-computed flux density in the yoke at the working point 127.5 Nm ($R=85.5\,\text{mm}$, $0^\circ \leq \theta \leq 45^\circ$, $L=0\,\text{mm}$).

### 5.2.2.3 Torque

The maximum average electromagnetic torque of a non-salient radial flux synchronous PM machine can be expressed by

$$T_{\text{em}} = 2\pi R^2 L B_{\delta l} K_1 = R A_{\text{rad}} B_{\delta l} K_1,$$

(5.5)
where $R$ is the air gap radius, $L$ is the length of the rotor, $A_{rad}$ is the radial air gap area, $B_{\delta 1}$ is the fundamental air gap flux density and $K_1$ is the fundamental current loading (RMS-values). The maximum average electromagnetic torque of a non-salient single-sided axial flux machine is given by

$$T_{em} = \pi R_o \left( R_o^2 - R_i^2 \right) B_{\delta 1} K_{o1} = R_o A_{ax} B_{\delta 1} K_{o1}, \tag{5.6}$$

where $R_o$ and $R_i$ is the outer and inner radii, respectively, $A_{ax}$ is the axial air gap area and $K_{o1}$ is the fundamental current loading at the outer radius. The outer radius and the radius of the radial air gap are approximately the same for the presented SMC-machine ($R_o \approx R$). This means that the maximum average electromagnetic torque of the SMC-machine, with one radial air gap and two axial air gaps is given by

$$T_{em} = 2 \pi R \left[ R L + \left( R^2 - R_i^2 \right) B_{\delta 1} K_{1} = R \left[ A_{rad} + 2 A_{ax} \right] B_{\delta 1} K_{1}. \tag{5.7}$$

The torque production advantage of the SMC-machine compared with a radial or an axial flux machine is accordingly due to its larger air gap area. This statement cannot be given without a discussion, since some questionable assumptions are made. The magnetic and the electric loadings must be equal for the machines. This is an overestimation for the presented machine, since cores based on iron laminations have better magnetic properties, as described above. The electric loading of the SMC-machine will therefore also be punished, since more iron is required, leaving less space for copper. This is not completely true, because the segmented SMC-core enables a very high slot fill factor and a very high thermal conductivity between the copper and the core, which means that the SMC-machine may actually have almost the same electric loading as a conventionally designed machine. Undoubtedly, the high magnetic loading is a problem, since the iron losses are much higher in SMC-cores. The calculated line-to-line no-load voltage at 1000 rpm is 48 volts (at 120 °C). The phase resistance and inductance are both very low of 10 mΩ (at 120 °C) and 8,3 mH, respectively. The low inductance results in a power factor close to unity. This means that a relatively small inverter is required, which is cost saving for the drive system. The finite element method computed electromagnetic torque versus the current loading at the radial air gap is given in Figure 5.18. The electromagnetic FEM-computations are performed using Flux2D [116].

![Figure 5.18](image-url)
Different torque ripple reduction techniques are applied in the electrical machine design. The permanent magnet arc of the radial part is equal to two slot pitches. The permanent magnets for the radial part are furthermore divided in three and each magnet is stepwise skewed by 1/3 slot pitch. The permanent magnets for the axial parts have an arc length of 2.5 slot pitches at their outer radii, while they have an arc length of 1.5 slot pitches at their inner radii, i.e. they have a shape with a built-in continuous skew of one slot pitch.

### 5.2.2.4 Losses and efficiency

The teeth iron losses are calculated by

\[
P_{\text{teeth,SMC}} = \left( k_{h,SMC} f B_{m,\text{teeth}}^2 + k_{\text{eddy},SMC} f^2 B_{m,\text{teeth}}^2 \right) m_{\text{teeth}},
\]

and the yoke iron losses are calculated by

\[
P_{\text{yoke,SMC}} = \left( k_{h,SMC} f B_{m,yoke}^2 + k_{\text{eddy},SMC} f^2 B_{m,yoke}^2 \right) m_{\text{yoke}},
\]

where \( k_{h,SMC} \) (12,5 \( \cdot \) 10^{-2} Ws/T^2/kg) is the hysteresis coefficient and \( k_{\text{eddy},SMC} \) (2 \( \cdot \) 10^{-5} Ws^2/T^2/kg) is the eddy current coefficient of the Soft Magnetic Composites material, \( f \) is the electrical frequency, \( B_m \) is the maximum flux density and \( m \) is the mass. The iron coefficients are found by a curve-fitting approach of the material data from the manufacturer [115]. The mass of the teeth is 6.1 kg, while the mass of the yoke is 2.9 kg. The maximum flux density is assumed constant independently of load and sinusoidal variation is also assumed. This assumption is roughly valid, since the machine is not field-weakened. The maximum flux densities in the teeth and the yoke are by the use of finite element method computations found to be 1.5 teslas and 1.3 teslas, respectively, which is seen in Figure 5.15 and Figure 5.17. The iron losses in the machine are as a result of the assumptions described above a pure function of the speed and they are given in Figure 5.19.

![Figure 5.19: Calculated teeth, yoke and total iron losses.](image)

The total losses include the iron losses, winding losses, mechanical losses and stray losses, according to Paper 9, and are plotted in the torque-speed diagram in Figure 5.20.
Figure 5.20: Calculated total losses (in watts) plotted in a torque-speed diagram.

The machine efficiency is calculated as (generator mode)

$$\eta = \frac{\omega_m T_{\text{shaft}} - P_{\text{total}}}{\omega_m T_{\text{shaft}}} \cdot 100\%,$$

(5.10)

where $\omega_m$ is the mechanical angular speed and $T_{\text{shaft}}$ is the shaft torque. The efficiency map of the machine based on the calculated losses is given in Figure 5.21. The maximum calculated efficiency is more than 96%.

Figure 5.21: The calculated efficiency of the machine in generator operation.
5.2.3 Thermal analysis

The speed-changing machine is as already mentioned cooled by means of forced air convection. The thermal analysis of it is performed with the use of the finite element software FEMLAB [117]. Another approach is for instance to do a lumped-parameter thermal analysis [118], but that is not performed here. The analysis is done in two steps. In the first step the velocity distribution in the region where the forced air convection occurs is obtained by solving the Incompressible-Navier-Stokes equations as given by [119]-[120]

\[
\begin{align*}
\frac{\partial \mathbf{u}}{\partial t} + \mathbf{u} \cdot \nabla \mathbf{u} &= -\frac{1}{\rho} \nabla p + \nu \nabla^2 \mathbf{u}, \\
\nabla \cdot \mathbf{u} &= 0,
\end{align*}
\]

where \( \mathbf{u} \) is the fluid velocity vector, \( \nu \) is the kinematic viscosity, \( p \) is the pressure and \( \rho \) is the mass density, where the first line represents the conservation of momentum, while the second line represents the conservation of mass. In the second step the temperature distribution is obtained by applying the conservation of energy equation as given by [121]

\[
\rho c_p \frac{\partial T}{\partial t} + \nabla \cdot \left( -k \nabla T + \rho c_p T \mathbf{u} \right) = q_c,
\]

where \( c_p \) is the thermal capacity of the fluid at constant pressure and \( k \) is the thermal conductivity. The expression within the brackets is the heat flux vector and \( q_c \) is the heat source. The mechanical fluid flow is not assumed to affect the heat transfer, since the temperature gradient is insignificant. Only one slot pitch of the machine is modeled due to thermal symmetry and the analysed model is shown in Figure 5.22.

Figure 5.22: The geometrical model used in the thermal finite element method computations.

The results from the computations in Figure 5.23 are showing the temperature distributions in the model for different power losses of 442, 769 and 1060 watts, respectively. The losses are taken from no-load measurements as described in the next chapter, and the thermal coefficients used in the computations are calibrated against measurement results. The thermal contact between the coil and yoke is assumed to be perfect, which means that the loss distribution has small influence on the temperature rise. The core and the winding in the laboratory prototype is vacuum pressure impregnated (VPI), which ensures a high conductivity between these parts. The calculated average temperature in the core versus the
volume flow rate of forced air convection at a machine power loss of 1060 watts is shown in Figure 5.24. The value of the air velocity of 13 m/s on the inlet that corresponds to approximately 8.7 m³/min (with an inlet temperature of 25 °C), gave a good agreement between FEM-calculations and measured temperatures [122]. The calculated average temperatures of the core at steady-state are 69, 101 and 129 °C, respectively, for the three loss conditions. The temperature of the winding is almost the same as for the core, because of the good thermal coupling between the two parts.

Figure 5.23: FEM-computed temperature distribution in one tooth and coil segment for the power loss (a) 442 watts, (b) 769 watts and (c) 1060 watts (in kelvins).

Figure 5.24: FEM-calculated average temperature in the machine versus the volume flow rate of forced air at a machine power loss of 1060 W.
5.2.4 Sliprings

The slipring unit is a vital component in the Four Quadrant Transducer drive system and its typical performance has been investigated in Paper 4 and Paper 7. A principle sketch of a slipring unit with three brushes per phase is illustrated in Figure 5.25.

![Principle sketch of a slipring unit](image)

**Figure 5.25: Principle sketch of a slipring unit (one phase is shown).**

The amount of losses associated with brush operation is influenced by a lot of different parameters like current density, peripheral speed, contact pressure, temperature, humidity, brush material and more. The combined effect of peripheral speed and contact pressure influence the lifetime of brushes the most [123]. The total losses in a slipring unit, equipped with copper-graphite brushes with a slipring diameter of 100 mm, versus both the speed and the current density (RMS) are exemplified in Figure 5.26.

![Total losses in a slipring unit](image)

**Figure 5.26: Total losses in a slipring unit equipped with copper-graphite brushes with a slipring diameter of 100 mm.**
It is experienced that big vibrations are unsuitable for brush operation. In severe cases, the slipring unit can get overheated and destroyed due to added friction. Special attention must be given to the mechanical design of the slipring unit if it is supposed to be used in applications with potentially high vibration risks, for instance in a hybrid electric vehicle. A comprehensive study of the performance of brushes made of silver-graphite, copper-graphite and pure graphite is explained in [124].

5.3 Torque-changing machine

The torque-changing machine has been thoroughly investigated in Papers 3, 5 and 6. The description in this section is therefore focused on the most interesting results from those papers along with some clarifications.

5.3.1 Design philosophy

The torque-changing machine must be designed for field-weakening operation and for high overload capability below base speed. The axial build of the machine has to be short, since the available space is very limited. The choice therefore fell on a synchronous permanent magnet machine with fractional pitch concentrated windings. The teeth are made of grain-oriented silicon iron sheets, because it has a higher saturation induction and lower losses in the grain orientation, compared with non-oriented steel, which is used for the yoke. Rectangular shaped copper conductors allow for a high slot fill factor. An outer rotor structure is chosen, since it gives a longer air gap radius than an inner rotor structure, but mainly because it gives the best utilization of the chosen machine topology in the Four Quadrant Transducer.

5.3.2 Electromagnetic analysis

5.3.2.1 Geometrical parameters

The winding layout and flux distribution of the machine at 20 kW and 2000 rpm along with its calculated magnetomotive force are shown in Figure 5.27. The machine configuration is a so-called modular PM machine with coils of the same phase winding around consecutive teeth. The relatively high content of harmonics in the MMF introduces alternating magnetic fields in the rotor. The flux lines in Figure 5.27 also illustrate this. The stator inner and outer radii are 61 mm and 102 mm. The air gap length is 1,5 mm and the length of the permanent magnets is 4,5 mm. The thickness of the rotor back is 12 mm. A rectangular shape of the slot is used, because the chosen conductor shape is rectangular. It is not advantageous for an inner stator structure that the tooth width increases with increasing radius though. In fact the complete opposite is to be preferred, since the amount of flux is higher closer to the yoke, due to slot leakage. The slot shape was chosen in order to test the concept with rectangular conductors and to identify the fill factor. The slot shape is more idealistic for outer stator configurations with high overload requirements. The presented machine could perform just as good with a rounded slot shape, round conductors (lower fill factor) and less iron in the teeth, but then the practical limitations of the proposed concept would not have been discovered.
Figure 5.27: (a) Winding layout and flux distribution at 2000 rpm and 20 kW with peak current in phase A. (b) The fundamental and total three-phase MMF and its harmonics for one turn per tooth coil and per unit current.

5.3.2.2 Stator losses

The iron losses are calculated using the finite element method with time stepping, hence the actual flux density variation over one electrical period can be found. The iron losses per mass unit are then calculated by \[ \rho_{\text{iron}} = \frac{1}{T_e} \int_{t=0}^{T_e} \left[ k_h \frac{d^2 B}{dt^2} + \frac{k_e}{8,67} \left( \frac{dB}{dt} \right)^{3/2} \right] dt \]  

The material coefficients are found by a curve-fitting approach on the material data from the manufacturer of the iron sheets based on measurements according to the Epstein frame method [126]. The material used in the calculations with a sheet thickness of 0.35 mm is the M235-35A from Cogent. This material is even used in the teeth, despite the use of grain-oriented 0.30 mm sheets in the laboratory prototype, because otherwise the numerical problem becomes too complex. The hysteresis coefficient \( k_h \) is 0.0165 W/T^2/kg, the conductivity \( \sigma \) is \( 1.69 \times 10^6 \) \( \Omega^{-1} \) m\(^{-1} \), the mass density \( \rho \) is 7600 kg/m\(^3 \) and the excess coefficient \( k_e \) is 0.00034 W/T^{3/2}/s^{2/3}/kg.

The finite element computed iron losses of the torque-changing machine at no-load are given in Figure 5.28 and at 20 kW output power (2000-6000 rpm) in Figure 5.29. The iron losses that were presented in Paper 5 and Paper 6 were higher than those given in Figure 5.29, due to an error in the translation of the excess coefficient. The iron losses in both the teeth and yoke increase with increasing speed at no load, since the magnetic condition is constant while the frequency varies. The iron losses in the teeth also increase with increasing speed at 20 kW output power in the field-weakening region, because the flux gets more distorted and the eddy current component increases, even if the maximum flux density decreases. However, the flux density waveform in the yoke is not changing that much, and since the machine is field-weakened the losses in the yoke are lower in the constant power region than at base speed. The iron losses in field-weakening operation are further analysed in Paper 8.
The rectangular copper conductors of the torque-changing machine prototype have the dimension two times three millimeters. This dimension is not idealistic for the higher operating frequency of the machine, which is around 700 Hz, due to the high AC-resistance. It is anyway chosen in order to limit the number of parallel coils and consequently the number of connection points in the manufacturing. The height should preferably be reduced in a machine intended for series-production. The calculated AC-resistance coefficients of the winding for the torque-changing machine for various conductor heights are shown in Figure 5.30 versus the frequency. The copper slot fill factor is assumed constant, which is a slight overestimation for decreasing heights.

![Figure 5.28: FEM-computed iron losses at no-load generating operation.](image)

![Figure 5.29: FEM-computed iron losses at 20 kW (2000-6000 rpm) field-weakening motor operation.](image)

![Figure 5.30: The calculated AC-resistance coefficient versus frequency for the torque-changing machine for various height dimensions of the copper conductor.](image)
5.3.2.3 Rotor losses, radial forces and noise generation

The tangential flux density component in a fixed point between two permanent magnets in the rotor back versus the rotor position at 20 kW for various speeds is shown in Figure 5.31. The flux density varies between 0.4 teslas and 1.3 teslas at 2000 rpm and slightly less for higher speeds in the constant power region. The tangential flux density component in the rotor back is also illustrated by the three-dimensional graph in Figure 5.32, showing the combined influence of time and space angle at 20 kW and 2000 rpm. The alternating magnetic fields in the rotor may induce losses in the permanent magnets and the rotor iron. In order to minimize these losses, the permanent magnets should be manufactured in many small pieces and the rotor iron should be laminated. Unfortunately, the magnetic field variations in the rotor were underestimated at the design stage, which means that the manufactured laboratory prototype did not follow these design principles. Instead, the rotor iron is solid and as a consequence the machine cannot fulfill the specifications.

\[ \text{Figure 5.31: Flux density (tangential component) in rotor back at 20 kW and various speeds.} \]

\[ \text{Figure 5.32: Flux density (tangential component) in rotor back at 20 kW and 6000 rpm.} \]
The high and asymmetric MMF of this machine also creates unbalanced radial forces on the rotor and the stator. The radial magnetic stress acting on the rotor core of the torque-changing machine at base operating conditions is illustrated in Figure 5.33. The FEM-drawings illustrate four equidistant time samples during a rotation equivalent to half an electrical period. It is shown that the highest radial force component moves all around the periphery of the machine during this time interval. The dominant frequency of the radial force is therefore equal to twice the electrical frequency. At base speed this frequency is 467 Hz. It is obvious that the radial force component will create annoying noise, since it is located in a frequency range where the human ear has a high sensitivity. Furthermore, in this upper base frequency range, surrounding equipment and walls will easily pick up the vibration and transport it through for instance buildings, since this tone is hard to dampen out. The 14 poles and 15 slots machine combination is consequently not recommended for use in a high performance drive system. The same is true for other concentrated winding machines, which are not balanced, normally modular machines with an odd number of slots, and especially machines with single-layer windings.

![Figure 5.33: Radial magnetic stress acting on the rotor core of the torque-changing machine at base operating conditions. The FEM-drawings illustrate four equidistant time samples during a rotation equivalent to half an electrical period.](image)

5.3.3 Thermal analysis

An aluminium cylinder on the inside of the stator mainly cools the torque-changing machine. The cylinder consists of eight axial series-connected cooling-channels filled with water. The water, which is driven by a pump, transports the heat away from the machine to an external cooler. The machine losses are mainly produced in the stator windings and core and are carried away in the radial direction to the cylinder by means of conduction in the solid parts. The three-dimensional thermal problem is solved by the use of FEM (FEMLAB) in three steps. First the velocity field inside the cooling-channels is calculated using the Incompressable-Navier-Stokes equations (5.10). The obtained velocity field is coupled with the conservation of energy equation (5.11). The three-dimensional multi-physical model as illustrated in Figure 5.34.a is then used to calculate the average temperature on the outer surface of the aluminium cylinder, which is in physical contact with the inner surface of the stator core. The obtained temperature is in the third and last step used as a boundary temperature in the two-dimensional model in Figure 5.35 to calculate temperature rises within the machine with the use of the conduction heat equation [121]

\[
\frac{\partial^2 T}{\partial x^2} + \frac{\partial^2 T}{\partial y^2} + \frac{\partial^2 T}{\partial z^2} + \frac{q_\text{r}}{k} = \frac{1}{\zeta} \frac{\partial T}{\partial t}
\]  

(5.14)

where \(\zeta\) is the thermal diffusivity, as given by
\[ \zeta = \frac{k}{\rho c}, \]  

(5.15)

where \( c \) is the specific heat capacity and \( q_G^* \) is the rate of energy per unit volume produced inside of the volume. The right term in (5.14) can be ignored for a stationary problem. The temperature distribution for each cooling-channel, assuming no heat conduction between the channels, is illustrated in Figure 5.34.b, where \( L_{active} \) is the core length of the machine and \( \Delta T_{channel} \) is an average temperature rise in one cooling channel. \( \bar{T}_{surface} \) is the average surface temperature of the cylinder, when perfect mixing of water in the end regions where the water changes direction is assumed. The inlet temperature in the next channel is the inlet water temperature plus the temperature rise in the previous channel.

Figure 5.34: (a) FEM-model of 1/16-part of actual cooling-cylinder, showing the aluminium frame (1) and half of the duct (2). The arrows are illustrating the velocity profile of the fluid. (b) Average temperature on the outer surface of the cooling-cylinder and the temperature distribution in the eight cooling channels.

The machine parts and boundary conditions used in the thermal FEM-computations are shown in Figure 5.35. All the electrical machine parameters are based on calculated values. The thermal coefficients used in the calculations are calibrated against measurement results from tests on a laboratory prototype machine though, as described in Chapter 6. A more detailed description of the thermal calculations can be found in [122].

The temperature distributions for four different simulated working points are shown in Figure 5.36 for an inlet water temperature of 22 °C and a water flow of 2.5 dm³/min. The working points are 95 Nm at 1 rpm, 20 kW at 2000 rpm, 20 kW at 4000 rpm and 20 kW at 6000 rpm and are estimated to be close to thermally rated conditions. The lower iron losses described in this chapter, compared to the iron losses as given in Paper 5 and Paper 6, are compensated for in these computations. It is found that the conductor experiences the highest temperature for all working conditions and the highest conductor temperature is 144 °C at the base speed of 2000 rpm. The temperatures are lower in the field-weakening region, due to lower copper losses. It is now possible to estimate the rated performance of the machine, assuming that it was manufactured without having the high rotor losses and AC-resistance losses, as explained earlier. If a maximum temperature rise of 105 K is chosen (class F insulation limit), then the highest permissible torque at base speed is 87 Nm. The torque density in motoring operation, based on the active weight of the machine, is then 5.3 Nm/kg. The calculated results are summarized in Table XVIII.
Figure 5.35: The geometry model of one pole of the torque-changing machine as used in the thermal FEM-computations.

Figure 5.36: FEM-computed temperature distribution in the PM machine at motoring operation with the working points (a) 95 Nm/1 rpm, (b) 20 kW/2000 rpm, (c) 20 kW/4000 rpm and (d) 20 kW/6000 rpm.
### Table XVIII: Calculated losses and temperatures for different working points.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Unit</th>
<th>1</th>
<th>2000</th>
<th>2000</th>
<th>4000</th>
<th>6000</th>
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</thead>
<tbody>
<tr>
<td>Speed</td>
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<td>2000</td>
<td>2000</td>
<td>4000</td>
<td>6000</td>
</tr>
<tr>
<td>Torque</td>
<td>[Nm]</td>
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<td>95</td>
<td>87</td>
<td>48</td>
<td>32</td>
</tr>
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<td>Power</td>
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<td>0</td>
<td>20</td>
<td>18</td>
<td>20</td>
<td>20</td>
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<td>Copper losses</td>
<td>[W]</td>
<td>708</td>
<td>708</td>
<td>600</td>
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<td>Iron losses</td>
<td>[W]</td>
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<td>122</td>
<td>137</td>
<td>175</td>
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<td>Total losses</td>
<td>[W]</td>
<td>708</td>
<td>830</td>
<td>722</td>
<td>365</td>
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<td>22</td>
<td>22</td>
<td>22</td>
<td>22</td>
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<td>Temperature, Yoke</td>
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<td>94</td>
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<td>65</td>
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<tr>
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<td>127</td>
<td>73</td>
<td>82</td>
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<td>Temperature rise</td>
<td>[K]</td>
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<td>121</td>
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<td>60</td>
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</table>

### 5.4 Summary

Chapter 5 has described the Four Quadrant Transducer drivetrain and its two electrical machine designs intended for a mid-sized passenger car. The system takes advantage of the best characteristics of the series and the parallel hybrid systems. The engine is for instance not mechanically connected to the wheels and the electrical components, especially the costly battery, have lower power ratings than the engine itself. The gearbox can be of a simple single stage type, which reduces the mechanical complexity and makes the traction performance of the vehicle smooth, without gear changes and drops in power. On the other hand, by introducing an extra gear stage and a lock-up function between the engine and the outgoing shaft, the electrical machines can be made smaller and the consumption at highway driving decreases. Simulations on a complete hybrid system show that fuel savings of more than 40% compared to a conventional vehicle can be achieved at city-traffic driving. The savings are modest at highway driving, since the engine is required to operate at high power during such conditions, and the help from the electrical system is negligible.

The speed-changing electrical machine of the system has been designed with a soft magnetic composite core and both axial and radial air gap fluxes. The best benefit of this concept is that it delivers a very high torque for a limited volume envelope. It is designed for automation manufacturing with its segmented stator parts. The machine also has a typical high power factor, which saves cost on the required inverter. The concept is limited to machines with high pole numbers. The presented 16 pole machine cooled by forced air convection has a calculated torque density of 5.1 Nm/kg in motoring operation.

The torque-changing electrical machine of the system has been designed with grain-oriented and non-oriented silicon iron laminated cores in the teeth and the yoke, respectively. The windings are concentrated and are made of rectangular copper conductors. The whole concept enables cost effective manufacturing by a segmentation of the stator parts. The machine is designed for overload conditions and the weight of the permanent magnets has been minimized in order to further save cost. The axial machine build is low thanks to short end-windings. The pole and slot combination of the machine is an extraordinary 14 and 15. The choice of this combination was based on a high fundamental winding factor of 0.95 and a negligible ripple torque, which is also verified. However, some drawbacks have been
discovered and analysed. The presented combination has typically high alternating magnetic fields in the rotor at high torque operation, which makes the machine susceptible to rotor losses. The phenomenon is induced by the unsymmetrical magnetomotive forces, which are also responsible for magnetic noise radiation. The 14 poles and 15 slots combination is consequently not recommended for use in high performance applications. The practical aspects and the potential cost effective production methods that are proposed are highly recommendable though. Neglecting the stray losses, the water-cooled machine concept is estimated to deliver a torque density of 5.3 Nm/kg in motoring operation.
6 Laboratory prototype

6.1 Introduction

The choice of machine topologies has opened up for a number of calculated risks. Relatively novel constructional methods like Soft Magnetic Composites, Gramme-windings, concentrated windings, modular permanent magnet machines, rectangular copper conductors, grain-oriented iron teeth and more have created new opportunities in the manufacturing process, but at the same time challenged what is practically feasible for a single-unit laboratory prototype. The choices of some constructional details have often been deliberately made in order to force forward potential weaknesses with the machine designs and to identify limitations in the production. At the same time big efforts have been spent on designing potentially cost effective machines for a hypothetical high volume production. The ideas are for instance high material utilization, low material waste, simple coils for automated manufacturing, segmentation of the core and forced convective cooling systems.

6.2 Manufacturing process

6.2.1 Speed-changing machine

Some pictures illustrating the manufacturing process of the laboratory prototype are shown in Figure 6.1 and the assembly of the machine is shown in Figure 6.2. The bobbin-wound copper coils are almost completely filling up the slot space, especially along the slot width of 4.4 mm, which is filled up by a copper conductor width of 4.0 mm. Each slot consist of 11 conductors with a height of 1.06 mm. Leaving a 1 mm margin of air in the slot top, the slot fill factor is incredible 0.78, for a slot with a total height of 14 mm. The active parts of the machine consisting of permanent magnets, copper, SMC and steel (for PM rotor back) have a weight of 1.3, 3.1, 9.0 and 6.3 kg, respectively, and a total weight of 19.7 kg. The cogging torque is low due to the shape of the permanent magnets in the axial air gaps and the three-times stepwise skewing of the permanent magnets in the radial air gap.

Figure 6.1: The Soft Magnetic Composites core and the rectangular copper conductor coils.
6.2.2 Torque-changing machine

The coils of both the speed-changing and the torque-changing machines consist of bobbin-wound rectangular conductors manufactured under tension in order to achieve very high slot fill factors. The manufacturing process of a coil for the torque-changing machine is illustrated in Figure 6.3 as an example. (This process can be automated in high-volume production.)

The core of the torque-changing machine is built-up by grain-oriented teeth and a non-oriented core. Each tooth consists of sheets that are glued together forming an individual part. The coil, which is already insulated, is then placed around the inner part of the tooth before the inner region of the tooth is glued to the yoke, which is formed by sheets that also is glued together to the same length as the tooth, 65 mm. The principle is shown in Figure 6.4.
The coil, as shown in Figure 6.3 and Figure 6.4, is actually made of two copper conductors in parallel, each with a dimension of two times three millimeters. The coil has ten turns. The active parts of the torque-changing machine consisting of permanent magnets, copper, silicon iron laminations and magnetic steel (for PM rotor back) have a weight of 0.9, 3.3, 8.0 and 4.3 kg, respectively, and a total weight of 16.5 kg. The permanent magnets are glued on the inside of a can of magnetic steel. No bandage is therefore required, which is normally used in inner rotor designs in order to keep the magnets on the rotor surface for increasing speed and centrifugal forces. The inner surface of the stator yoke is mounted on an aluminium cylinder that is water-cooled. This cylinder is in connection with the housing through an end-shield, which also is water-cooled. The assembly of the torque-changing machine is shown in Figure 6.5.

![Figure 6.5: The assembly of the torque-changing machine.](image1)

The main components of both machines that make the Four Quadrant Transducer laboratory prototype are illustrated in Figure 6.6. The speed-changing machine with the sliprings on the same shaft is seen to the left, the permanent magnet rotors are in the middle and the torque-changing machine is seen to the right. A selection of various smaller components is seen in the front, for instance a resolver (position sensor), bearings and copper-graphite brushes.

![Figure 6.6: The main components of the Four Quadrant Transducer laboratory prototype.](image2)
6.3 Measurements

The measurement results as described in this section are mainly based on Paper 6 and Paper 9.

6.3.1 Speed-changing machine

6.3.1.1 Measurement set-up

The measurement set-up is shown in Figure 6.7. A torque transducer measures the mechanical power and a power meter coupled to an oscilloscope measures the voltages. A fan of 185 watts provides the forced air-convection. The air inlets to the Four Quadrant Transducer housing consist of four holes that are connected to the fan through tubes with a diameter of 50 millimeters. The air outlets also consist of four holes that are in direct contact with the ambient, as shown in Figure 6.7. The water-cooling system is inactive during the tests of the speed-changing machine. An induction machine with variable speed drives the speed-changing machine in the generator tests.

Figure 6.7: Measurement set-up for the laboratory prototype, which is partly cooled by forced air from a fan and partly by a built-in water-cooling system (mainly for the torque-changing machine).

6.3.1.2 Winding fault

Unfortunately, the winding is manufactured with a fault in one of the phases (discovered after the VPI-process), which creates circulating currents and extra losses. This means that the machine is only tested at no-load generating condition, in order to avoid further damage. Two alternative explanations of the fault are most likely. In the first alternative, one coil in a phase is wrongly connected (switch of plus and minus terminals). In the second alternative, a short-circuit between turns in the same phase is present. A short-circuit between one phase winding and a non-magnetic steel end-plate is verified. However, this phase-to-ground short-circuit is insulated from the shaft, which means that the fault is locally limited, but this could cause the second alternative fault. The electromagnetic performance of the machine is therefore not verified. However, the thermal performance of the machine is verified. The core, the coils and the couplings to the shaft are partly glued and partly vacuum pressure impregnated. It is therefore extremely difficult to repair the faulty phase winding without
damaging the winding even more. Furthermore, it would be very costly and time-consuming to make an attempt to save the winding. These factors have led to the decision of not repairing the machine within the scope of this thesis work.

6.3.1.3 Measurement results

The winding fault is illustrated by the measured unsymmetrical no-load phase back-EMFs in Figure 6.8. The phase displacement between the peak voltages of the two phases with the highest amplitudes (A and C) differs in time in an irregular way. The phase displacement alternates between approximately 177 and 183 degrees, instead of the regular 120 and 240 degrees for a three-phase winding. None of the three phase waveforms are identical, even if two are more alike than the third. The waveforms have additionally a bended shape, which is characteristically for a loaded machine. This indicates a braking torque that is a result of circulating currents in the phase windings.

![Figure 6.8: The measured no-load phase back-EMFs at a speed of approximately 1000 rpm. (The winding is manufactured with a fault in one of the phases, which creates circulating currents and extra losses.)](image)

The measurement results from the no-load tests are given in Table XIX. The temperature rise is here defined as the difference between the core (SMC) and the inlet air temperatures. The temperatures are measured using resistance temperature detectors or RTDs (PT-100). The good thermal contact between the core and the winding makes the temperature rise in the different parts almost independent of the loss distribution, as explained earlier. This means that the maximum power loss for a chosen maximum temperature rise can be defined. The maximum power loss is found to be 1060 watts for a temperature rise of 105 kelvins. It is now possible to estimate the rated performance of the machine based on the combined measured and calculated data. The calculated total losses are plotted in the torque-speed map in Figure 5.20. The machine can operate at higher loss conditions for a limited amount of time. This is most likely sufficient, since high power operation is only required during accelerations of the driven vehicle. The winding of the machine is furthermore capable of withstanding heat shock treatments of up to 220 °C without insulation damage. The permanent magnets are not heated up to these high temperatures, since the transient overload interval is relatively short.
Table XIX: Measurement results from no-load tests.

<table>
<thead>
<tr>
<th>Speed [rpm]</th>
<th>1005</th>
<th>1499</th>
<th>1903</th>
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</thead>
<tbody>
<tr>
<td>Torque [Nm]</td>
<td>5.1</td>
<td>5.9</td>
<td>6.4</td>
</tr>
<tr>
<td>Power losses [W]</td>
<td>442</td>
<td>769</td>
<td>1060</td>
</tr>
<tr>
<td>Phase voltage, A, RMS [V]</td>
<td>22.5</td>
<td>33.8</td>
<td>43.1</td>
</tr>
<tr>
<td>Phase voltage, B, RMS [V]</td>
<td>10.1</td>
<td>16.3</td>
<td>21.9</td>
</tr>
<tr>
<td>Phase voltage, C, RMS [V]</td>
<td>22.8</td>
<td>34</td>
<td>41.5</td>
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</table>

<table>
<thead>
<tr>
<th>Temperatures</th>
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<td>Permanent magnets [°C]</td>
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<tr>
<td>Soft Magnetic Composites [°C]</td>
</tr>
<tr>
<td>Copper conductors [°C]</td>
</tr>
<tr>
<td>Ambient, air inlet [°C]</td>
</tr>
<tr>
<td>Air outlet [°C]</td>
</tr>
<tr>
<td>Temperature rise [K]</td>
</tr>
</tbody>
</table>

The rated speed is assumed to be 2000 rpm, where the rated torque is 106 Nm. This gives a torque density in generating operation of 5.4 Nm/kg (5.1 Nm/kg in motoring operation), which is in class with the water-cooled torque-changing machine with concentrated windings and silicon iron laminated core [Paper 6]. This figure can be compared with air-cooled induction motors (internal fan) and servomotors (natural convection) of typically 1.2 and 1.8 Nm/kg, respectively [127]. Earlier manufactured naturally cooled motors with iron powder cores have torque densities of 1.2 and 2.7 Nm/kg, respectively, where the former is a combined axial-radial flux rare-earth PM motor with air gap winding and the later is a radial flux ferrite PM motor with teeth, both designed for an electric bicycle application [127].

6.3.2 Torque-changing machine

The torque-changing machine is experiencing an alternating magnetic field in the rotor when loaded, as described earlier. Since the rotor back is manufactured as a solid iron can, the rotor has losses when loaded. It is therefore difficult to separate the measured total losses into loss components at high power conditions, which also risk demagnetization of the permanent magnets. The thermal performance of the machine is therefore evaluated in two steps. In the first step the iron losses and their influence on the heat generation is characterized by no-load and partial load tests in generating operation. In the second step the copper losses are in a similar way evaluated by load tests at low speed in motoring operation. The low speed ensures that the rotor losses are negligible.

6.3.2.1 Generating operation

The laboratory set-up and the measurement results for generating operation are shown in Figure 6.9 and Table XX, respectively. The machine is water-cooled with a flow of roughly 3 liters per minute. The losses are found by the direct method, which means that the torque, speed and electrical power are measured. The current is not controlled in the generator tests. It is only a function of the induced EMFs and the impedances in the electric circuits. Resistors are used as load. The measured results for the generating working point 31.5 Nm at 2000 rpm, as shown in Table XX, verify that the rotor has losses at high frequencies. The rotor temperature is almost 20 °C higher than for the stator core.
6.3.2.1.1 No-load back-EMF

The measured no-load back-EMF at 2000 rpm is shown in Figure 6.10. It is quite sinusoidal due to the special combination of 14 poles and 15 slots. However, it is not exactly the same waveforms in the measured curve and the FEM-computed curve (As seen in Paper 6). The difference is probably due to the grain-oriented iron material in the teeth of the real machine. The teeth were modeled with a non-oriented material in the calculations in order to simplify the numerical complexity of the problem.

Table XX: Test data from generating operation

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Unit</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Torque</td>
<td>[Nm]</td>
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<tr>
<td>Speed</td>
<td>[rpm]</td>
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<td>Mechanical power</td>
<td>[W]</td>
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<td>Phase voltage (RMS)</td>
<td>[V]</td>
<td>132,4</td>
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<td>Phase current (RMS)</td>
<td>[A]</td>
<td>0</td>
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<tr>
<td>Electrical power</td>
<td>[W]</td>
<td>0</td>
</tr>
<tr>
<td>Losses</td>
<td>[W]</td>
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</tr>
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<td>2,5</td>
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<td>Temperatures</td>
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<tr>
<td>End-winding</td>
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</tr>
<tr>
<td>Tooth</td>
<td>[°C]</td>
<td>36</td>
</tr>
<tr>
<td>Yoke</td>
<td>[°C]</td>
<td>35</td>
</tr>
<tr>
<td>Rotor iron</td>
<td>[°C]</td>
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</tr>
<tr>
<td>Cooling cylinder</td>
<td>[°C]</td>
<td>15</td>
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<tr>
<td>Inlet water</td>
<td>[°C]</td>
<td>9</td>
</tr>
<tr>
<td>Temperature rise</td>
<td>[K]</td>
<td>26</td>
</tr>
</tbody>
</table>
6.3.2.1.2 Resistance

The measured phase DC-resistance of 20.5 mΩ at room temperature is 29% higher than the analytically calculated 15.9 mΩ. The reason for the discrepancy is due to tolerances in the wire cross-section and slightly different resistivity of 17.24 nΩm compared to the analytically used 16.70 nΩm (5%-units of the increase) and relatively long connection wires and terminal cables (24%-units of the increase).

6.3.2.1.3 Inductance

The unsaturated phase inductance is measured 5.4 mH at 100 Hz. The FEM-computed phase inductance is 5.2 mH, where the end-winding inductance is disregarded. The end-winding inductance is therefore very small, less than 4% of the total inductance. The inductance is measured by using a sinusoidal power generator (ELGAR 1751B) with a frequency capability in the range 50-10000 hertz. The tested machine is connected as illustrated in Figure 6.11 and its rotor is locked in fixed positions (preferably in the d- and q-directions). A power meter measures the voltages and currents. The q-inductance is found from the following expression

$$ L_q = \frac{S_q}{2\pi f I_q^2} \frac{\sqrt{1-(\cos \phi)^2}}{3/2}, $$

(6.1)

where $S$ is the apparent power, $f$ is the electrical frequency, $I$ is the current (RMS), $\phi$ is the angle between the voltage and the current and the notation $q$ means that the measurements are performed with the rotor locked in the q-direction. The factor $3/2$ is due to the connection of the phase windings, as seen in Figure 6.11. The d-inductance is found in a similar manner. The measured phase inductance (average inductance between the d- and q-directions, since it is almost non-salient) versus frequency for the torque-changing machine is given in Figure 6.12.
6.3.2.1.4 Losses induced by current harmonics

The high frequency resistance of the machine can also be measured using the connection diagram in Figure 6.11. The resistance in the $q$-direction is found from the following expression (same procedure as for the inductance measurements)

$$R_q = \frac{S_q}{I_q} \cdot \frac{(\cos \phi)_q}{3/2}.$$  \hspace{1cm} (6.2)

The measured phase resistance versus frequency for the torque-changing machine is given in Figure 6.13. At very low frequency the resistance is very close to the measured DC-resistance of the winding, which confirms the reliability of the method.
It is important to note that the resistance in Figure 6.13 does not only represent the winding losses, but also the core losses and other stray losses like for instance in the permanent magnets and the housing. This resistance can be used for calculating losses induced by current harmonics. The $P_{R}$-losses can be found when the current spectrum is known, by multiplying the current squared with the resistance at that particular frequency. The average losses for the \( n \)th current harmonic component with a frequency of \( f_n \) are approximately given by

\[
P_n = 3 \cdot \frac{R_n(f_n) + R_q(f_n)}{2} \cdot I_n^2.
\] (6.3)

This method is very accurate when the studied frequency is much higher than the fundamental frequency driving the machine, because then the rotor is almost at standstill compared to the fluxes induced by the high frequency currents, exactly like in the proposed measurement method.

### 6.3.2.2 Motoring operation

The test bench for the motor tests is shown in Figure 6.14. An induction machine with variable speed loads the torque-changing machine. The shaft torque is measured by a torque transducer, which also measures the speed. The test results are given in Table XXI. The measurements in motoring operation are carried out at low speed in order to make the described rotor losses negligible. The low rotor back temperature verifies this assumption, which is around 20 °C lower than the stator core and around 70 °C lower than the winding temperature at a torque of 81.7 Nm (at thermal steady-state). The low rotor temperature is also of great advantage for the permanent magnets, which have high margins against demagnetization, and is consequently promoting the concept. The temperature rise is defined as the difference between the end-winding temperature and the inlet water temperature. The measured efficiency is in the range of 76-95 % for the operating points in Table XXI, which are nice figures considering that these are low power conditions.
Figure 6.14: Measurement set-up in motoring operation.

Table XXI: Test results from motoring operation.

<table>
<thead>
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<th>Value</th>
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<td>Speed</td>
<td>[rpm]</td>
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<td>Fund. phase voltage (RMS)</td>
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<td>Losses</td>
<td>[W]</td>
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<tr>
<td>Water flow</td>
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</table>

Temperatures

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<th></th>
<th>[°C]</th>
<th>Value</th>
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<td>Tooth</td>
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<td>52</td>
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<td>Yoke</td>
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<td>52</td>
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<tr>
<td>Rotor iron</td>
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<td>49</td>
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<td>Cooling cylinder</td>
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<tr>
<td>Inlet water</td>
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<td>20</td>
</tr>
<tr>
<td>Temperature rise</td>
<td>[K]</td>
<td>9</td>
</tr>
</tbody>
</table>

6.4 Summary

The laboratory prototype has shown that it is possible to manufacture high performance electrical machines with high material utilization and potential for automated production. The described concepts offer cost effective solutions for future drive systems in automotive and industrial applications. A number of weaknesses with the presented constructions have also been characterized, which should serve as guidelines for creating more optimized machines. A winding fault made it impossible to verify the electromagnetic performance of the SMC machine in practice.
7 Conclusions and future work

7.1 Conclusions

This thesis has focused on the design and analysis of permanent magnet machines for a novel hybrid electric vehicle drive system called the Four Quadrant Transducer. A number of electrical machine aspects have been identified, including cores of soft magnetic composites, fractional pitch concentrated windings, core segmentation, novel machine topologies and cost effective production methods. The main objective has been to analyse and judge the many unconventional machine aspects of which some may have the potential to improve the performance and reduce the cost of permanent magnet machines. Another objective has been to study the effects of the use of fossil fuels and describe them with a new perspective and thereby make one small contribution to the debate about energy issues. The novel Four Quadrant Transducer drive system has been the platform of which the major work has been based.

Much focus has been spent on the theory of concentrated windings for permanent magnet machines. The potential parasitic effects and methods to improve the torque performance have been described. Other topics that have been given a high priority are material and power loss studies. An important contribution to the understanding of iron losses during field-weakening operation has been made. A comprehensive use of finite element modeling has been done in the analysis combined with measurements on several laboratory prototypes.

The Four Quadrant Transducer drivetrain and its two electrical machines intended for a mid-sized passenger car has been studied. The gearbox can be of a simple single stage type, which reduces the mechanical complexity and makes the traction performance of the vehicle smooth, without gear changes and drops in power. Simulations on a complete hybrid system show that fuel savings of more than 40% compared to a conventional vehicle can be achieved at city-traffic driving. The savings are modest at highway driving, since the engine is required to operate at high power during such conditions, and the help from the electrical system is negligible. The laboratory prototypes have shown that it is possible to manufacture high performance electrical machines with high material utilization and potential for automated production. The described concepts offer cost effective solutions for future drive systems in automotive and industrial applications. A number of weaknesses with the presented constructions have also been characterized, which should serve as guidelines for creating more optimized machines. The research achievements should not be limited for the Four Quadrant Transducer system alone, but some of the findings should also advantageously be introduced in other drive applications with equally high demands on performance.

7.2 Future work

The laboratory prototype of the machine with soft magnetic composites core was manufactured with a winding fault. The magnetic performance of the core material could
consequently not be verified. It should be of greatest interest to repair the winding in a near future, test the machine, and make a final evaluation of the machine concept. An implementation of the Four Quadrant Transducer drive system in a real prototype vehicle should be very valuable. Practical tests under real driving conditions and environmental interactions are invaluable in order to optimize the total system.
References


References


References


References


References


References


[116] Flux2D, Finite element software, Version 7.60/6, Cedrat, France.


List of symbols

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<th>Symbol</th>
<th>Description</th>
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List of symbols

- **S** Apparent power [VA]
- **T** Temperature [°C]
- **T** Torque [Nm]
- **T_{em}** Electromagnetic torque [Nm]
- **T_{ICE}** Torque from the engine [Nm]
- **T_{load}** Load torque [Nm]
- **T_{max}** Maximum torque [Nm]
- **T_{p}** Time period [s]
- **T_{Shaft}** Shaft torque [Nm]
- **\overline{T}_{surface}** Average surface temperature [°C]
- **T_{ST}** Torque of the stator machine (torque-changing) [Nm]
- **T_{wheels}** Torque at the wheels [Nm]
- **t** Time [s]
- **T_{C(B_r)}** Temperature coefficient of remanence [%]
- **T_{C(H_{c,d})}** Temperature coefficient of coercivity of polarization [%]
- **U** Phase voltage [V]
- **U** Gear ratio [-]
- **U_{drivetrain}** Gear ratio of drivetrain [-]
- **U_{final drive}** Gear ratio of final drive [-]
- **U_{max}** Maximum line-line voltage, RMS [V]
- **u** Fluid velocity vector [m/s]
- **W** Total magnetic energy [J]
- **x** Distance in x-direction [m]
- **\dot{x}** Velocity in x-direction [m/s]
- **\ddot{x}** Acceleration in x-direction [m/s²]
- **y** Distance in y-direction [m]
- **z** Direction in z-direction [m]
- **z** Number of wires enclosed by p:th conductor layer [-]
- **z_{total}** Total number of wires in round conductor [-]
- **\Delta T_{rise}** Temperature rise [K]
- **\Delta T_{channel}** Average temperature rise in one cooling channel [K]
- **α** Inversed penetration depth [1/m]
- **δ** Load angle [°]
- **ϕ** Power angle [°]
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<thead>
<tr>
<th>Symbol</th>
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<td>$\rho_{air}$</td>
<td>Mass density of air</td>
<td>[kg/m$^3$]</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>Conductivity</td>
<td>[1/(Ωm)]</td>
</tr>
<tr>
<td>$\tau_m$</td>
<td>Permanent magnet pitch</td>
<td>[m]</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Electric angular speed</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Angular speed (mechanical)</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{DRM}$</td>
<td>Angular speed of double rotor machine (speed-chang.)</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{ICE}$</td>
<td>Angular speed of engine</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{load}$</td>
<td>Angular speed of load</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_m$</td>
<td>Mechanical angular speed</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{p_{\text{max}}}$</td>
<td>Angular speed at maximum power</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{ST}$</td>
<td>Angular speed of stator machine (torque-changing)</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{T_{\text{max}}}$</td>
<td>Angular speed at maximum torque</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\omega_{wheels}$</td>
<td>Angular speed of wheels</td>
<td>[rad/s]</td>
</tr>
<tr>
<td>$\xi$</td>
<td>Reduced conductor height factor</td>
<td>[-]</td>
</tr>
<tr>
<td>$\Psi$</td>
<td>Flux linkage</td>
<td>[Vs]</td>
</tr>
<tr>
<td>$\Psi$</td>
<td>Internal proximity effect factor</td>
<td>[-]</td>
</tr>
<tr>
<td>$\Psi_m$</td>
<td>Magnet flux linkage</td>
<td>[Vs]</td>
</tr>
<tr>
<td>$\Psi_{air}$</td>
<td>Armature flux linkage</td>
<td>[Vs]</td>
</tr>
</tbody>
</table>
Appendix A – Driving cycle diagrams

Figure 7.1 The operating window of the ICE (bottom left), the DRM (top right) and the ST (bottom right) for the European driving cycle, EC.

Figure 7.2 The operating window of the ICE (bottom left), the DRM (top right) and the ST (bottom right) for the Japanese driving cycle, Japan 10-15.
Figure 7.3 The operating window of the ICE (bottom left), the DRM (top right) and the ST (bottom right) for the American driving cycle, FTP75.

Figure 7.4 The operating window of the ICE (bottom left), the DRM (top right) and the ST (bottom right) for the European highway driving cycle, INRETS Auto 2.
Appendix B – 4QT Drawings

Figure 7.5: Cross-sectional drawing of the speed-changing machine of the 4QT.

Figure 7.6: Cross-sectional drawing of the torque-changing machine of the 4QT.
Figure 7.7: Longitudinal cross-sectional drawing of the Four Quadrant Transducer.
Figure 7.8: Drawing of one SMC-segment for the speed-changing machine.
Figure 7.9: Drawing of the radial solid core back of the permanent magnet rotor for the speed-changing machine.
Figure 7.10: Drawing of one of the axial core backs of the permanent magnet rotor for the speed-changing machine.
Figure 7.11: Drawing of one of the axial core backs of the permanent magnet rotor for the speed-changing machine that is coupled to the outgoing shaft and also functions as a fan.
Figure 7.12: Drawing of the stator core of the torque-changing machine.
Figure 7.13: Drawing of the solid core back of the permanent magnet rotor for the torque-changing machine.
Electromagnetic Transducer for Hybrid Electric Vehicles

Published in the Proceedings of the Nordic Workshop on Power and Industrial Electronics (NORPIE), Stockholm, Sweden, 12-14 August 2002.
Abstract—This paper describes a novel electromagnetic transducer called the Four Quadrant Transducer (4QT) for hybrid electric vehicles. The system consists of one electrical machine unit (including two rotors) and two inverters, which enable the vehicle’s Internal Combustion Engine (ICE) to run at its optimum working points regarding efficiency, almost independently of the changing load requirements at the wheels. In other words the ICE is operated at high torque and low speed as much as possible. As a consequence, reduced fuel consumption will be achieved.

The basic structure of the Four Quadrant Transducer system, simulation results and ideas about suitable topologies for designing a compact machine unit are reported. The simulated system of a passenger car is equipped with a single step gearbox making it simple and cost effective. Since the engine is not mechanically connected to the wheels and the electrical components have lower power ratings than the engine itself, the system takes advantage of the best characteristics of the series- and the parallel hybrid, respectively. The proposed concept looks promising and fuel savings of more than 40% compared with a conventional vehicle can be achieved.

Index Terms—Hybrid Electric Vehicles, Transducer

I. INTRODUCTION

ELECTRICAL machine drive systems for Hybrid Electric Vehicles (HEV) have been designed and studied in many academic- and industrial research programs during the last decade. The research interest within this area is still rapidly growing due to public awareness about environmental issues in general, but in particular due to the continuously tougher legislations on fuel consumption and emissions made by governments around the world. Future legislations will force car manufacturers to develop more advanced drivelines, where the hybrid topology is one alternative, which seems to fulfill the requirements without sacrificing performance or existing production facilities. The commercially available HEVs (e.g. Toyota Prius and Honda Insight) already demonstrate that the HEV concept is a realistic alternative to today’s conventional cars.

A novel electromagnetic transducer [1] called the Four Quadrant Transducer for hybrid electric vehicles will be presented in this paper. The concept is developed from the work on the Integrated Energy Transducer [2].

II. HYBRID ELECTRIC SYSTEMS

A. The series- and parallel hybrid systems

The driveline of hybrids consists of both an internal combustion engine and an electrical machine drive system. Two main topologies are the series and parallel hybrids, which are shown in figures 1 and 2, respectively. In order to explain the benefits of the Four Quadrant Transducer it is preferable to start by describing the characteristics of these two traditional hybrids.

The main advantage of the series hybrid is that the ICE can run at a certain torque and speed independently of the load requirement, since it is not mechanically connected to the wheels. This means that the ICE can be operated at its most efficient working point, especially at low loads, so that the fuel consumption and the emissions will be low. The gearbox can be made simple by using only one or two steps. One disadvantage is that the power ratings of all the components need to be of the same order of magnitude, since they are connected in series, which results in a big and expensive system. Another disadvantage is that the power is converted several times from the ICE to the wheels, leading to extra losses and higher fuel consumption at highway driving, when high power is required [2].

The advantage of the parallel hybrid is that the electrical components are fewer and can have lower power ratings than the ICE, because the power is not converted as in the series system, but added in parallel. The ICE can operate on its own at highway driving for better system efficiency. The disadvantage is, since the ICE is mechanically connected to the wheels, the ICE speed cannot be chosen freely. However, the ICE torque can be independently chosen by support from the electrical machine. This sometimes means higher fuel consumption and definitely higher emissions compared to the series hybrid at city traffic driving. In some parallel systems it is possible to disconnect the ICE using a clutch. The vehicle can thus be run entirely electric, but this will require a high-energy battery to acquire a reasonable driving range, which

Fig. 1. Principle sketch illustrating the series hybrid system.

The advantage of the parallel hybrid is that the electrical components are fewer and can have lower power ratings than the ICE, because the power is not converted as in the series system, but added in parallel. The ICE can operate on its own at highway driving for better system efficiency. The disadvantage is, since the ICE is mechanically connected to the wheels, the ICE speed cannot be chosen freely. However, the ICE torque can be independently chosen by support from the electrical machine. This sometimes means higher fuel consumption and definitely higher emissions compared to the series hybrid at city traffic driving. In some parallel systems it is possible to disconnect the ICE using a clutch. The vehicle can thus be run entirely electric, but this will require a high-energy battery to acquire a reasonable driving range, which
will add to the cost and weight of the vehicle. The gearbox is often conventional with several steps, due to the mechanical connection with the ICE [3].

Fig. 2. Principle sketch illustrating the parallel hybrid system.

B. The Four Quadrant Transducer hybrid system

The main characteristics of the 4QT hybrid system are based on the advantages from the series- and the parallel hybrid systems, respectively. The main characteristics are:

- The ICE is not mechanically connected to the wheels, which allows the ICE to run at a speed independent of the speed of the vehicle.
- The torque from the ICE is supported from the electrical machine unit, which means that the ICE can operate at torque values independent of the torque requirements at the wheels.
- The power ratings of the electrical machine unit and inverters do not need to be equal to the power of the ICE.
- The rated battery power can be chosen very low, since the electrical machine unit can operate as motor and generator simultaneously.
- The gearbox can be of a simple single step type, since only the electrical machine is mechanically connected to the wheels.

The basic layout of the 4QT hybrid system is shown in figure 3.

Fig. 3. Principle sketch illustrating the Four Quadrant Transducer hybrid system.

The electrical machine unit of the 4QT consists of two parts. One part is called the Double Rotor Machine (DRM), which is equipped with windings via slip rings. It is mechanically connected to the shaft of the ICE. The second part is called the Stator (ST), which is also equipped with windings. The DRM and the ST have one rotor in common, which is mechanically connected to the wheels via a gearbox. This rotor is preferably of permanent magnet or reluctance type.

The mission of the DRM is to make the speed of the ICE independent from the speed requirement at the wheels. This is possible by applying an AC current to the windings of the DRM. The DRM must be designed for the maximum torque of the ICE, since the torque of the DRM equals the torque of the ICE at steady state. The mission of the ST is to add or subtract the torque delivered by the ICE. The principle of operation is illustrated by the power flow diagram in figure 4.

Fig. 4. Power flow diagram of the Four Quadrant Transducer hybrid system.

The power delivered to the wheels of the vehicle is given by (the component losses are neglected for convenience)

\[ P_{\text{wheels}} = P_{\text{ICE}} + P_{\text{Battery}} = P_{\text{ICE}} + P_{\text{DRM}} + P_{\text{ST}}. \]  

(1)

The torque delivered to the wheels is given by

\[ T_{\text{wheels}} = G \cdot T_{\text{load}} = G \left[ T_{\text{ICE}} + T_{\text{ST}} \right] \]  

(2)

where \( G \) is the gear ratio. The angular speed of the wheels is given by

\[ \omega_{\text{wheels}} = \frac{1}{G} \omega_{\text{load}} = \frac{1}{G} \omega_{\text{ST}} = \frac{1}{G} \left[ \omega_{\text{ICE}} + \omega_{\text{DRM}} \right]. \]  

(3)

The wheel power can also be expressed

\[ P_{\text{wheels}} = \omega_{\text{wheels}} T_{\text{wheels}} = \left[ \omega_{\text{ICE}} + \omega_{\text{DRM}} \right] \left[ T_{\text{ICE}} + T_{\text{ST}} \right]. \]  

(4)

This expression shows that it is possible to operate the ICE at a fixed working point even if the load is changing simply by adjusting the speed difference between the two shafts of the 4QT and the torque of the ST.

Every ICE has an optimal operation line (o.o.l.) as a function of engine speed regarding fuel consumption. Figure 5 illustrates how the 4QT enables the ICE to operate on its
optimal operation line for two different load conditions with equal power requirements. From the working point of the ICE it is possible to move into any four quadrants by assistance of the hybrid system, whatever the load requires, which function has also given name to the 4QT.

![Diagram illustrating the flexibility of the 4QT](image)

**Fig. 5.** A diagram illustrating the flexibility of the 4QT in keeping the speed and the torque of the ICE on its optimal operation line independently of the changing load requirements.

The 4QT can be categorized as a power split hybrid system, due to its characteristic simultaneous motoring and generating capabilities. The previously presented power split systems [4-6] are strongly dependent on complex mechanical transmissions like a planetary gearbox or a Continuously Variable Transmission (CVT). The 4QT is unique in the sense that no such complex mechanical transmission is needed, but still the advantages of the power split system are exploited.

### III. Simulations

A complete hybrid electric vehicle system is simulated in Matlab/Simulink. The basic data are shown in table 1. The simulated vehicle is a passenger car and the gearbox is a simple single speed.

<table>
<thead>
<tr>
<th>Table 1. The Basic Data of the Hybrid Electric Vehicle Used in the Simulations.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Engine power</td>
</tr>
<tr>
<td>Engine volume</td>
</tr>
<tr>
<td>Mass of vehicle</td>
</tr>
<tr>
<td>Gearbox</td>
</tr>
<tr>
<td>Acceleration</td>
</tr>
<tr>
<td>Top speed</td>
</tr>
</tbody>
</table>

A gear ratio of 4,0:1 is chosen to achieve a low electrical machine power at maximum vehicle speed. A mechanical lock-up device between the ICE shaft and the outgoing 4QT shaft is possible. This would remove most of the losses from the electrical components at highway driving. At highway driving, high continuous power from the ICE is needed and the 4QT system will therefore not be able to reduce the consumption. A lock-up device is not used in the simulations, however. A more advanced multi speed gearbox can also be used to reduce the size of the 4QT machines. Only a single speed gearbox is used in the simulations to make it as simple and cost effective as possible.

Three driving cycles according to table 2 are simulated.

<table>
<thead>
<tr>
<th>Table 2. Basic Data of the Simulated Driving Cycles.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Driving cycle</td>
</tr>
<tr>
<td>Nationality</td>
</tr>
<tr>
<td>Speed curve</td>
</tr>
<tr>
<td>Type</td>
</tr>
<tr>
<td>Cycle time</td>
</tr>
<tr>
<td>Distance</td>
</tr>
<tr>
<td>Aver. speed</td>
</tr>
<tr>
<td>Max. speed</td>
</tr>
</tbody>
</table>

The simulated vehicle speed for the driving cycles EC, Japan 10-15 and FTP 75 are shown in figure 6.

![Graphs of vehicle speed for different driving cycles](image)

**Fig. 6.** The simulated vehicle speed for the driving cycles EC, Japan 10-15 and FTP 75.

As can be seen in figure 6, the three driving cycles have all a relatively low average speed and the speed is transient. These are the best conditions for the 4QT system to be able to reduce the fuel consumption. At highway driving the potential is completely different. This is because high continuous power is needed from the ICE, as mentioned earlier, and the vehicle
speed is relatively constant. The working point of the ICE can no longer be chosen freely. The fuel consumption can therefore not be reduced and the power flow in the electrical components should be low in order to have low losses in the system. The greatest fuel savings are achieved at city traffic driving as for most hybrid systems. The simulation results of the fuel consumption for the 4QT system are shown in figure 7.

Fig. 7. Calculated fuel consumption of the 4QT system compared to the conventional vehicle.

The 4QT system is compared with a similar conventional vehicle (ICE only) with the same weight. The fuel saving is 40%, 44% and 43% for the driving cycles EC, Japan 10-15 and FTP 75, respectively. The five speed gearbox, the engine flywheel, the start motor and the generator in the conventional vehicle are replaced by the 4QT system, the traction battery and the single speed gearbox in the novel hybrid. The weight difference can be neglected if a small traction battery is used. A detailed analysis of the system behaviour is shown in table 3.

TABLE 3. ANALYSIS OF THE SYSTEM BEHAVIOUR FOR THREE DRIVING CYCLES.

<table>
<thead>
<tr>
<th>Driving cycle</th>
<th>EC</th>
<th>Japan 10-15</th>
<th>FTP 75</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fuel consumption, corrected for SOC [l/100 km]</td>
<td>4.1</td>
<td>3.6</td>
<td>4.0</td>
</tr>
<tr>
<td>Fuel saving compared to conv. vehicle [%]</td>
<td>40</td>
<td>44</td>
<td>43</td>
</tr>
<tr>
<td>Duration in electric mode [s]</td>
<td>946.8</td>
<td>559.9</td>
<td>2064.6</td>
</tr>
<tr>
<td>Driven distance in electric mode [km]</td>
<td>5.4</td>
<td>3.0</td>
<td>11.3</td>
</tr>
<tr>
<td>Average speed in electric mode [km/h]</td>
<td>20.5</td>
<td>19.3</td>
<td>19.7</td>
</tr>
<tr>
<td>Duration in hybrid mode [s]</td>
<td>273.2</td>
<td>75.1</td>
<td>412.4</td>
</tr>
<tr>
<td>Driven distance in hybrid mode [km]</td>
<td>5.6</td>
<td>1.2</td>
<td>6.4</td>
</tr>
<tr>
<td>Average speed in hybrid mode [km/h]</td>
<td>74.1</td>
<td>55.9</td>
<td>55.5</td>
</tr>
<tr>
<td>Maximum battery power during cycle [kW]</td>
<td>13.9</td>
<td>13.6</td>
<td>15.8</td>
</tr>
<tr>
<td>Minimum speed of DRM [rpm]</td>
<td>-937</td>
<td>-793</td>
<td>-1285</td>
</tr>
<tr>
<td>Maximum speed of DRM [rpm]</td>
<td>4020</td>
<td>2341</td>
<td>3113</td>
</tr>
<tr>
<td>Maximum torque of DRM [Nm]</td>
<td>135.2</td>
<td>97.9</td>
<td>135.1</td>
</tr>
<tr>
<td>Minimum power of DRM [kW]</td>
<td>-8.5</td>
<td>-7.1</td>
<td>-16.8</td>
</tr>
<tr>
<td>Maximum power of DRM [kW]</td>
<td>22.9</td>
<td>7.5</td>
<td>12.7</td>
</tr>
<tr>
<td>Minimum speed of ST [rpm]</td>
<td>4232</td>
<td>2484</td>
<td>3221</td>
</tr>
<tr>
<td>Maximum speed of ST [rpm]</td>
<td>-87.9</td>
<td>-83.5</td>
<td>-111.3</td>
</tr>
<tr>
<td>Maximum torque of ST [Nm]</td>
<td>128.8</td>
<td>105.2</td>
<td>157.6</td>
</tr>
<tr>
<td>Minimum power of ST [kW]</td>
<td>-33.0</td>
<td>-21.0</td>
<td>-27.0</td>
</tr>
<tr>
<td>Maximum power of ST [kW]</td>
<td>11.7</td>
<td>11.6</td>
<td>13.6</td>
</tr>
</tbody>
</table>

The duration in electric mode in table 3 is defined as the total cycle time minus the duration in hybrid mode. I.e. it includes even the time when the vehicle is at stand still.

Hybrid mode means that the ICE is on. It can be seen in table 3 that the ICE is frequently turned off. The strategy is to turn off the engine at low speed and load, when braking and at stand still. For the simulated driving cycles, which have relatively low power demands, the vehicle is run in electric mode for the majority of time. Furthermore, the required battery power is lower than 16 kW. One machine generates 33 kW at the most, but since the other machine simultaneously acts as motor, the battery only needs to handle the power difference and can consequently be dimensioned for lower power than the electrical machines, respectively.

If mapping the working point of the DRM for a lot of driving cycles and maximum acceleration condition, an overall operating window can be found, as shown in figure 8. A similar mapping can be done for the ST, as shown in figure 9.

The speed of the DRM is a slip speed and equal to the speed difference between the out- and ingoing shafts of the 4QT. The maximum power is transient. No field weakening is needed for the DRM.

The maximum torque and power in figure 9 are transient. The ST needs to be field weakened above 2000 rpm, according to figure 9.
The average required performance of the electrical machines of the 4QT is calculated in two minutes interval for different driving cycles. The information is used to estimate the required continuous performance of the machines as shown in table 4.

<table>
<thead>
<tr>
<th>Machine</th>
<th>DRM</th>
<th>ST</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>Torque</td>
<td>120</td>
<td>60</td>
</tr>
<tr>
<td>Speed</td>
<td>2000</td>
<td>4000</td>
</tr>
</tbody>
</table>

It is evident when comparing table 4 with the overall operation windows, that the required continuous performance is much lower than the maximum performance. This is especially the case for the torque of the ST, which has a maximum of 170 Nm, but a continuous value of only 60 Nm. Such data can be used to cost optimise the electrical machines. It is well known that electrical machines are magnetically designed by the required transient performance, respectively thermally designed by the required continuous performance.

IV. MACHINE DESIGN

The available space for the electrical machine in a 4QT system for passenger cars is very limited. Front wheel drive is most common today. This makes the axial build of the machine limited by the width of the engine compartment, which also shall fit the engine and the gearbox, lengthwise. The outer diameter of the machine is limited by the distance to the wheel shaft. Furthermore, it is of course always required in a hybrid vehicle that all components shall have as low weight as possible. These requirements lead to the conclusion that the electrical machine unit needs to be compact. The proposed laboratory prototype to be built in the near future is therefore equipped with permanent magnet machines with high pole numbers.

The electrical machines can have many different configurations and three examples are illustrated in figure 10. The configuration to be chosen should be decided from the maximum allowable machine dimensions and the required machine performance.

The axial/radial-radial flux configuration as shown in figure 10 has been chosen for the laboratory prototype. The choice is based on the required performance as given by the simulation results. The DRM needs to develop a high continuous torque and effort has therefore been made to design a machine with a big air gap surface. Since the available space is very limited, a 3D flux topology with a Soft Magnetic Composites (SMC) iron core is proposed. A toroid winding ensures short end windings. The ST will instead be made with iron laminations with better magnetic properties, because it needs to generate a very high transient torque and low magnetic saturation is wanted. The ST will be equipped with tooth windings to keep the end windings and the axial build short. What is also worth mentioning regarding the three topologies in figure 10 is that the first two concepts require equal pole numbers of the two machines, since they are magnetically coupled to each other. The third topology can have different pole numbers on the electrical machines, because the machines work separately. Freedom of choosing the pole numbers independently can be advantageous if the machines run at different speeds. At the same time possible transient magnetic phenomenon causing stray losses can be avoided. A CAD drawing of the laboratory prototype machine unit is shown in figure 11.

Fig. 11. A CAD drawing of the laboratory prototype machine unit.

Attention must be put on the cooling aspects for the machine unit. There are several possible methods. Internal air cooling, forced air cooling, water cooling of the housing and direct oil cooling of windings and core or combinations of these are some alternatives. The mentioned cooling methods can all be implemented relatively easy in a hybrid application, since a conventional car is already equipped with air-, oil- and water cooling systems. The best compromise regarding cost and performance should be chosen.

A test bench has been built in the Electrical Machine Laboratory at the Royal Institute of Technology, where future laboratory prototypes will be tested.

V. CONCLUSION

A novel electrical machine drive system called the Four Quadrant Transducer for hybrid electric vehicles has been presented. Ideas about suitable topologies for designing a
compact machine unit are discussed. The complete hybrid system has been simulated and compared with a similar conventional vehicle, or non-hybrid. The results show that fuel savings of more than 40% compared to a conventional vehicle can be achieved.

REFERENCES

Winding Factors and Joule Losses of Permanent Magnet Machines with Concentrated Windings

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Winding Factors and Joule Losses of Permanent Magnet Machines with Concentrated Windings

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Abstract—The torque to MMF ratio of a permanent magnet machine with concentrated windings is normally much lower than for the more traditionally used distributed windings. BLDC and AC machines with concentrated windings usually have a slot pitch of only 2/3 the length of the pole pitch, which results in a poor fundamental winding factor of 0.866. This can be compared to the ideal winding factor of one, which can easily be acquired using distributed windings. However, by choosing better combinations of the pole and slot numbers for a machine with concentrated windings, the winding factor can be substantially increased. Moreover, it is also possible to achieve low cogging torque without skewing, simply by selecting appropriate combinations. The presented theory includes both single- and double-layer configurations of concentrated windings. The theory is verified by using FEM analysis.

This paper describes methods for designing high performance permanent magnet machines with concentrated windings.

Index Terms—Winding factors, permanent magnet machines, concentrated windings, EMF phasors, Joule losses, cogging torque, end windings.

II. CALCULATION OF WINDING FACTORS

The average electromagnetic torque of a surface mounted permanent magnet machine with concentrated windings fed with sinusoidal currents is given by

\[ T = \frac{1}{4\pi} k_w n_t n_i Q \hat{B}_{g10} A_g \hat{i} \cos(\theta), \]  

where \( k_w \) is the fundamental winding factor, \( n_t \) is the number of layers in each slot (1 or 2), \( n_i \) is the number of turns around each tooth, \( Q \) is the number of slots, \( \hat{B}_{g10} \) is the peak value of the fundamental no-load air gap flux density, \( A_g \) is the air gap area, \( \hat{i} \) is the peak value of the phase current to the machine and \( \theta \) is the current angle. \( \theta \) is equal to zero for q-current only.

The winding factor is proportional to the electromagnetic torque. I.e. an electrical machine with a low winding factor needs to compensate its lower torque with higher current or more number of turns, which both are inversely proportional to the winding factor. Thus, compared to a machine equipped with the reference winding (see section I), the resistive losses will increase. For example a machine with a winding factor of 0.866 has 15.5 percent higher current density and 33.3 percent higher Joule losses compared to the reference machine for the same torque, assuming that the machines have equal slot fill factors and comparable magnetic design and that the end windings are disregarded.

Fig. 1. Winding layouts for four-pole PM machines with a) distributed windings, b) single-layer concentrated windings and c) double-layer concentrated windings.
A. Theory for concentrated windings

The following characteristics and constraints are put on the concentrated windings in this paper [1]:

- The number of phases, \( m \), is equal to three.
- The winding system is symmetrical: \( Q/m \) is an integer.
- \( Q/p < m \), where \( p \) is the number of poles.
- \( Q \neq p \).
- The number of layers per slot, \( n_t \), is 1 or 2.
- The number of winding elements is \( S = n_t Q \).
- The slot pitch is constant.

The winding factor can be written

\[
k_w = k_p \cdot k_d \cdot k_{skew}
\]

(2)

where \( k_p \) is the pitch factor, \( k_d \) is the distribution factor and \( k_{skew} \) is the skewing factor. The pitch factor can be calculated straightforwardly for any winding, but the distribution factor is not obviously calculated for certain types of fractional slot concentrated windings. A general calculation method that incorporates both the pitch and distribution factors is therefore presented. The skewing factor will be discussed later, but machines with concentrated windings are normally not skewed.

To be able to calculate the winding factors for different pole and slot combinations, the electromotive force (EMF) phasor for each winding element needs to be defined. The winding element, \( i \), has the EMF phasor (in per unit):

\[
\vec{E}_i = e^{j(\gamma_i)}.
\]

(3)

Its argument is given by

\[
\gamma_i = \frac{\pi p}{Q} \cdot i.
\]

(4)

It is convenient to draw the phasors. The total number of phasors that has to be calculated is always \( Q \), even if the number of layers per slot is two, because two elements in the same slot have equal phasors. In a three-phase system, the phase EMF phasors are symmetrically displaced 120° from each other according to figure 2a. All winding elements must be assigned to the most appropriate phase winding. The winding element, \( i \), is assigned to the phase winding that has its phasor closest or oppositely closest to the phasor of \( i \). An important constraint for concentrated windings is of course that the two closest winding elements next to a tooth belongs to the same phase, in order to enable a tooth coil. Consequently, these two winding elements have opposite winding directions. The winding elements assigned to each phase must be connected (in series or parallel) depending on their phasor orientation. The resulting phase EMF phasor is the sum of the element phasors assigned to that phase, as given by (in per unit)

\[
\vec{E}_{phase, pu} = \sum_{S/3} \vec{E}_i.
\]

(5)

where \( S \) is the number of winding elements. The winding factor is found by dividing the magnitude of the resulting phase EMF phasor by the number of winding elements per phase, as given by

\[
k_w = \frac{\left| \vec{E}_{phase, pu} \right|}{S/3}.
\]

(6)

An example is given in Fig. 2b, showing the winding element phasors and the resulting phase EMF phasor for a four pole and six slot machine with double-layer concentrated windings (see also Fig. 1c). It is practical to start the numbering of the slots and winding elements by zero.

The maximum winding factor of one is achieved if all the winding element phasors for a phase point in exactly the same direction, which is true for a distributed winding with one slot per pole and phase (see Fig. 1a). The resulting phase EMF phasors would then correspond to Fig. 2a. A winding factor of one is not possible with a concentrated winding.
B. Winding factors

Fundamental winding factors for single-layer concentrated windings for different pole- and slot combinations are given in Table 1 and Fig. 4. The highest winding factor for each pole number is shown in bold in the table.

<table>
<thead>
<tr>
<th>Q</th>
<th>p</th>
<th>2</th>
<th>4</th>
<th>6</th>
<th>8</th>
<th>10</th>
<th>12</th>
<th>14</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td></td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>6</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.866</td>
<td>0.866</td>
<td>0.500</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>9</td>
<td></td>
<td>*</td>
<td>0.736</td>
<td>0.667</td>
<td>0.960</td>
<td>0.960</td>
<td>0.667</td>
<td>0.218</td>
<td>0.177</td>
</tr>
<tr>
<td>12</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.866</td>
<td>0.960</td>
<td>*</td>
<td>0.966</td>
<td>0.866</td>
<td></td>
</tr>
<tr>
<td>15</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.247</td>
<td>0.383</td>
<td>0.866</td>
<td>0.808</td>
<td>0.957</td>
<td>0.957</td>
</tr>
<tr>
<td>18</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.473</td>
<td>0.676</td>
<td>0.866</td>
<td>0.844</td>
<td>0.960</td>
<td></td>
</tr>
<tr>
<td>21</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.248</td>
<td>0.397</td>
<td>0.622</td>
<td>0.866</td>
<td>0.218</td>
<td>0.177</td>
</tr>
<tr>
<td>24</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.430</td>
<td>0.561</td>
<td>0.960</td>
<td>0.966</td>
<td>0.866</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 4. Winding factors for concentrated windings with one layer per slot.

Fundamental winding factors for double-layer concentrated windings are given in Table 2 and Fig. 5.

<table>
<thead>
<tr>
<th>Q</th>
<th>p</th>
<th>2</th>
<th>4</th>
<th>6</th>
<th>8</th>
<th>10</th>
<th>12</th>
<th>14</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td></td>
<td>0.866</td>
<td>0.866</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>6</td>
<td></td>
<td>*</td>
<td>0.866</td>
<td>0.866</td>
<td>0.500</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>9</td>
<td></td>
<td>*</td>
<td>0.617</td>
<td>0.866</td>
<td>0.945</td>
<td>0.945</td>
<td>0.764</td>
<td>0.473</td>
<td>0.175</td>
</tr>
<tr>
<td>12</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.866</td>
<td>0.933</td>
<td>*</td>
<td>0.933</td>
<td>0.866</td>
<td></td>
</tr>
<tr>
<td>15</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.481</td>
<td>0.621</td>
<td>0.866</td>
<td>0.906</td>
<td>0.951</td>
<td>0.951</td>
</tr>
<tr>
<td>18</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.543</td>
<td>0.647</td>
<td>0.866</td>
<td>0.902</td>
<td>0.931</td>
<td></td>
</tr>
<tr>
<td>21</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.468</td>
<td>0.565</td>
<td>0.521</td>
<td>0.866</td>
<td>0.851</td>
<td></td>
</tr>
<tr>
<td>24</td>
<td></td>
<td>*</td>
<td>*</td>
<td>0.463</td>
<td>0.760</td>
<td>0.866</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Fig. 5. Winding factors for concentrated windings with two layers per slot.

C. MMF-waves

Once the winding layout is defined, it is possible to calculate the magnetomotive force (MMF) created by the stator currents. The MMF, \( M \), is related to the magnetic field intensity, \( H \), for a coil with \( N \) turns carrying a current \( I \), as given by Ampère’s circuital law [2]

\[
M = NI = \oint H \cdot d\ell
\]

Some parameter definitions of an MMF-wave created by a single coil are described in Fig. 6. The coil pitch for a single-layer winding, \( \tau_c \), can also be used for calculations of the MMF-waves for the double-layer winding, because all the flux is assumed to pass only through the teeth.

The MMF-wave of order \( n \) for a coil fed with sinusoidal current is expressed by [2]

\[
M_{(n)} = \frac{4N}{np\pi} \sin \left( \frac{n\pi}{2} \cdot \frac{\tau_c}{\tau_p} \right) \cos(n(\gamma - \vartheta)) \cdot I \cos(\omega t) =
\]

\[
= \frac{4N}{np\pi} \sin \left( \frac{n\pi}{2} \cdot \frac{\tau_c}{\tau_p} \right) \cdot \frac{1}{2} \left[ \cos(\omega t - n(\gamma - \vartheta)) + \cos(\omega t + n(\gamma - \vartheta)) \right]
\]

(8)
where $\omega$ is the angular electrical speed and $t$ is time. Equation (8) implies that the MMF mathematically can be separated into two waves, rotating in opposite directions. The waveform pattern must be repeated at least once over the whole periphery of the machine, due to the need of symmetry, which means that the minimum harmonic order is $2/p$. The amplitude of the fundamental MMF-wave is reduced for a short-pitched coil, as shown in (8), by a term called pitch factor:

$$k_{p(e)} = \sin\left(\frac{n\pi}{2} \frac{\tau_c}{\tau_p}\right). \tag{9}$$

The MMF-pattern is not identical for each pole for fractional slot windings. This is compensated for by the electrical displacement factor as given by

$$n_y = \left(\frac{x}{6q} - \frac{1}{3}\right) \tag{10}$$

where $x$ is the physical phase displacement in number of slots. The three-phase MMF of order $n$ for a machine with symmetrical phase windings is

$$M_{3ph(n)} = \sum_{c=1}^{N_c} \sum_{p=1}^{n_c} 4N_p \sin\left(\frac{n\pi}{2} \frac{\tau_c}{\tau_p}\right) \left(M^+ + M^\right) \tag{11}$$

where $\tau_c$ is the coil span of the $c$:th coil, $n_c$ is the number of coils, while $M^+$ and $M^-$ are the forward and backward rotating MMF-waves, respectively, as given by

$$M^+ = I \cos(\omega t - n(\gamma - \vartheta)_c) \left[\frac{1}{2} + \cos\left(\frac{2\pi}{3}(n-1)+2\pi \cdot n \cdot n_c\right)\right] \tag{12}$$

$$M^- = I \cos(\omega t + n(\gamma - \vartheta)_c) \left[\frac{1}{2} + \cos\left(\frac{2\pi}{3}(n+1)+2\pi \cdot n \cdot n_c\right)\right] \tag{13}$$

The three-phase MMF and its harmonic content for the machines in Fig. 1a, 1b and 1c are shown in Fig. 7, Fig. 8 and Fig. 9, respectively. The current loading is kept constant for all three machines in the analysis. The peak-peak value of the total phase-MMF for the distributed winding is the reference and equal to one per unit. The influence of slotting is disregarded.

A characteristic for concentrated windings is that they generate both odd and even MMF-waves. I.e. the leakage inductance is increased compared to a machine with distributed windings, because all the extra MMF-harmonics create additional flux in the machine. Some concentrated windings also produce subharmonics, as for the single-layer winding in Fig. 8. An increased leakage inductance results in higher core losses and a need for higher inverter rating, but lower field weakening current in the constant power region.
Most machines with concentrated windings have trapezoidal EMF-waveforms and are fed with rectangular currents. These machines have simpler control algorithms and position sensors. The required core material must be slightly increased, due to the higher flux in the machine. Machines with sinusoidal currents, on the other hand, should have sinusoidal air gap flux density and have typically lower torque ripple.

III. JOULE LOSSES

The Joule losses dealt with in this paper concern DC losses only. AC losses due to skin effect, alternating magnetic fields or magnetic asymmetries are neglected. The Joule losses for one phase can therefore be written

$$P_{\text{winding}} = R I^2 - \frac{L_C}{c_f k_w^2}$$  \hspace{1cm} (14)

where $R$ is the phase resistance, $I$ is the RMS value of the current, $L_C$ is the average conductor length and $c_f$ is the slot fill factor. As can be seen, the winding losses are proportional to the conductor length and inversely proportional to the slot fill factor and the winding factor squared.

The average conductor length can be written

$$L_C = k_c 2\pi r_w + L,$$  \hspace{1cm} (15)

where $k_c$ is a parameter that depends on the geometry of the coil, $r_w$ is the average radius of the winding (the distance from the centre of the machine to the middle of the slot) and $L$ is the active length of the machine. For simplicity reasons the slot width and tooth width are chosen equal, i.e. a half slot pitch each. The constants qualitatively chosen for the different winding types are shown in Table 3.

<table>
<thead>
<tr>
<th>TABLE 3</th>
<th>WINDING CONSTANS</th>
</tr>
</thead>
<tbody>
<tr>
<td>$c_f$</td>
<td>Distributed 0,35</td>
</tr>
<tr>
<td></td>
<td>Concentrated, 1 0,50</td>
</tr>
<tr>
<td></td>
<td>Concentrated, 2 0,46</td>
</tr>
<tr>
<td>$k_c$</td>
<td>$1,6 Q/p$</td>
</tr>
<tr>
<td></td>
<td>$1,36$</td>
</tr>
<tr>
<td></td>
<td>$0,93$</td>
</tr>
</tbody>
</table>

The fill factors of the concentrated windings are higher, than that of the distributed, because it is possible to manufacture their stators in segments. Segmented designs can be manufactured with higher degree of automation and allow for better control of the coils during the production process, which increases the fill factor. Using two layers per slot means that two phases share the slot space and more insulation is needed. Thus, the double-layer winding has a lower slot fill factor than the single-layer winding.

A relative comparison of the winding Joule losses of machines with the three winding types is shown in Fig. 10. (The combinations with the highest winding factors from Table 1 and 2 are used.) The average winding radius is chosen equal to the active length in this study.

![Fig. 10. The Joule losses of three winding configurations relative the Joule losses of a two pole machine with distributed windings.](image)

The iron losses are not considered, so one should be careful in comparing Joule losses of different pole numbers (which of course should have different magnetic design layouts). The reductions of the Joule losses for machines with concentrated windings in relation to the Joule losses for a machine with distributed windings of the same pole number are shown in Fig. 11.

![Fig. 11. The reduction of the Joule losses for machines with concentrated windings compared to the Joule losses for a machine with distributed windings of the same pole number. (Negative values means increased Joule losses.)](image)

IV. COGGING TORQUE MINIMIZATION

A stator with the reference winding (see section I) can easily be skewed one slot pitch to cancel out its cogging torque. The drawback is that the fundamental winding factor and the EMF will drop 4.5 percent, as given by the fundamental skewing factor (which is a part of the winding factor) [2]

$$k_{\text{skew}} = \sin \left( \frac{z\pi p}{2Q} \right),$$  \hspace{1cm} (16)

where $z$ is the skew in number of slots.

It is not advantageous to skew a stator with concentrated windings, because the number of slots per pole and phase is lower than one. Other methods of reducing the cogging torque...
can be used [3]-[9], but normally the reduction techniques often contribute to a machine with weaker performance. A proposal is to select a pole and slot combination with a built-in geometric asymmetry that will give a high average performance as well as a low cogging torque.

The number of cogging periods per rotor revolution is given by the least common multiple (LCM) of the pole and slot numbers [9]. Since higher cogging frequency most certainly means lower cogging amplitude, a good strategy is to select a high LCM. The LCM for different combinations is shown in Table 4.

<table>
<thead>
<tr>
<th>Q</th>
<th>2</th>
<th>4</th>
<th>6</th>
<th>8</th>
<th>10</th>
<th>12</th>
<th>14</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>6</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>6</td>
<td>*</td>
<td>12</td>
<td>24</td>
<td>30</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>*</td>
</tr>
<tr>
<td>9</td>
<td>*</td>
<td>36</td>
<td>72</td>
<td>90</td>
<td>36</td>
<td>126</td>
<td>144</td>
<td></td>
</tr>
<tr>
<td>12</td>
<td>*</td>
<td>24</td>
<td>60</td>
<td>*</td>
<td>84</td>
<td>48</td>
<td></td>
<td></td>
</tr>
<tr>
<td>15</td>
<td>*</td>
<td>30</td>
<td>120</td>
<td>30</td>
<td>60</td>
<td>210</td>
<td>240</td>
<td></td>
</tr>
<tr>
<td>18</td>
<td>*</td>
<td>*</td>
<td>72</td>
<td>90</td>
<td>36</td>
<td>126</td>
<td>144</td>
<td></td>
</tr>
<tr>
<td>21</td>
<td>*</td>
<td>*</td>
<td>168</td>
<td>210</td>
<td>84</td>
<td>42</td>
<td>336</td>
<td></td>
</tr>
<tr>
<td>24</td>
<td>*</td>
<td>*</td>
<td>*</td>
<td>120</td>
<td>*</td>
<td>168</td>
<td>48</td>
<td></td>
</tr>
</tbody>
</table>

It is shown in Table 4 that the pole and slot combinations which are multiples of 2/3 all have a low LCM, which strongly indicate that they have high cogging. For higher pole numbers it is easier to find combinations with both high winding factors and high LCMs. For example the combination 14 poles and 15 slots has a winding factor of 0,951 (see Table 2) and a LCM as high as 210. It is thus a high performance PM machine with concentrated windings. A FEM calculated no-load flux distribution of such a machine is shown in Fig. 12.

The cogging torque of the machine in Fig. 12, with an air gap diameter of 204 mm and a slot opening width of 2 mm, is calculated for one slot pitch using the finite element software FLUX2D®. (The active length is chosen to be one meter for the computation.) The result is shown in Fig. 13.

The peak-peak value of the cogging torque is lower than 0,4 Nm/m, which is only a fractional percentage of the nominal torque. It is shown in Fig. 13 that one slot pitch has 14 cogging periods, same value as the pole number, which means 210 periods per rotor revolution. The result is an extremely high cogging frequency with low amplitude.

V. AXIAL BUILD OF MACHINES

It is important in some applications that the axial build of the machine is short (e.g. electric vehicles) [3]. One parameter that has a considerable impact on the axial build is the length of the end windings, especially for machines with relatively short active lengths. Machines with distributed windings have typically long end windings, because the coil of a phase must cross the other phase coils. Since the phases are located closely together, more insulation material is used, which will add to the axial build of the end windings. The concentrated winding has phase coils that are wound around separate teeth, which means that the axial build is short. The single-layer concentrated winding has almost twice the axial end winding build, compared to double-layer windings. This is because the coil around one tooth consists of all conductors in the slot next to it, while for double-layer windings it consists of only half of these conductors. Fig. 14 illustrates the different axial build of the machines with equal active lengths.

The importance of the axial build of the end-windings is as already mentioned of course depending on the active length of the machine. The longer the active length, the less important are the end-windings. The length of the end-windings is a function of the pole- and slot numbers, the average radius of the winding and the distribution of the coils.

A machine with double-layer concentrated windings has the greatest potential to be the most compact unit of the three described types of electrical machines.
VI. CONCLUSION

A theory for calculation of winding factors for electrical machines with concentrated windings has been presented. The winding factors influence on the Joule losses has been discussed. It is shown that by selecting appropriate combinations of the pole and slot numbers, low Joule losses and low cogging torque can be acquired.
Design of Compact Permanent Magnet Machines for a Novel HEV Propulsion System

Design of Compact Permanent Magnet Machines for a Novel HEV Propulsion System

Freddy Magnussen, Dr. Peter Thelin, Prof. Chandur Sadarangani
Royal Institute of Technology, SE-100 44 Stockholm, +46 8 790 77 75

Abstract

This paper presents the design of two permanent magnet machines for a novel propulsion system for hybrid electric vehicles called the Four Quadrant Transducer (4QT). The presented machines are designed for a medium-sized passenger car with front wheel drive. It is therefore essential that the machines are compact, since the available space in the engine compartment is very limited.

The first machine has a 3D-flux topology and uses Soft Magnetic Composites (SMC), i.e. iron powder, as the core material. The core is segmented in 48 parts, which individually consists of teeth and a rounded back. A rectangular conductor is toroidally wound around the back of each segment resulting in short end-windings and a high slot fill factor of 0.78.

The second electrical machine design includes features that are unique for its power class, e.g. a segmented core with grain-oriented silicon iron teeth and rectangular copper conductors. The result is a slot fill factor of 0.74, a high fundamental winding factor for a tooth winding of 0.95, a cogging torque close to zero and high efficiency.

Electromagnetic analyses of the proposed machines are performed with analytical calculations and by using the 3D- and the 2D- finite element methods, including torque ripple and eddy current calculations. A laboratory prototype has been built.

Keywords: Motor design, permanent magnet motor, propulsion system, finite element calculation.

1. System description

The Four Quadrant Transducer (4QT) propulsion system is a power-split hybrid concept and consists of two electrical machines integrated into one unit [1]-[2]. The armature windings and core of the first machine are mounted on a shaft mechanically connected to the output shaft of the internal combustion engine (ICE). The armature windings and core of the second machine are static. The two machines have a common rotor with permanent magnets, which is connected to the final gear. The 4QT hybrid system is illustrated in Figure 1.

![Figure 1: Layout of the Four Quadrant Transducer hybrid system (left). The electrical machine topology in the designed machine prototype (right).](image-url)
The main mission of the 4QT system is to assist the internal combustion engine in such a way that the ICE can be run along its optimum operation line (O.O.L), where the system efficiency of the vehicle is optimized. Any speed difference between the wanted speed on the output shaft and the engine speed, for a specific output power, is handled by the electrical machine connected to the engine by supplying AC-currents to its windings via slip rings. This speed-changing machine is called the Double Rotor Machine (DRM). In the same manner, the second electrical machine handles any torque difference. This torque-changing machine is called the Stator Machine (SM). The output power to the wheels is delivered from either the ICE or the battery or both. The peak power of the battery is chosen to be less than 20 kW, while the peak power of the ICE is 80 kW. The optimum operation line for the studied engine is shown in Figure 2.

![Optimum Operation Line of an ICE](image)

**Figure 2:** An optimum operation line of an internal combustion engine.

The studied concept is equipped with a final gear of constant gear ratio of 4:1 only, in order to make it as simple and transparent as possible for characterization purposes. Other fixed gear ratios, multiple gear ratios or shaft lock-up functions are other obvious alternatives, which can reduce the torque requirements of the 4QT, but at the same time add mechanical complexity to the system. The operating limits of the two electrical machines, for the conditions as mentioned in [2], are given in Figure 3. The maximum speed of the output shaft of the 4QT is 6300 rpm and equivalent to a vehicle speed of 180 km/h.

![Operating Limits of Two Electrical Machines](image)

**Figure 3:** Operating limits of the two electrical machines.

## 2. Design of the 3D-flux permanent magnet machine using SMC (DRM)

The speed-changing electrical machine (DRM) must deliver the same torque as the internal combustion engine at steady state, which is maximum 140 Nm. The maximum speed is around 4000 rpm. The speed of the DRM is defined as the speed difference between the output shaft of the 4QT and the engine shaft. No field weakening operation is needed. The machine is required to work both as a motor and as a
generator, according to the load specification shown in Figure 3. The peak power requirement of 53 kW is in generator mode. The continuous power requirement is roughly 20 kW.

2.1 Materials

The DRM has a 3D-flux topology and uses Soft Magnetic Composites (SMC), i.e. iron powder, as the core material. Iron powder has lower permeability and higher hysteresis losses than silicon steel sheets [3], but the powder has other unique advantages. The best advantage is the 3D-isotropically magnetic properties. Iron powder can also be pressed into complex parts and allow the coil and core to be fitted closely together. The heat transfer will often be superior when also taking into account the 3D-isotropically thermal properties of SMC. The rated torque of a well-designed SMC machine can therefore be increased compared to a laminated machine [4]. The B-H-curve and the loss characteristic of the iron powder SOMALOY™ 500 with 0.5% of the lubricant Kenolube from the manufacturer Höganäs AB is shown in Figure 4.

![Figure 4: Magnetic flux density versus magnetic field strength (left), and core loss versus frequency at a peak flux density of 1.0 T (right) for the iron powder SOMALOY™ 500.](image)

2.2 Electromagnetics

The permanent magnets (NdFeB) in the rotor with sixteen poles produce airgap flux in both the axial and radial directions. The core material must therefore carry flux in three dimensions. This kind of topology was first reported in [5], where a slotless prototype for an electric bicycle was built. The machine presented in this paper increases the concept’s torque density capability by adopting a slotted structure. Forty-eight segments build up the core that consists of teeth and back. In order to increase the torque density even further, a rectangular conductor is toroidally wound around the back of each segment, called Gramme-winding, making a coil with very short end-windings. The core and coils are shown in Figure 5.

![Figure 5: The dimensions of the core segment and the assembled 48 segments with coils of the DRM.](image)
Due to the segmentation, which allows for the rectangular conductor shape, the slot fill factor can be extremely high (0.78) and the overall machine size will not be limited to the capability of the tool for pressing the iron powder. The electrical machine parameters of the DRM are shown in Table 1.

Table 1: Electrical machine parameters of the Double Rotor Machine.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer radius of radial flux surface $R_1$</td>
<td>105 mm</td>
</tr>
<tr>
<td>Outer radius of axial flux surface $R_2$</td>
<td>104 mm</td>
</tr>
<tr>
<td>Inner radius of axial flux surface $R_i$</td>
<td>73 mm</td>
</tr>
<tr>
<td>Radius of upper core back $R_{back1}$</td>
<td>5 mm</td>
</tr>
<tr>
<td>Radius of lower part of core back $R_{back2}$</td>
<td>11 mm</td>
</tr>
<tr>
<td>Axial length of armature core $l_{CRM}$</td>
<td>68 mm</td>
</tr>
<tr>
<td>Active axial length $l_{active}$</td>
<td>66 mm</td>
</tr>
<tr>
<td>Length of straight part of core back $l_{back}$</td>
<td>18 mm</td>
</tr>
<tr>
<td>Height of slot $hsl$</td>
<td>14 mm</td>
</tr>
<tr>
<td>Width of slot $wsl$</td>
<td>4.4 mm</td>
</tr>
<tr>
<td>Airgap length $\delta$</td>
<td>1.5 mm</td>
</tr>
<tr>
<td>Length of permanent magnets $l_m$</td>
<td>3.5 mm</td>
</tr>
<tr>
<td>Average length of conductor $l_c$</td>
<td>150 mm</td>
</tr>
<tr>
<td>Total airgap area $A_{delta}$</td>
<td>78000 mm²</td>
</tr>
<tr>
<td>Number of poles $p$</td>
<td>16</td>
</tr>
<tr>
<td>Number of slots $Q$</td>
<td>48</td>
</tr>
<tr>
<td>Number of slots per pole and phase $qs$</td>
<td>1</td>
</tr>
<tr>
<td>No-load flux density (peak fundam.) $B_{0}$</td>
<td>0.70 T</td>
</tr>
<tr>
<td>No-load voltage @ 3600 rpm (RMS) $E_0$</td>
<td>189 V</td>
</tr>
<tr>
<td>Phase resistance @ 120 °C $R$</td>
<td>10.2 mΩ</td>
</tr>
<tr>
<td>Synchronous inductance $L_s$</td>
<td>0.11 mH</td>
</tr>
<tr>
<td>Number of turns per slot $ns$</td>
<td>12</td>
</tr>
<tr>
<td>Number of parallel coils per phase $cs$</td>
<td>4</td>
</tr>
<tr>
<td>Height of conductor $h$</td>
<td>1.06 mm</td>
</tr>
<tr>
<td>Width of conductor $w$</td>
<td>4.0 mm</td>
</tr>
<tr>
<td>Slot fill factor $cf$</td>
<td>0.78</td>
</tr>
<tr>
<td>Fund. winding factor, radial part $kwr$</td>
<td>0.96</td>
</tr>
<tr>
<td>Maximum efficiency $\eta$</td>
<td>&gt;0.94</td>
</tr>
</tbody>
</table>

The FEM-calculated flux density in the machine at full load is shown in Figure 6. Only 1/32-part of the machine is modelled, due to symmetry.

Figure 6: The 3D-FEM calculated flux density (Tesla) in the SMC machine at full load. One pole of the machine is also shown in a radial cut and an axial cut.

As can be seen in Figure 6, the intersection between the teeth and the core back is saturated. This region of the machine is the weakest part of the magnetic design. It is possible to increase the area of this intersection by decreasing the slot depth or the slot width, but at the same time the available space for the copper winding is also decreased. If the number of poles is increased, the required areas for the intersection and the core back itself are reduced (to the cost of a higher electrical frequency for the same speed) and the inner radius of the winding is increased, which means that the average conductor length is shortened. As a result, only machines with high number of poles (>10 poles) are practically feasible with the presented concept.

2.3 Torque ripple

Different torque ripple reduction techniques are applied in the electrical machine design. The permanent magnet arc of the radial part is equal to two slot pitches. The permanent magnets for the radial part are furthermore divided in three and each magnet is stepwise skewed by 1/3 slot pitch. The permanent magnets for the axial parts have an arc length of 2.5 slot pitches at their outer radii, while they have an arc.
length of 1.5 slot pitches at their inner radii. I.e. they have a shape with a built-in continuous skew of one slot pitch. The orientations of the permanent magnets are shown in Figure 6.

3. Design of the PM machine with concentrated windings (SM)

The torque-changing machine (SM) must deliver a peak torque of 170 Nm between standstill and the base speed of 2000 rpm, according to the load specification shown in Figure 3. This torque is just needed during accelerations or for starting in steep slopes and is not thermally dimensioning. The machine is therefore physically downscaled (the active weight is 17 kg) to reduce weight and cost and especially designed for overload conditions. At normal operating conditions the efficiency is overall high. The maximum speed is 6300 rpm, which means that field weakening operation is implemented. The machine is required to work both as motor and generator.

3.1 Materials

The core is segmented with 0.30 mm grain-oriented silicon iron sheet teeth and 0.35 mm non-oriented sheet stator back. A grain-oriented material has superior performance in respect of less saturation and lower losses in the direction of the grain orientation. The performance in the perpendicular direction is slightly poorer than for a non-oriented material, as shown in Figure 7. The grain-oriented material can therefore advantageously be used for the teeth, where the main flux is in the radial direction.

![Figure 7: Peak magnetic polarizations versus peak magnetic field strength for 0.30 mm grain- and 0.35 mm non-oriented silicon iron sheets (left). Core losses versus frequency at a peak flux density of 1.0 T for the same iron sheets tested according to the Epstein frame method (right).](image)

Due to the segmentation of the stator core, it is possible to use pre-made coils in the slots. The coils are made up of rectangular copper conductors manufactured under tension. The result is a high slot fill factor of 0.74. Due to the relatively large cross-section area of 2×3 mm² of the conductors, the eddy-currents induced at increasing frequency are not negligible. Methods to calculate the additional joule losses caused by skin effect and internal proximity effects and consequently how to reduce them are described in chapter 4.

3.2 Electromagnetics

BLDC- and synchronous PM machines with concentrated windings usually have a slot pitch of only 2/3 the length of the pole pitch [6], which results in a poor fundamental winding factor of 0.87. This can be compared to the ideal winding factor of unity, which can easily be acquired using distributed windings. The proposed machine shown in Figure 8 has an optimized pole-slot combination (14 poles and 15 slots)
to achieve a high winding factor for a tooth winding of 0.95 and a cogging torque almost equal to zero, without skewing (as shown in Figure 9).

Figure 8: The design of the Stator Machine showing from left: the dimensions, the no-load flux distribution and the winding distribution.

There are no symmetric conditions along the airgap, i.e. the machine needs to be modelled over its complete cross-section in the FEM-software. This characteristic is the reason for the extremely low cogging torque. Since the machine is unsymmetrical, this could lead to an undesired radial force component, which could contribute to additional noise, but this effect is not studied in detail. The electrical machine parameters of the SM are shown in Table 2.

Table 2: Electrical machine parameters of the Stator Machine.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Airgap diameter D_{air}</td>
<td>204 mm</td>
</tr>
<tr>
<td>Outer diameter of rotor D_o</td>
<td>240 mm</td>
</tr>
<tr>
<td>Inner diameter of stator D_i</td>
<td>122 mm</td>
</tr>
<tr>
<td>Active length l_{active}</td>
<td>65 mm</td>
</tr>
<tr>
<td>Height of slot h_{sl}</td>
<td>22.5 mm</td>
</tr>
<tr>
<td>Width of slot w_{sl}</td>
<td>13.5 mm</td>
</tr>
<tr>
<td>Airgap length delta</td>
<td>1.5 mm</td>
</tr>
<tr>
<td>Length of permanent magnets lm</td>
<td>4.5 mm</td>
</tr>
<tr>
<td>Height of conductor h</td>
<td>2.0 mm</td>
</tr>
<tr>
<td>Width of conductor w</td>
<td>3.0 mm</td>
</tr>
<tr>
<td>Average length of conductor L_c</td>
<td>110 mm</td>
</tr>
<tr>
<td>Total airgap area A_{delta}</td>
<td>41700 mm²</td>
</tr>
<tr>
<td>Number of poles p</td>
<td>14</td>
</tr>
<tr>
<td>Number of slots Q</td>
<td>15</td>
</tr>
<tr>
<td>No-load flux density (peak fundam.) B_{phy}</td>
<td>0.73 T</td>
</tr>
<tr>
<td>No-load voltage @ 2000 rpm (RMS) E_{q0}</td>
<td>118 V</td>
</tr>
<tr>
<td>Phase resistance @ 120 °C R</td>
<td>22.2 mOhm</td>
</tr>
<tr>
<td>Synchronous inductance L_s</td>
<td>0.5 mH</td>
</tr>
<tr>
<td>Number of turns per tooth nt</td>
<td>10</td>
</tr>
<tr>
<td>Number of winding layers per slot n_l</td>
<td>2</td>
</tr>
<tr>
<td>Number of parallel tooth coils c_t</td>
<td>2</td>
</tr>
<tr>
<td>Slot fill factor c_f</td>
<td>0.74</td>
</tr>
<tr>
<td>Fundamental winding factor kw</td>
<td>0.95</td>
</tr>
<tr>
<td>Maximum efficiency η</td>
<td>&gt;0.95</td>
</tr>
</tbody>
</table>

3.3 Torque ripple

The FEM-calculated cogging torque with the corresponding harmonic content is shown in Figure 9.

Figure 9: FEM-calculated cogging torque with the corresponding harmonic content for the designed permanent magnet machine with concentrated windings.

The FEM-calculated electromagnetic torque at nominal load with only q-current is shown in Figure 10.
The highest torque ripple component has six times higher frequency than the fundamental electrical frequency. The 18th harmonic is the second largest torque ripple component followed by the 12th harmonic, i.e. they are all multiples of six. The peak-to-peak torque ripple is 3.0 Nm or 3.5% of the estimated nominal torque of 85 Nm, which is regarded as a low value for the suggested hybrid vehicle application. The FEM-calculated no-load voltages are shown in Figure 11.

Figure 11: FEM-calculated phase (A) and line-line (BC) EMF-curves with corresponding harmonic content (peak values) at base speed.

4. Eddy current calculations

Eddy currents are induced in conductors carrying alternating current. In the cross-section of such a conductor, the centre has the maximum inductance. The inductance decreases in the radial direction meeting an inductance minimum at the periphery of the conductor. The result is that the current density is not uniform over the cross-section. The current density has a minimum in the centre and maximum at the periphery of the conductor. This current distribution increases the effective resistance of the conductor. The effect, called the skin-effect, must be considered when large conductor cross-sections are used and/or high frequencies are applied. Since relatively large rectangular conductor cross-sections are used in the presented electrical machines, the eddy currents can therefore not be neglected. The DC resistance of the phase winding must be multiplied with a resistance factor, which takes into account the eddy currents. The average resistance factor for a conductor layer in a slot with rectangular conductor cross-section is given by [7]

\[ k_{sp,slot} = \varphi(\zeta) + \frac{\gamma^2 - 1}{3} \cdot \psi(\zeta) \]  

(1)

where \( \gamma \) is the number of conductor layers in the slot and
\[
\varphi(\xi) = \xi \frac{\sinh 2\xi + \sin 2\xi}{\cosh 2\xi - \cos 2\xi}
\]
(2)

\[
\psi(\xi) = 2\xi \frac{\sinh \xi - \sin \xi}{\cosh \xi + \cos \xi}
\]
(3)

where the dimensionless, so-called “reduced conductor height” is

\[\xi = \alpha \cdot h\]
(4)

with

\[
\alpha = \sqrt{\frac{xw \pi \mu_0 f}{s \rho_{Cu}}} = \sqrt{\frac{xw}{s} \cdot \frac{1}{\delta_p}}
\]
(5)

where \(x\) is the number of conductors in the \(p\):th layer, \(w\) is the conductor width, \(h\) is the conductor height, \(s\) is the width of the slot and \(\delta_p\) is the penetration depth. Figure 12 illustrates the relationships of the slot geometry and shows the penetration depth of a copper conductor.

Figure 12: Definition of conductor and slot dimensions (left). The penetration depth of a copper conductor versus frequency at a temperature of 120 °C (right).

The first term in Equation 1 describes the skin-effect, which is caused by alternating current in the studied conductor. The second term describes the internal proximity effect, which is caused by AC-currents in other conductors in the same slot. The external proximity effect, which is caused by external alternating magnetic fields, e.g. from AC-currents in neighbouring slots or permanent magnets in motion, is disregarded in this paper, since such fields are assumed to be in the teeth and not in the slots.

The height of a rectangular conductor or the diameter of a round conductor should always be smaller (preferably significantly smaller) than the penetration depth in order to limit the eddy current losses in the conductor. If so, the value of \(\xi\) will be lower than one and Equation 1 can be simplified into

\[
k_{sp,slot} = 1 + \frac{y^2 - 0.2}{9} \cdot \xi^4 \quad (0 \leq \xi \leq 1).
\]
(6)

For the end-windings of an electrical machine, where the conductors are located in air, the eddy-current losses are lower and the resistance factor is given by

\[
k_{sp,end-windings} = 1 + \frac{y^2 - 0.8}{36} \cdot \xi^4 \quad (0 \leq \xi \leq 1).
\]
(7)

The winding losses of a three-phase machine including the skin and internal proximity effects are
where $R$ is the phase DC-resistance at actual temperature and $I$ is the phase RMS-current.

5. Torque ripple calculations

Torque ripple can appear with or without the presence of currents. If currents are applied the torque ripple is caused by interactions between permanent magnet generated flux and space harmonics of the winding layout and time harmonics in the drive current.

Torque ripple in a permanent magnet synchronous machine is caused by non-sinusoidal back-EMF or current waveforms and permeance variations in the airgap due to slotting or rotor saliency. The torque produced by permeance variations can exist without currents, and is then called cogging torque, or with currents, and is then called reluctance torque. In a surface mounted permanent magnet machine the reluctance torque is almost zero, because the rotor has almost no saliency (the relative permeability of rare earth permanent magnets are close to unity). The main torque in such a machine is called electromagnetic torque, and it is generated by rotor flux from the permanent magnets linked with the stator windings in combination with the stator currents. The instantaneous electromagnetic torque for a three-phase electrical machine is given by [8]

$$T_{em}(t) = \frac{e_x(t) \cdot i_x(t) + e_y(t) \cdot i_y(t) + e_z(t) \cdot i_z(t)}{\omega_{mech}}$$  \quad (9)

where $e_x(t)$ and $i_x(t)$ is the instantaneous values of the back-EMF and current in phase $A$, respectively, and $\omega_{mech}$ is the angular speed of the rotor. For a machine with a balanced Y-connected three-phase winding, its phase back-EMF and phase current contains only odd harmonics. Furthermore, the third current harmonic is eliminated, due to the Y-connection. Since the phase parameters are displaced 120 electrical degrees from each other, the instantaneous electromagnetic torque will contain an average component and harmonics of order of six and can be expressed as [8]

$$T_{em}(t) = T_0 + \sum_{n=4}^{\infty} T_{6n} \cos(m6\omega t)$$  \quad (10)

where $T_0$ is the average torque and $T_{6n}$ is the peak value of a harmonic torque component, given by

$$T_0 = \frac{3}{2\omega_{mech}} \cdot \sum_{n=1}^{\infty} E_n I_n \begin{cases} (k = 0,1,2,3,...) \\ (n \neq 3,6,9,12,...) \end{cases}$$  \quad (11)

$$T_{6n} = \frac{3}{2\omega_{mech}} \cdot \sum_{n=1}^{\infty} (E_{n+6n} + E_{n-6n}) I_n \begin{cases} (k = 0,1,2,3,...) \\ (n \neq 3,6,9,12,...) \end{cases}$$  \quad (12)

where $E_n$ is the amplitude of the $n^{th}$ time harmonic of the phase EMF, which is produced by the $n^{th}$ space harmonic of the airgap flux density and $I_n$ is the amplitude of the $n^{th}$ time harmonic of the phase current, which depends on the current waveform.

Torque ripple calculations are preferably performed using finite element analysis. The torque ripple factor (in percent) can be defined as the ratio of the peak-to-peak torque ripple to the average torque [8]

$$T_{RF} = \frac{\Delta T_p}{T_0} \cdot 100\% .$$  \quad (13)
6. Laboratory prototype

6.1 Cooling aspects

The 4QT-prototype is equipped with two cooling systems. Firstly, the inner contour of the stator core of the SM and the complete outer housing of the 4QT are cooled by a water system. Secondly, the inner part of the coils of the DRM and the end-windings of the SM are directly cooled by forced external air, as shown in Figure 13.

![Figure 13: Illustration of the two cooling systems in the prototype with the directions of the water flow and the forced external main air flow.](image)

The maximum air inlet temperature will be 40 °C (nominal 20 °C) to take reasonable weather conditions into account. The maximum water inlet temperature is not finally defined. Concerning the mandatory cooling of the sensitive standardized power transistors and capacitors in the inverters, a water temperature above 70 °C is not recommended. The same cooling system should advantageously also be used for the housing of the electrical machines. Both machines are equipped with winding insulations, which are capable of temperature transients of up to 220 °C. This gives a maximum allowable transient temperature difference from the hot spot in the winding to the air ambient of 180 °C and to the water ambient of 150 °C, respectively. The maximum allowed average temperature of the winding systems influences the lifetime of the drive system. In order to allow for transiently high peak performance and follow common practice in high quality electrical machinery, the maximum average temperature of the winding systems should be around 130-150 °C. A general rule states that the lifetime of the insulation is reduced by 50% for every 10 K rise of temperature above the rated value at continuous operation [9]. The permanent magnets have very high coercive field strength (2865 kA/m @ 20 °C) in order to withstand demagnetization at high temperatures and safe operation of up to 180 °C is estimated.

6.2 Estimated cost of active materials

The costs of the active materials in the 4QT-prototype are estimated in Table 3. The manufacturing costs asymptotically approach the total material costs for very large production volumes [10]. In [10] a recent quote of $13/kg for Chinese permanent magnets (NdFeB) is reported. A more pessimistic value of $40/kg is chosen in this paper, also taking into account the very high coercive field strength of the permanent magnets that is needed at high temperature conditions, which increases the cost.

Table 3: Cost estimation of the active materials in the 4QT-prototype.

<table>
<thead>
<tr>
<th>Material</th>
<th>DRM</th>
<th>SM</th>
<th>4QT</th>
<th>Unit cost [$/kg]</th>
<th>Cost [$]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent magnets</td>
<td>1,3</td>
<td>0,9</td>
<td>2,2</td>
<td>40</td>
<td>88</td>
</tr>
<tr>
<td>Copper</td>
<td>3,1</td>
<td>3,3</td>
<td>6,4</td>
<td>3,0</td>
<td>19</td>
</tr>
<tr>
<td>Soft Magnetic Composites</td>
<td>9,0</td>
<td>0</td>
<td>9,0</td>
<td>1,7</td>
<td>15</td>
</tr>
<tr>
<td>Iron/Steel</td>
<td>6,3</td>
<td>12,3</td>
<td>18,6</td>
<td>1,0</td>
<td>19</td>
</tr>
<tr>
<td>Sum</td>
<td>19,7</td>
<td>16,5</td>
<td>36,2</td>
<td></td>
<td>141</td>
</tr>
</tbody>
</table>
7. Conclusions

In this paper the designs of two different compact electrical machines, intended for a new hybrid electric vehicle propulsion system named 4QT, have been presented. The speed-changing machine of the 4QT comprise of a novel machine design using a Soft Magnetic Composites core and rectangular conductors. It is shown that the new potentials offered by iron powder can result in innovative machine concepts with high performance, but the presented topology is unsuitable for machines with low pole numbers. The design of the torque-changing machine of the 4QT includes features that are unique for its power class, e.g. a segmented core with grain-oriented silicon iron teeth and concentrated windings with rectangular copper conductors. A slot fill factor of 0.74, a high fundamental winding factor for a tooth winding of 0.95, a cogging torque close to zero and high efficiency are achieved.

8. Acknowledgments

The support from ABB, for the manufacturing of the laboratory prototype and from Höganäs AB, for the support of the SMC components as well as the financial support from the Swedish Energy Agency is gratefully acknowledged.

9. References

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Measurements on Slip Ring Units for Characterization of Performance

Abstract

A slip ring unit is required in a novel hybrid electric vehicle drive system, called the Four Quadrant Transducer, for transferring AC currents to the rotating windings of an electrical machine [1]-[3]. Component compactness, cost effectiveness and reliability are of vital importance in vehicles, especially for passenger cars. Therefore, it is crucial to define the characteristics of slip ring units to be able to optimise the design for this special application with a relatively short expected lifetime compared with industrial drives. This paper describes the measured performance of three different slip ring units.

Keywords
Slip rings, brushes, measurements.

1 INTRODUCTION

This paper describes the measurements carried out on slip ring units of three different sizes. The set-up of the components, the testing methods and the results are explained and analysed. The slip rings are driven by a variable speed drive and the currents through them are fed from an inverter. The performances are monitored for various current- and speed conditions. The voltage, the current, the speed and the temperature are the measured parameters. The temperature is logged until a stable slip ring temperature has been reached and the procedure is repeated for a large number of working points.

The presented measurements differ from earlier experience of slip rings, which are limited to DC or low frequency currents and relatively low speed commonly used in rotor windings of traditional electrical machines. In the measurements presented in this paper slip rings are fed with AC currents up to a fundamental frequency of 600 Hz and driven up to a mechanical speed of 5500 rpm. The idea of the experiment is to simulate rotating armature windings exciting a permanent magnet rotor with high pole numbers.

A principle sketch of a slip ring unit is shown in figure 1.

2 PHYSICS OF SLIP RINGS

The physical, chemical and mechanical phenomena that are present in brush operation can be generally described reasonably simple. The details of what happens when two surfaces are forced together are in reality of quite complex nature. It is not within the scope of this paper to present these theoretical details, which for instance are described in [4]-[5]. A short summary of the elementary physics of sliding contacts will be given.

Figure 1: Principle sketch of a slip ring unit (1 phase).

2.1 Contact voltage drop

In sliding contacts the contact surface appears at every region of the face of the brush, but every single region is not in contact with the slip ring at every instant. At most, the contact surface represents only about 0.1 % of the apparent brush area [4]. This actual surface can be subdivided into smaller areas, depending upon the finish of the brush and the slip ring and also on the operating conditions. Many experiments have shown that there are somewhere between five and twenty separate simultaneously electrically conducting areas for each brush [4]. These spots are highly conducting bridges through the otherwise insulating film or patina formed by the process called “fritting”. The insulating film is made up of the oxides that were formed on the slip ring while it was exposed to air before it was brought into contact. When mechanical contact starts, a voltage gradient across the film generates small spots where current conduction occur. This current is sufficient to heat a small region of the film and the adjacent conducting material, so that a microscopic quantity of the slip ring metal or brush material (whichever is lower melting) is melted. The force of metallic adhesion then draws the molten metal into the form of a fine column joining the brush and the slip ring. This bridge formation is called “film breakdown” and when it happens the conduction becomes metallic. In addition, current passes also through the moisture- and oxide films by means of...
the tunnel effect, which has essentially ohmic resistance characteristics.

The total contact resistance is made up of the resistance of the arrays of conducting spots at the contact areas, the tunnel resistance and the constriction resistances in the brush and the slip ring.

2.2 Friction

By definition, the coefficient of friction is

\[ \mu = \frac{F_t}{F_n} \]  

where \( F_t \) is the tangential friction force and \( F_n \) is the force normal to the contact surface, for one brush. The friction loss is

\[ P_\mu = N_B \cdot F_t \cdot \nu \]  

where \( N_B \) is the number of brushes and \( \nu \) is the peripheral speed.

The friction coefficient of a graphite brush against a copper ring is typically 0.1-0.2 [4]. It is very dependent of temperature, atmosphere and current density. Without current or at low load conditions the molten metal process described earlier will not occur and the friction will be higher. In vacuum, no oxidation can take place. Operation in vacuum would be catastrophic for the slip ring unit. It is therefore important that the unit is run under correct atmospheric conditions.

The friction losses are of course proportional to the brush area in contact with the slip ring. A reduction of the brush cross-section or the number of parallel brushes will reduce the friction losses, but then the current density and electrical losses will increase. It is therefore possible to choose an optimum for a specific slip ring unit at a specific working point, which is the best compromise between the two loss components.

3 MEASUREMENTS

3.1 Test objects

The tested slip rings and brush holders are manufactured by BGB Engineering Ltd. of England and the copper graphite brushes by Carbex AB of Sweden. The material data are shown in table 1.

<table>
<thead>
<tr>
<th>Component</th>
<th>Material</th>
<th>Manufacturer</th>
<th>Product name</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slip ring</td>
<td>Copper</td>
<td>BGB Eng.</td>
<td></td>
</tr>
<tr>
<td>Brush holder</td>
<td>Brass</td>
<td>BGB Eng.</td>
<td>Paratec system</td>
</tr>
<tr>
<td>Brush</td>
<td>Copper graphite</td>
<td>Carbex</td>
<td>Cu35M</td>
</tr>
</tbody>
</table>

The three units have different slip ring diameters of 100, 80 and 62 mm, respectively. Each unit has three phases (or modules) and each module has three brushes in parallel. The brush size is similar for all units. Its cross-sectional area is 16×32 mm². The length of one slip ring is 20 mm, which obviously is 4 mm longer than the brush. The recommended maximum peripheral speed of the brush is 40 m/s and the recommended maximum current density is 14 A/cm². The density of the brush material is 3700 kg/m³.

Each of the three units is equipped with three PT-100 temperature elements. These are glued or fastened on the cooling element of the brush holder, between the spring bracket and the brush and on the edge of the slip ring, respectively.

The largest slip ring unit is designed for a rated current of 200 A and a speed of 3000 rpm, but shall also be able to operate at 6000 rpm for shorter periods of time. The two smaller units are not designed for a special operating condition, but are still measured to characterize the relation between size and performance.

During the experiments it was experienced that there was a risk of overheating the two smaller slip ring units if they were continuously run at high speed. Therefore it was decided that the majority of the tests should be performed up to a speed of 2000 rpm with all three slip ring units connected. For tests at speeds up to 5500 rpm, only the slip ring unit with the ring diameter 100 mm was connected with its brushes.

3.2 Laboratory set-up

The three power slip rings and one signal slip ring are mounted on two shafts, which are driven by a variable speed drive system. The system is an asynchronous ABB Integral Motor 112 with integrated inverter and 4 kW output power, which has a maximum speed of almost 6000 rpm. The shafts are coupled together with two couplings and they are supported by four bearings in addition to the bearings of the electric motor. The purpose of the signal slip ring is to transfer the information from the temperature sensors glued on the power slip rings to the data acquisition unit. The laboratory set-up is shown in figure 2.

![Figure 2: Laboratory set-up.](image-url)
fan is blocked. This in order to have equal cooling conditions for the test objects.
The power slip rings are electrically connected in series. In addition, the modules of the three-phase (3 modules) slip ring unit are also series connected. This is done by short-circuiting two slip rings. For the largest slip ring unit, only two modules are active. The voltage is measured across two modules for each slip ring unit by a digital power meter. The current is measured via shunts by the same power meter.

An inverter generates the current through the power slip rings. Inductors (30 mH) are connected in series with the power slip rings to increase the AC load and to smooth the current waveform. It is therefore easy to control the fundamental frequency as well as the current amplitude. The inverter is fed from a DC source with a voltage in the range 20-300 V depending on the load. The switching frequency is 14.6 kHz.

The inductances of the slip ring units are modelled as pure resistances, which is very close to the truth for this kind of measurements at stationary conditions.

The electrical power into the electric motor is measured. It is therefore possible to monitor power variations caused by friction variations and vibrations. The friction losses of the brushes can also be estimated pretty well by measuring the difference in losses with and without the brushes in contact with the slip rings. The instruments used during the measurements are given in table 2.

### Table 2: List of instruments.

<table>
<thead>
<tr>
<th>Instrument</th>
<th>Model</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power meter 1</td>
<td>Yokogawa WT 1030M</td>
</tr>
<tr>
<td>Power meter 2</td>
<td>Yokogawa 2533</td>
</tr>
<tr>
<td>Current shunts</td>
<td>Norma Triax (6-300 A)</td>
</tr>
<tr>
<td>Temperature instrument</td>
<td>Hewlett Packard 34970A</td>
</tr>
<tr>
<td>Temperature sensors</td>
<td>Heraeus M-FK 422 (PT-100)</td>
</tr>
<tr>
<td>Digital tachometer</td>
<td>Ono Sokki HT-430</td>
</tr>
</tbody>
</table>

#### 3.3 Test methods

A comprehensive description of test methods for the measurement of brush characteristics is published by IEC [6].

The temperature rise is in this paper defined as

\[
\Delta T = T_{\text{ring}} - T_{\text{ambient}}
\]

(3)

where \( T_{\text{ring}} \) is the slip ring temperature and \( T_{\text{ambient}} \) is the ambient temperature. The ring temperature is not necessarily the hottest spot of the slip ring unit. For very high current densities the maximum temperature is expected to be in the middle of the brush. If the friction losses are more dominating than the electrical losses, the maximum temperature will be located closer to the contact surface.

The definition of a stable temperature is that it does not change more than one degree per hour (Kelvin). This condition has not been achieved for all performed tests, but the temperatures have been reasonably stable. The reason for the discrepancy is due to the nature of brush operation, which for instance can be expressed in a temperature drop after a long time of running with increasing temperature.

Slip rings with different diameters will of course experience different peripheral speeds for the same rotational speed. The peripheral speeds for the slip ring diameters 100, 80 and 62 mm are given as functions of rotational speed in figure 4.

#### 4 RESULTS

Whenever not clearly noted, the graphs below are valid for the unit with the slip ring diameter 100 mm.

##### 4.1 Frequency

The electrical power loss per phase (3 brushes) as a function of fundamental frequency for the speed 1000 rpm is shown in figure 5.
current phenomena seem to have negligible influence. The fact that the electrical power loss increases with current is not a surprise, however.

The temperature rise is shown as a function of frequency for the speed 1000 rpm in figure 6.

The temperature rise shows the same characteristics as the electrical power loss. What is also interesting in figure 6, is that the temperature rise is lower for a current of 100 A than for 50 A. It can depend on that higher current means lower friction loss up to a certain extent. The electrical and mechanical properties in brush operation are strongly related.

### 4.2 Current density

The electrical power loss per phase as a function of DC current density is shown in figure 7.

The temperature curves in figure 9 are quite typical. The temperatures of the brush and the ring are almost equal when stability is reached. The temperature of the cooling element of the holder is naturally lower. The reason why all the temperatures in figure 9 do not start at the ambient temperature is that another working point was previously tested.

### 4.3 Slip ring diameter

The electrical power loss per phase as a function of slip ring diameter is shown in figure 10.

The instantaneous temperatures at the current density 13 A/cm$^2$ and the speed 1000 rpm are shown in figure 9.

The temperature rise as a function of DC current density.

The instantaneous temperatures at the current density 13 A/cm$^2$ and the speed 1000 rpm are shown in figure 9.

The temperature rise is shown as a function of frequency for the speed 1000 rpm in figure 6.

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4.3 Slip ring diameter

The electrical power loss per phase as a function of slip ring diameter is shown in figure 10.
The electrical power loss as a function of slip ring diameter seems to have a random pattern. Manufacture tolerances and individual differences can play a role. The brushes are all similar, so the difference in electrical power loss is not expected to be large. The temperature rise is shown in figure 11.

Figure 11: Temperature rise as a function of slip ring diameter for the current density 6.5 A/cm² (100 A, DC).

Figure 11 shows that slip rings with smaller diameters get hotter than larger diameters at increasing speed. Rings with smaller diameters have lower peripheral speed and should have lower friction losses for the same friction coefficient and contact surface. The reason is that a smaller ring has a smaller surface for heat dissipation and it is more difficult to achieve a good air stream for cooling through the narrow shaft.

4.4 Peripheral speed

The temperature rise as a function of peripheral speed for the current density 3.3 A/cm² is shown in figure 12. The electrical power loss is very low for this working point (<5 W), so the temperature rise is mainly due to the friction loss.

Figure 12: Temperature rise as a function of peripheral speed for the current density 3.3 A/cm² (50 A, 200 Hz).

The temperature rise increases dramatically with increasing peripheral speed.

4.5 Friction

Friction loss per phase is shown as a function of peripheral speed in figure 13. The friction loss seems to be proportional to the peripheral speed as illustrated by the dotted line in figure 13. That was already expected from Equation (2).

Figure 13: Friction loss per phase (3 brushes in parallel) as a function of peripheral speed.

4.6 Vibrations

A vibration phenomenon was accidentally discovered. Copper wires electrically connect the slip rings. Due to the high rotational speed at one occasion (4000 rpm), these wires that were connected between the slip rings of diameters 80 and 62 mm were bent outwards and touched the static part outside of them. This of course resulted in considerable vibrations. The brushes of the two smaller units were disconnected at that time, but the largest unit was in use. The slip ring and brush temperatures of the largest unit immediately started to increase rapidly. The test had to be stopped to prevent the unit to be overheated. The heat that was produced by the wires hitting the static part was isolated from the largest slip ring by a support bearing, so the temperature rise in it had to arise from additional friction in itself due to the vibrations. The fact that vibrations can be very harmful to brushes is also described in literature [4].

5 CONCLUSIONS

The characteristics of slip ring units have been demonstrated by measurement results from numerous working conditions. It has been shown that the electrical and mechanical properties are closely related, which means that they interfere with each other. For a specific brush and slip ring configuration at a specific working point, it is possible to find an optimum where the sum of the electrical power loss and the friction loss is at a minimum.

If slip rings are to be run at high speed, it can be advantageous to either reduce the number of brushes or decrease the apparent brush area. This will increase the electrical power loss, but the friction loss will be reduced proportionally. Slip rings with relatively small diameters and poor cooling capability are found to be particularly sensitive to high peripheral speed, but not as much sensitive to high current density.

It is experienced that big vibrations are unsuitable for brush operation. In severe cases, the slip ring unit can get overheated and destroyed due to added friction, which once again is a function of the electrical and mechanical behaviour put together. Special attention must be given
to the mechanical design of the slip ring unit if it is supposed to be used in applications with potentially high vibration risks, for instance in a hybrid electric vehicle.

ACKNOWLEDGEMENTS

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REFERENCES


Performance Evaluation of Permanent Magnet Synchronous Machines with Concentrated and Distributed Windings Including the Effect of Field-weakening

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PERFORMANCE EVALUATION OF PERMANENT MAGNET SYNCHRONOUS MACHINES WITH CONCENTRATED AND DISTRIBUTED WINDINGS including the EFFECT OF FIELD-WEAKENING

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Keywords: Permanent magnet machines, torque ripple, power capability, field-weakening.

ABSTRACT
The aim of this paper is to evaluate different concentrated fractional pitch winding designs in comparison to a distributed full pitch winding design with one slot per pole per phase as a reference. The rotor design is fixed and the permanent magnets are surface mounted, i.e. the rotor has negligible magnetic saliency. The studied parameters are power capability considering inverter capacity, field-weakening performance, torque ripple (including cogging), iron losses, winding losses and thermal behaviour. The results of the machine performance comparisons are based on a comprehensive use of finite element analysis tools. Measurement results from tests on a laboratory prototype are used to define heat coefficients.

LIST OF SYMBOLS

- **B** Flux density \([T]\)
- **B_{max}** Maximum flux density \([T]\)
- **E** Electromotive force \([V]\)
- **f** Electric frequency \([Hz]\)
- **I** Phase current \([A]\)
- **I_{lim}** Inverter limited peak phase current \([A]\)
- **i_d** Direct-axis current \([A]\)
- **i_q** Quadrature-axis current \([A]\)
- **k_e** Excess loss coefficient \([W/(T^2s^2kg)]\)
- **k_f** Stacking factor \((0 \leq k_f \leq 1)\) \([-]\)
- **k_h** Hysteresis loss coefficient \([Ws/(Tkg)]\)
- **k_{rise}** Empirical thermal coefficient \([K/W]\)
- **L_d** Direct-axis inductance \([H]\)
- **L_q** Quadrature-axis inductance \([H]\)
- **L_s** Synchronous phase inductance \([H]\)
- **n** Rotational speed \([rpm]\)
- **n_{max}** Maximum rotational speed \([rpm]\)
- **P** Power \([W]\)
- **P_{iron}** Iron losses \([W]\)
- **P_{winding}** Winding losses \([W]\)
- **p** Number of poles \([-]\)
- **p_{iron}** Iron losses per weight unit \([W/kg]\)
- **q** Number of slots per pole per phase \([-]\)
- **R** Resistance \([\Omega]\)
- **T** Torque \([Nm]\)
- **T_p** Time period \([s]\)
- **t** Time \([s]\)
- **U** Phase voltage \([V]\)
- **U_{max}** Maximum line-line voltage, RMS \([V]\)
- **w_{sheet}** Iron sheet thickness \([m]\)
- **\Delta T_{rise}** Temperature rise \([K]\)
- **\delta** Load angle \([\degree]\)
- **\phi** Power angle \([\degree]\)
- **\theta** Current angle \([\degree]\)
- **\rho** Weight density of iron \([kg/m^3]\)
- **\sigma** Conductivity of iron \([1/(\Omega m)]\)
- **\omega** Electric angular speed \([rad/s]\)
- **\Psi_m** Magnet flux linkage \([Vs]\)
- **\Psi_{ar}** Armature flux linkage \([Vs]\)

INTRODUCTION
The interest in permanent magnet (PM) machines with concentrated windings is growing due to lower winding cost through extended automated manufacturing and the feasibility of more compact designs compared to the conventional machine designs with distributed windings. This is in particular true for servo and traction applications with high production volumes.

The aim of this paper is to evaluate different concentrated fractional pitch winding designs in comparison to a distributed full pitch winding design with one slot per pole per phase as a reference. The machines in the analysis have 14 poles with equal outer rotor designs and active dimensions. The permanent magnets are surface mounted, i.e. the rotor has negligible magnetic saliency. The studied parameters are power capability considering inverter capacity, field-weakening performance, torque ripple (including cogging), iron losses, winding losses and thermal behaviour. Several authors have comprehensively studied power capabilities for different kinds of rotor configurations during the 1990’s [1]-[5]. In this paper the rotor design is fixed and the parameters are the winding type and the number of slots.

The results of the machine performance comparisons are based on a comprehensive use of finite element analysis tools (Flux2D\(^1\) and Ace\(^2\)). One of the machine designs with concentrated windings intended for a novel hybrid electric propulsion system, called the Four

---

1 Software developed by Cedrat, France.
2 Software (non-commercial) developed by ABB, Sweden.
Quadrant Transducer [6], has been manufactured as a laboratory prototype. Measurement results from tests on the prototype are used to define heat coefficients.

**MACHINE DATA**

The main data of the three studied machine designs are given in Table 1, the distributions of their phase windings are shown in Table 2 and the machine geometries are shown in Fig. 1. Machine designs with concentrated windings can advantageously be manufactured with segmented core teeth. Bobbin-wound coils can then easily be assembled with the teeth and thereafter with the core back. The coils can be made up of rectangular conductors to increase the copper slot fill factor and a value of e.g. 0.74 is possible [6]. The slots of the machines with concentrated windings (machines B and C in Fig. 1) are therefore designed with a rectangular shape. Their total stator slot areas are equal. The machine with distributed windings (machine A in Fig. 1) has a traditional stator design with constant tooth width and random-wound coils of round copper wires. The copper slot fill factor is typically 0.37 for a machine of this dimension and type (only half the value of the machines with rectangular conductors), but the total stator slot area of machine A is 44% larger than for machines B and C, since its constant tooth width results in less iron and more slot space.

Machine design A can be modelled in the finite element tools by a segment of one pole pitch, due to the magnetic symmetry conditions that are present. In the same manner design B can be modelled by two pole pitches, while design C does not have any symmetry-planes and must consequently be modelled by a complete 360° layout, as illustrated by Fig. 1.

The electric circuit parameters of the three machines, as shown in Table 1, are based on a chosen equal base speed of 2000 rpm and an inverter rating of 20 kVA. This means that the machines will have different numbers of phase turns for the same maximum voltage, since the electromotive forces (EMF) and inductances differ. This, however, enables a comparison of the machines’ power capabilities, which will be seen in the next chapter.

**Table 1: Dimensions of the three studied machine designs.**

<table>
<thead>
<tr>
<th>Machine</th>
<th>A</th>
<th>B</th>
<th>C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Active length</td>
<td>65</td>
<td>65</td>
<td>65</td>
</tr>
<tr>
<td>Stator outer diameter</td>
<td>204</td>
<td>204</td>
<td>204</td>
</tr>
<tr>
<td>Stator inner diameter</td>
<td>122</td>
<td>122</td>
<td>122</td>
</tr>
<tr>
<td>Rotor outer diameter</td>
<td>240</td>
<td>240</td>
<td>240</td>
</tr>
<tr>
<td>Air gap length</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>Magnet length</td>
<td>4.5</td>
<td>4.5</td>
<td>4.5</td>
</tr>
<tr>
<td>Slot opening width</td>
<td>2.0</td>
<td>2.0</td>
<td>2.0</td>
</tr>
<tr>
<td>Average conductor length</td>
<td>130</td>
<td>101</td>
<td>110</td>
</tr>
<tr>
<td>Average end-winding length</td>
<td>65</td>
<td>36</td>
<td>45</td>
</tr>
<tr>
<td>Area of stator slot</td>
<td>156</td>
<td>217</td>
<td>303</td>
</tr>
<tr>
<td>Magnet arc angle</td>
<td>17.1</td>
<td>17.1</td>
<td>17.1</td>
</tr>
<tr>
<td>Mass of permanent magnets</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
</tr>
<tr>
<td>Mass of copper</td>
<td>2.8</td>
<td>3.1</td>
<td>3.3</td>
</tr>
<tr>
<td>Mass of iron yoke</td>
<td>2.9</td>
<td>2.9</td>
<td>2.9</td>
</tr>
<tr>
<td>Mass of iron teeth</td>
<td>4.1</td>
<td>5.1</td>
<td>5.1</td>
</tr>
<tr>
<td>Number of poles</td>
<td>14</td>
<td>14</td>
<td>14</td>
</tr>
<tr>
<td>Number of slots</td>
<td>42</td>
<td>21</td>
<td>15</td>
</tr>
<tr>
<td>No. of slots per pole and phase</td>
<td>1 0.5</td>
<td>0.36</td>
<td></td>
</tr>
<tr>
<td>Number of turns per phase</td>
<td>58</td>
<td>57</td>
<td>50</td>
</tr>
<tr>
<td>Slot fill factor</td>
<td>0.37</td>
<td>0.74</td>
<td>0.74</td>
</tr>
<tr>
<td>Fundamental winding factor</td>
<td>1</td>
<td>0.87</td>
<td>0.95</td>
</tr>
<tr>
<td>Phase resistance at 120°C</td>
<td>50</td>
<td>27</td>
<td>22</td>
</tr>
<tr>
<td>Synchronous phase inductance</td>
<td>0.30</td>
<td>0.46</td>
<td>0.52</td>
</tr>
<tr>
<td>Base speed (inverter rating 20 kVA)</td>
<td>2000</td>
<td>2000</td>
<td>2000</td>
</tr>
<tr>
<td>Phase EMF, at base speed, RMS</td>
<td>84</td>
<td>71</td>
<td>68</td>
</tr>
</tbody>
</table>

**Table 2: The distributions of the phase windings of the three studied machines.**

<table>
<thead>
<tr>
<th>Slot #:</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
<th>11</th>
<th>12</th>
<th>13</th>
<th>14</th>
<th>15</th>
</tr>
</thead>
<tbody>
<tr>
<td>Machine A:</td>
<td>-B</td>
<td>+A</td>
<td>-C</td>
<td>+B</td>
<td>-A</td>
<td>+C</td>
<td>-B</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Machine B:</td>
<td>-A/+B</td>
<td>-B/+C</td>
<td>-C/+A</td>
<td>-A/+B</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Machine C:</td>
<td>+C/-A</td>
<td>+A/+A</td>
<td>-A/-A</td>
<td>+A/+A</td>
<td>+A/-B</td>
<td>+B/+B</td>
<td>-B/-B</td>
<td>+B/-B</td>
<td>+B/C</td>
<td>+C+C-C/-C</td>
<td>+C/+C</td>
<td>-C/-C</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Fig. 1. Geometries of the studied non-salient outer rotor PM machines with 14 poles and equal rotors: Machine A with distributed windings (q=1), machine B with concentrated double-layer windings (q=0.5) and machine C with concentrated double-layer windings (q=0.36).**
A change in the inverter rating actually means a change of the base speed, if the numbers of turns are left constant. The inverter rating is increased in the chapter about losses later in this paper in order to increase the constant power region. The theoretical base speed is therefore higher than 2000 rpm in that particular comparison. Vector diagrams for non-salient pole permanent magnet machines are shown in Fig. 2.

Fig. 2. Vector diagrams for non-salient PM machines in motoring (left) and generating (right) operations.

**POWER CAPABILITY CONSIDERING INVERTER CAPACITY**

The power capability of the different machine designs considering inverter capacity will be compared using a per unit system as defined in [3]-[5]. The normalised dq-model in the rotor reference frame has the magnet flux in the d-axis direction while the q-axis is leading it by an electrical angle of 90°. The inductances and flux linkages are modelled as constants and the machines are considered to be lossless in this section, i.e. the input power equals the output power. The effects of saturation and losses are described later. The subscript \( n \) is used to indicate normalised parameters in this paper. The base values of the electric circuit parameters are given by:

\[
\begin{align*}
U_{\text{base}} &= \frac{\sqrt{2}}{\sqrt{3}} U_{\text{max}} \\
I_{\text{base}} &= I_{\text{lim}} = \sqrt{i_d^2 + i_q^2} \\
\Psi_{\text{base}} &= \sqrt{\Psi_m + L_d i_d^2 + (L_q i_q)^2} \\
o_{\text{base}} &= \frac{U_{\text{base}}}{\Psi_{\text{base}}} \\
L_{\text{base}} &= \frac{\Psi_{\text{base}}}{I_{\text{base}}} \\
T_{\text{base}} &= 3 \cdot \frac{U_{\text{base}} \cdot I_{\text{base}}}{p \cdot o_{\text{base}}} \\
n_{\text{base}} &= \frac{60 \cdot o_{\text{base}}}{\pi \cdot p}.
\end{align*}
\]

**Field-weakening operation**

Permanent magnet machines can have a theoretical infinite constant power region, which was presented by [1]. Non-salient PM machines have finite speed if \( \Psi_m > L_d i_{\text{lim}} \) and theoretically infinite speed if \( \Psi_m \leq L_d i_{\text{lim}} \). Optimum current control strategies for maximum torque-per-ampere for the two machine types are illustrated by the synchronous flux linkage vector trajectories in Fig. 3. The operating point of the non-salient PM machine must always be within the voltage and current limiting boundaries in Fig. 3. Maximum synchronous flux linkage condition occurs at maximum torque operation, which is current limited. The constant torque region ends at base speed, because of voltage limitation, and is represented by point A. The synchronous flux linkage needs to be weakened above base speed in inverse proportion to the speed in order to keep the fundamental voltage constant, which is represented by region A-B. Finite speed machines have reached their maximum speed at point B, where the q-current and consequently the torque is zero. For infinite speed machines, point B represents an operating point where the direct axis armature flux linkage precisely counteracts the magnet flux linkage. The d-current can therefore be kept constant in region B-C for the purpose of current minimization and reducing the q-current further increases the speed. Region B-C is a true constant power region. The per unit torque at base speed (point A) and below is

\[
T_n = \Psi_{\text{max}} \sqrt{1 - L_{ds}^2}
\]

and the per unit power is

\[
P_n = n_n T_n.
\]

Fig. 3. Optimum synchronous flux linkage vector trajectory for maximum torque-per-ampere of non-salient permanent magnet machines with finite speed (upper) and infinite speed (lower) potential, respectively.
The per unit torque as a function of speed in region A-B is

\[ T_{n} = \sqrt{1 - L_{sn}^2} = \left(1 - \frac{1}{2} \frac{L_{m}}{L_{sn}} \right)^2. \]  \hspace{1cm} (10)

Finite speed non-salient PM machines have reached their maximum speed at point B, which is given by

\[ n_{\max, B} = \frac{1}{\Psi_{\max}} - L_{sn}. \]  \hspace{1cm} (11)

The per unit torque as a function of speed for infinite speed machines in region B-C is

\[ T_{n} = \frac{1}{n_{B}} \frac{\Psi_{\max}}{L_{sn}^2} = \frac{1}{n_{B}} \sqrt{1 - L_{sn}^2}. \]  \hspace{1cm} (12)

The normalised torque and power capabilities versus speed of the studied machine designs for an inverter rating of 20 kVA and a base speed of 2000 rpm are shown in Fig. 4 and Fig. 5, respectively. The machines have quite different power capabilities mainly due to the inductance differences of the machine types. The power capability is also influenced by the fundamental winding factor, which is proportional to the electromotive force and even to the torque in a surface-mounted PM-machine. A permanent magnet machine is moving towards being a theoretical infinite speed machine, when its inverter rating is increasing.

**TORQUE RIPPLE**

Torque ripple in permanent magnet machines can even be present without winding excitation and is then called cogging torque. The results from the FEM-calculated (Finite Element Method) cogging torques are shown in Fig. 6. The permanent magnet arc pitch equals two slot pitches and one slot pitch for machine A and machine B, respectively. The method of designing the length of the magnet arc pitch as an integer value in number of slot pitches is frequently used in permanent magnet machine design to minimize the magnitude of the cogging torque. Despite the implemented reduction technique and a relatively large physical air gap of 1.5 mm, it is seen in Fig. 6 that machine A has substantial cogging torque with the peak-to-peak value 10 Nm and six periods per pole pair pitch. Machine B has comparatively 1.2 Nm and the same number of periods. Machine C has completely negligible cogging. It is worth mentioning that a machine with distributed windings can be skewed by one slot pitch to cancel out the cogging torque and reduce the torque ripple. The drawback is a resulting lower fundamental winding factor (0,96 if q=1). Machines with concentrated windings are normally not skewed, since the fundamental open-circuit voltage would drastically drop. In the following torque ripple analysis it is assumed that none of the machines are skewed.

The FEM-calculated torque ripple is calculated at four current loading values of 10,3, 20,7, 41,3 and 82,7 A/mm (RMS) in addition to the no-load calculation. The results for machines A, B and C are shown in Fig. 7, Fig. 8 and Fig. 9, respectively. The windings are fed with pure sinusoidal currents with the current angle θ (see Fig. 2) equal to zero. The torque ripple is therefore caused by machine parameters only, i.e. the inverter has no influence. This means that pulsating torque variations can only depend on non-sinusoidal open-circuit voltages or air gap permeance variations. The torque ripple factor of machine A is decreasing with increasing current loading and is in the range 21-49%. The torque ripple factor of machine B is in the range of 13-22%, while it is only 3-6% for machine C. The highest torque ripple component for all machines is the 6th, i.e. its frequency is six times the fundamental electrical frequency. This component is mainly caused by the product of the fundamental current component and the 5th and 7th harmonic components in the no-load flux, but also the cogging torque contributes in machines A and B, since their cogging torques have the same frequency. The results are numerically represented in Table 3, where the average torque, the peak-peak torque and the torque
ripple factor are given. The FEM-calculated no-load voltages at a speed of 2000 rpm are shown in Fig. 10.

![Fig. 6. Cogging torque versus rotor position for the studied machines. The cogging torque of machine C is almost equal to zero.](image)

![Fig. 7. Calculated torque ripple versus rotor position of machine A for a current loading of 0, 10.3, 20.7, 41.3 and 82.7 A/mm (RMS).](image)

![Fig. 8. Calculated torque ripple versus rotor position of machine B for a current loading of 0, 10.3, 20.7, 41.3 and 82.7 A/mm (RMS).](image)

![Fig. 9. Calculated torque ripple versus rotor position for machine C for a current loading of 0, 10.3, 20.7, 41.3 and 82.7 A/mm (RMS).](image)

Table 3: FEM-calculated torque ripple for various current loadings.

<table>
<thead>
<tr>
<th>Machine</th>
<th>A</th>
<th>B</th>
<th>C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Torque</td>
<td>Tav</td>
<td>∆T</td>
<td>kTRF</td>
</tr>
<tr>
<td>0</td>
<td>10,0</td>
<td>0,0</td>
<td>0,0</td>
</tr>
<tr>
<td>10,3</td>
<td>21,9</td>
<td>10,7</td>
<td>48,9</td>
</tr>
<tr>
<td>20,7</td>
<td>43,7</td>
<td>12,8</td>
<td>29,3</td>
</tr>
<tr>
<td>41,3</td>
<td>86,4</td>
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<td>21,8</td>
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<tr>
<td>82,7</td>
<td>165,1</td>
<td>35,2</td>
<td>21,3</td>
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</table>

Fig. 10. FEM-calculated phase EMFs at 2000 rpm for the machines.

**EFFECTS OF SATURATION AND LOSSES**

The losses are calculated at base speed and in the field-weakening region. The friction losses (bearing and windage) will be neglected, since they are basically the same in all the three studied machines and do not contribute with important information to the comparative analysis. The conductors are assumed to have sufficiently small cross-section areas, so that skin and proximity effects can be disregarded. Rotor losses are neglected. The machine parameters used in this analysis are found in Table 1, but the current rating of the inverter is unlimited, so that the constant power region can be extended for any machine.

**Saturation**

The direct and quadrature-axes inductances are almost equal in surface-mounted PM machines, since the permanent magnets have a relative permeability close to unity. The inductance of machine A is almost constant versus current, while the inductances of machines B and C are higher and saturate linearly with current. The reason is a relatively high slot leakage inductance in the machines with concentrated windings. This slot leakage inductance is reduced when the current is increased, because the tooth tips get saturated. In the machine with distributed windings, several slots lie between the sides of each phase coil and the slot leakage inductance is therefore lower. The slot leakage inductance increases by decreasing number of slots (for constant slot opening widths), since more iron at the expense of air reduces the magnetic reluctance, which explains why machine C has higher inductance than machine B. The FEM-calculated inductances versus current are shown in Fig. 11 in per unit values. The saturation of the slot leakage inductance for the concentrated winding machines reduces the induced voltage at high torque operation.
The empirical coefficient $k_{\text{rise}}$ depends on the machine dimensions, the cooling method and the winding type.

$$
\Delta T_{\text{rise}} = k_{\text{rise}} \sqrt{P_{\text{winding}} (P_{\text{winding}} + P_{\text{iron}})}.
$$

(15)
rotor. This alternating field is due to a relatively high sub-harmonic component in the magnetomotive force (MMF) induced by the non-symmetrical magnetic design of machine C. The sub-harmonic component will induce rotor losses, which are current dependent. These losses can be effectively reduced by laminating the rotor and manufacture the permanent magnets in small pieces. Unfortunately, this phenomenon was underestimated at the design stage, so the rotor core of the prototype was made of solid magnetic steel. Loading the machine prototype above rated speed is therefore risking demagnetisation of the permanent magnets, due to the heat produced by the rotor losses.

The teeth in the laboratory prototype are made of grain-oriented silicon iron sheets in order to increase the magnetic performance of the machine, regarding losses and overload capability, compared to the non-oriented iron, which is used in the core back.

**Summary**

The studied machines have shown quite different characteristics in performance despite the fact that their rotors and active dimensions are equal. Simplified relative comparisons of the acquired results are shown in Table 5 below. The main reason for the differences in the power density is due the differences in slot fill factors (made possible by segmented iron teeth and rectangular shaped conductor cross-sections) and end-winding lengths (where the concentrated windings are superior over the distributed windings). The machine with q=0,36 has higher power density than the machine with q=0,5, because it has a higher fundamental winding factor, i.e. higher torque for the same current loading. It is also superior regarding torque ripple, because of the non-symmetrical magnetic design of the machine. Care should be taken to avoid parasitic effects like rotor losses, which can arise from this design.

<table>
<thead>
<tr>
<th>Machine</th>
<th>A</th>
<th>B</th>
<th>C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Speed [RPM]</td>
<td>2000</td>
<td>4000</td>
<td>6000</td>
</tr>
<tr>
<td>Winding losses [W]</td>
<td>1006</td>
<td>819</td>
<td>1777</td>
</tr>
<tr>
<td>Iron losses [W]</td>
<td>235</td>
<td>346</td>
<td>350</td>
</tr>
<tr>
<td>Total losses [W]</td>
<td>1241</td>
<td>1165</td>
<td>2127</td>
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<tr>
<td>Efficiency</td>
<td>57</td>
<td>59</td>
<td>110</td>
</tr>
<tr>
<td>Power factor below base speed</td>
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<td>+</td>
<td>O</td>
</tr>
<tr>
<td>Power factor above base speed</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Power density</td>
<td>++</td>
<td>++</td>
<td>++</td>
</tr>
<tr>
<td>Ripple torque</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Cogging torque</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Efficiency</td>
<td>O</td>
<td>+</td>
<td>++</td>
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</tbody>
</table>

**CONCLUSIONS**

The performances of surface-mounted permanent magnet machines with both distributed and concentrated windings have been evaluated in this paper. It is shown that machines with concentrated windings have considerably higher inductance, mainly due to higher slot leakage. The power capabilities are therefore dramatically different. Machines with distributed windings have the best utilization of the inverter rating at base speed. Machines with concentrated windings have a wider field-weakening region, assuming identical rotors. Furthermore, it is shown that the machine design with q=0,36 has typically 85% and 74% lower torque ripple than the machine with q=1 and q=0,5, respectively. The machines with concentrated windings have superior thermal performance in the constant power region due to their lower winding losses.

**REFERENCES**

Paper 6

Analysis of a PM Machine with Concentrated Fractional Pitch Windings

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Analysis of a PM Machine with Concentrated Fractional Pitch Windings

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Abstract—Permanent magnet machines with concentrated fractional pitch windings for servo and traction applications are growing in popularity. This paper deals with analysis of a novel synchronous permanent magnet machine using concentrated windings and grain-oriented and non-oriented silicon iron sheets for the teeth and yoke, respectively. Parasitic effects like unbalanced radial forces and alternating magnetic fields in the rotor, due to a non-symmetrical magnetic design of the presented machine, are described. Measurements on a laboratory prototype are performed and the results are compared to finite element method computation results to identify the heat transfer coefficients of the electrical machine.

Index Terms—Permanent magnet machines, concentrated windings, thermal calculations, parasitic effects, measurements.

I. INTRODUCTION

The interest in permanent magnet (PM) machines with concentrated windings is growing due to their simple winding structure, which enables cost effective automation methods in the manufacturing process [1]-[2]. This is in particular true for servo and traction applications with high production volumes. The concentrated winding machines have potentially more compact designs compared to the conventional machine designs with distributed windings, due to shorter end-windings and axial build. BLDC- (Brush-Less Direct Current) and synchronous PM machines with concentrated fractional pitch windings usually have a poor fundamental winding factor of typically 0.87. This paper deals with analysis of a novel synchronous permanent magnet machine with a fundamental winding factor of 0.95 using concentrated windings and grain-oriented and non-oriented silicon iron sheets for the teeth and yoke, respectively.

II. ELECTRICAL MACHINE DESIGN

The stator core of the presented machine is segmented with 0.30 mm grain-oriented silicon iron sheets teeth and 0.35 mm non-oriented silicon iron sheets stator back. A grain-oriented material has superior performance in respect of less saturation and lower losses in the direction of the grain orientation. The performance in the perpendicular direction is slightly poorer than for a non-oriented material, as shown in Fig. 1 and Fig. 2. The grain-oriented material is therefore used for the teeth, where the main flux is in the radial direction. Due to the segmentation of the stator core, it is possible to use pre-made coils in the slots. The coils are made up of rectangular copper wires manufactured under tension. The result is a high slot fill factor of 0.74. The proposed machine shown in Fig. 3 has 14 poles and 15 slots to achieve a high fundamental winding factor for a tooth winding of 0.95 and very low torque ripple, without skewing [3]. There are no symmetric magnetic conditions along the air gap, i.e. the machine needs to be modelled over its complete cross-section in the FEM-software. This characteristic is the reason for a resulting close to sinusoidal back-EMF (electromotive force) and extremely low cogging torque [4]. Since the machine is non-symmetrical, some parasitic effects like unbalanced radial force components and magnetic field variations in the rotor arise and are described in Section III. The machine in Fig. 3 has an outer and inner stator diameter of 204 mm and 122 mm, respectively, while the outer rotor diameter is 240 mm. The active length is 65 mm.

Fig. 1. Peak magnetic polarizations versus peak magnetic field strength for 0.30 mm grain- (upper for the oriented direction and lower for the perpendicular direction) and 0.35 mm non-oriented (middle) silicon iron sheets.

Fig. 2. Core losses versus frequency at a sinusoidal peak flux density of 1.0 T for 0.30 mm grain- (lower for the oriented direction and upper for the perpendicular direction) and 0.35 mm non-oriented (middle) silicon iron sheets tested according to the Epstein frame method.
III. PARASITIC EFFECTS

A. Magnetic noise

Tangential forces, radial forces and magnetostrictive forces in the machine air gap cause magnetic noise from a radial flux electrical machine. Tangential forces act on the stator and rotor to produce a desired torque, but also - depending on the magnetic design - undesired torque ripple components, including cogging. Radial forces are attractive forces between the stator and the rotor, while magnetostrictive forces act to stretch the iron in the direction of the magnetic field. Radial forces are by far the most dominating source of magnetic noise in radial flux induction machines [5] and PM machines. The same theory is of course also applicable for axial flux machines, but the radial forces described above are then equivalent to axial forces. Maxwell’s stress tensor equation gives the magnetic force density in the air gap

$$\sigma(\theta, t) = \frac{1}{2\mu_0} \left[ B_r(\theta, t) B_t(\theta, t) - B_t(\theta, t) B_r(\theta, t) \right],$$  \hspace{1cm} (1)

where $\theta$ is an angular coordinate, $t$ is time, $B_r$ is the radial component of the air gap flux density, $B_t$ is the tangential component of the air gap flux density and $\mu_0$ is the permeability of free space. The active iron material in a typical, moderately saturated electrical machine is many times more permeable than air. Therefore, the flux enters and leaves the stator and rotor surfaces almost perpendicularly. This means that the radial component of the air gap flux density is much larger than the tangential component, which can be neglected. The radial magnetic stress from Equation 1 can then be simplified into

$$\sigma(\theta, t) = \frac{B_t(\theta, t)^2}{2\mu_0}.$$ \hspace{1cm} (2)

The FEM-computed normal component of the no-load flux density versus the angular position for the machine presented in Fig. 3 is shown in Fig. 4. The angle zero is here defined to be at 9 o’clock in Fig. 3 and the angular position increases counter-clockwise.

The flux density due to a winding excitation for a current loading of 927 A/cm (RMS) with peak current in phase A, which here means current in the quadrature axis direction, is also shown in Fig. 4. This is an extreme overload operating condition and the armature reaction is therefore very high. The flux density mainly induced by phase B (0-120° in Fig. 4) is amplifying the no-load flux density. The flux density mainly generated by phase A (120-240°) is partly amplifying partly dampening the no-load flux density. Likewise, phase C (240-360°) is dampening the no-load flux density. The resulting radial magnetic stress in the air gap is shown in Fig. 5. It is for instance shown that the radial stress is very high at around 140°, but very low on the opposite side of the air gap at 320°. This results in an unbalanced radial force called an unbalanced magnetic pull, which represents a unidirectional deflection of the stator (and rotor) [5]. The effect is exemplified by considering the stator or rotor to be a metal ring with a mode 1 radial vibration shape in Fig. 6. Some other modes are shown in the same figure. The radial forces usually cause undesired magnetic noise. The sound level depends on several factors like rigidity of the machine parts, machine dimensions and location of the natural frequencies of the machine structures. Small machines are in general not so sensitive to higher vibration modes as big machines, but especially mode number 1 can be influential even in this case, because relatively low vibration frequencies can easily coincide with the natural frequencies of the machine structures and result in resonance. The presented machine is very sensitive to the mode 1 vibration, since it has a relatively thin and weak outer rotor structure, which will function as an undesired loudspeaker. Inner rotor designs are generally more rigid against this phenomenon, but the stator can still be influenced if the yoke is thin (as for high pole numbers).

B. Rotor losses

The flux density caused by the armature reaction in Fig. 4 has the same waveform as the magnetomotive force (MMF). Since the MMF is not symmetrical in relation to the waveform
of the no-load flux density, alternating magnetic fields appear in the rotor. The tangential component of the flux density, in the middle of the rotor iron between two permanent magnets, due to the armature reaction is shown in Fig. 7 versus time at an angular speed of 2000 rpm. The flux density at no-load is of course a pure DC component (0.85 T), but superposed AC components with substantial amplitudes are introduced for increasing MMF values. The alternating magnetic fields in the rotor may induce losses in the permanent magnets and the rotor iron. In order to minimize these losses, the permanent magnets should be manufactured in many small pieces and the rotor iron should be laminated. Unfortunately, the magnetic field variations in the rotor were underestimated at the design stage, which means that the manufactured laboratory prototype did not follow these design principles. Instead, the rotor iron is solid and as a consequence the machine cannot fulfil the specifications as given in [3]. The elapsed time in Fig. 7 equals one electrical period. The waveforms are not all starting and ending at the same flux density value. The reason is that the MMF has sub-harmonics with the lowest being the 1/7th. Another machine configuration with a high fundamental winding factor and a low cogging torque is the one with the combination 8 poles and 9 slots [6], which also has the same weaknesses as described for the presented machine. In general, it is not advantageous with concentrated three-phase windings and an odd number of slots if the number of poles almost equals the number of slots (e.g. the pole/slot-combinations: 8/9, 14/15 and 20/21). Furthermore, machines with single layer concentrated windings have normally more MMF harmonics, including sub-harmonics, than double layer concentrated windings machines and are consequently more sensitive to the discussed parasitic effects.

IV. MEASUREMENTS

A. Laboratory prototype

The studied laboratory prototype and its different active parts are shown in Fig. 8. The sheets for a tooth are glued together to form a component, which are then glued onto the stator yoke. The materials used for the active parts of the machine are given in Table I. The machine is equipped with several PT-100 temperature sensors, and some of their locations are shown in Fig. 9. The rotor temperature is also measured by using a slip ring unit for signal transmission. Due to the parasitic effects described previously in Section III, the machine is risking demagnetisation of the permanent magnets if loaded hard above rated speed. The tests with high torque are therefore performed at very low angular speeds.

B. Laboratory set-up

Two different set-ups are used. The first is used for generator tests with resistors as load, while the second is used for motor tests with an inverter as the current source. The losses are found by the direct method in both set-ups, which means that the torque, speed and electrical power are measured. The laboratory set-up for the generator measurements is shown in Fig. 10. The current is not controlled in the generator tests. It is only a function of the induced EMFs and the impedances in the electric circuits. In motor operation, the machine is fed with quadrature axis current, only.

Fig. 5. Radial magnetic stress in the air gap for the current loading 927 A/cm (RMS) with peak current in phase A.

Fig. 6. Some radial vibration mode shapes for the stator and rotor.

Fig. 7. Computed tangential component of the flux density in the rotor iron (x=114 mm and y=0 mm in Fig. 3) versus time at an angular speed of 2000 rpm for different current loadings (0, 103, 207, 413 and 827 A/cm, RMS).
Table I
Material data of the active machine parts

<table>
<thead>
<tr>
<th>Active machine part</th>
<th>Manufacturer</th>
<th>Material code</th>
</tr>
</thead>
<tbody>
<tr>
<td>Permanent magnets</td>
<td>Vacuumschmelze, Germany</td>
<td>VACODYM 688 AP</td>
</tr>
<tr>
<td>Silicon iron teeth</td>
<td>Cogent, UK</td>
<td>M105-30P</td>
</tr>
<tr>
<td>Silicon iron yoke</td>
<td>Cogent, Sweden</td>
<td>M235-35A</td>
</tr>
<tr>
<td>Rectangular copper wires</td>
<td>Dahréntråd, Sweden</td>
<td>DAMID Class 200</td>
</tr>
</tbody>
</table>

C. Measurement results
The measured phase resistance of 20.5 mΩ at room temperature is 29% higher than the analytically calculated 15.9 mΩ. The reason for the discrepancy is due to tolerances in the wire cross-section and slightly different resistivity of 17.24 nΩm compared to the analytically used 16.70 nΩm (5%) and relatively long connection wires and terminal cables (24%). The measured (thick line) and the three FEM-computed (thin lines) no-load voltages at a permanent magnet temperature of 59 °C and an angular speed of 2000 rpm, which is the base speed, are shown in Fig. 11. The curve with the highest amplitude is the originally computed EMF. The curve with the second highest amplitude takes into account that the manufactured machine has magnetic wedges on both sides for production purposes. This increases the leakage flux and the amplitude is reduced by 2%. The smallest of the computed curves takes into account additionally the tolerances of the magnetic remanence from the manufacturer of the permanent magnets. The minimum guaranteed remanence of 1.03 T is 5% lower than the typical remanence of 1.08 T. The maximum value of the measured EMF is higher than for the smallest of the computed curves, but its RMS value is 3.6% less. The reason for this is the difference in waveforms as seen in Fig. 11, which is probably caused by the difference in the modelled and the actual tooth materials. The FEM-computed EMFs are modelled with non-oriented iron in the teeth, as for the yoke, due to a numerically much simpler task. The manufactured machine has, as already mentioned, grain-oriented iron for the teeth. The lamination stacking factor is according to the manufacturer of the teeth 0.965, but was set to unity in the computations. The test results from the generating operations are shown in Table II and the test results from the motoring operations are shown in Table III. The temperature rise is defined as the difference of the measured end-winding temperature and the inlet water temperature. (The normal definition of temperature rise is the difference between the average winding temperature and the ambient temperature.) The hot spot temperature in the end-winding is normally 5-10 K higher than the average temperature of the winding.

The measured results for the generating working point 31.5 Nm at 2000 rpm, as shown in Table II, verify that the rotor has losses at high frequencies. The rotor temperature is almost 20 K higher than for the stator core. The motor tests are performed at very low speeds. The rotor losses are therefore almost negligible, which is also indirectly shown by studying the temperatures in Table III. The motor tests are used to identify the winding losses’, while the no-load generator tests are used to identify the iron losses’ influence on the heat generation in the machine. The thermal performance of the machine, assuming the machine is made without any manufacturing defects, can therefore be evaluated, which is shown in the next chapter.
V. THERMAL ANALYSIS

A. Theory of heat transfer

There are three ways that heat may be transferred between machine parts at different temperatures — by conduction, convection and radiation. Heat exchange due to conduction takes place in the solid parts of the machine, e.g. copper, steel, and insulators. Heat transfer by means of convection occurs within a temperature gradient between solid parts and cooling media, as air or water. Heat transfer due to radiation is insignificant in the studied machine, which is water-cooled, and is therefore not considered in this paper. The heat flow caused by conduction is described by the differential conduction equation [7]:

$$\frac{\partial^2 T}{\partial x^2} + \frac{\partial^2 T}{\partial y^2} + \frac{\partial^2 T}{\partial z^2} + \frac{\dot{q}'_G}{k} = \frac{1}{\alpha} \frac{\partial T}{\partial t},$$

where

$$\alpha = \frac{k}{\rho c}$$

and $\dot{q}'_G$ is the rate of heat generated per unit volume, $T$ is the temperature, $\alpha$ is the thermal diffusivity, which depends only on the physical properties of the material, $\rho$ is the material density, $c$ is the thermal capacity and $k$ represents the thermal conductivity. For a stationary problem, the right-hand side term is zero. Heat transfer by convection occurs at different temperatures between solid parts of the electrical machine and a fluid or an air coolant. Convection can be classified as forced, caused by a fan or a pump, and natural, due to mass density gradients, caused by temperature differences. Newton’s Law of Cooling determines the rate of heat transfer by convection [7]:

$$\dot{q}_c = h_c A (T_s - T_\infty),$$

where $q_c$ is the rate of heat transfer by convection, $T_s$ and $T_\infty$ are the surface temperature and the temperature of the fluid, respectively, and $h_c$ is an average heat transfer coefficient over area $A$. A quantitative evaluation of the convection heat transfer coefficient is a complex problem, as $h_c$ is dependent on many variables such as geometry of the surface, flow characteristics, velocity and physical properties of a fluid, and temperature difference ($T_s - T_\infty$). The value of the convection heat transfer coefficient can also vary along the cooling surface. However, for most engineering applications the average value $\bar{h}_c$ is often used instead. Several methods are available for the evaluation of convection heat transfer coefficients. Dimensional analysis combined with experiments is a mathematically simple method and has found a wide range of applications [7]. Dimensional analysis combines several variables to form a number of dimensionless groups. The main disadvantage of the method is that the results obtained from the calculations are incomplete.

---

**TABLE II**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Unit</th>
<th>Value</th>
</tr>
</thead>
<tbody>
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<td>Speed</td>
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<td>Mechanical power</td>
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<td>Yoke</td>
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<td>Inlet water</td>
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**TABLE III**

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<tr>
<td>Speed</td>
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<td>Mechanical power</td>
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<td>Fund. phase voltage (RMS)</td>
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<tr>
<td>Phase current (RMS)</td>
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<td>Electrical power</td>
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<tr>
<td>Losses</td>
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<tr>
<td>Water flow</td>
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<td>End-winding</td>
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<tr>
<td>Tooth</td>
<td>[°C]</td>
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</tr>
<tr>
<td>Yoke</td>
<td>[°C]</td>
<td>22</td>
</tr>
<tr>
<td>Rotor iron</td>
<td>[°C]</td>
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</tr>
<tr>
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<td>[°C]</td>
<td>18</td>
</tr>
<tr>
<td>Inlet water</td>
<td>[°C]</td>
<td>22</td>
</tr>
<tr>
<td>Temperature rise</td>
<td>[K]</td>
<td>9</td>
</tr>
</tbody>
</table>

---

Fig. 11. The measured (thick curve) and FEM-computed no-load voltages at a speed of 2000 rpm and at a permanent magnet temperature of 59 °C.
and quite useless without experimental verification. Moreover, the number of variables that influence the physical process of heat transfer should be properly selected beforehand. However, once the pertinent variables are known, dimensional analysis can be applied to most problems. The Buckingham \( \pi \) theorem is used to determine dimensionless groups [7]. The most important dimensionless groups are outlined below.

The Reynolds number is a dimensionless ratio of inertia to viscous forces defined as

\[
\text{Re} = \frac{U_\infty L}{\nu},
\]

where \( U_\infty \) is the velocity of the fluid, \( L \) is a characteristic length, which for a tube is the diameter, and \( \nu \) is the kinematic viscosity. The magnitude of the Reynolds number characterizes the flow of the cooling fluid, i.e. a low Reynolds number corresponds to laminar flow, while a high Reynolds number indicates turbulent flow. The transition from laminar to turbulent flow happens at approximately \( \text{Re} \approx 2300 \) [8]. The heat transfer coefficients are much higher in turbulent than in laminar flow (around 30-80 times higher) [8]. The Nusselt number is a ratio of the heat entering the fluid from the surface to the heat conducted away by the fluid. The Nusselt number is defined as

\[
\text{Nu} = \frac{\dot{h}_s L}{k}.
\]

The Prandtl number is a ratio of the viscous boundary layer thickness to the thermal boundary layer thickness. The Prandtl number is defined as

\[
\text{Pr} = \frac{\nu}{\alpha} = \frac{c_p \mu_b}{k},
\]

where \( \mu_b \) is the dynamic viscosity of the fluid at the average or bulk temperature. The Grashof number is a ratio of buoyancy to viscous forces. The Grashof number is defined as

\[
\text{Gr} = \frac{g \beta (T_s - T_\infty) L^3}{\nu^2},
\]

where \( g \) is acceleration due to gravity, and \( \beta \) the volumetric thermal expansion coefficient. The Grashof number characterizes the flow in natural convection and indicates whether it is laminar or turbulent. The Graetz number is defined as

\[
\text{Gz} = \frac{\pi}{4} \text{Re} \cdot \text{Pr} \left( \frac{L}{l} \right),
\]

where \( l \) here is the distance from the entrance of the cooling-channel. The Graetz number is used to define heat transfer coefficients in laminar tube flow. The temperature of a fluid flowing inside a tube will change along the tube. It is therefore not evident how to define the temperature difference to use in Equation 5. If the tube surface temperature varies significantly, the logarithmic mean temperature difference is used to calculate the rate of heat flow

\[
q_e = \dot{m} C_p \left( \frac{\Delta T_\text{out} - \Delta T_\text{in}}{\ln(\Delta T_\text{out} / \Delta T_\text{in})} \right),
\]

where \( \Delta T_\text{in} \) and \( \Delta T_\text{out} \) are the temperature differences at the inlet and outlet, respectively. For turbulent flow in short tubes \((10 < l_{\text{tube}} / L < 400)\) the Nusselt number is given by the empirical relation [9]

\[
\text{Nu} = 0.036 \cdot \text{Re}^{0.8} \cdot \text{Pr}^{1/3} \left( \frac{L}{l_{\text{tube}}} \right)^{1/18},
\]

where \( l_{\text{tube}} \) is the length of the tube. For laminar flow, when the temperature field is fully developed, the Nusselt number is constant, which happens for \( \text{Gz} < 10 \). For the entrance region of a tube in laminar flow, the Nusselt number is

\[
\text{Nu} = 1.86 \cdot \text{Gz}^{1/3} \left( \frac{\mu_s}{\mu_b} \right)^{0.14},
\]

where \( \mu_s \) is the dynamic viscosity of the fluid at the surface temperature. With known Nusselt number, the heat transfer coefficient can easily be calculated from the definition in Equation 7. The velocity profile of a fluid at an entrance region of a tube, fully developed laminar flow and fully developed turbulent flow are shown in Fig. 12.

B. Finite element method simulations

The machine is thermally analysed using FEM. The stator is mounted along its inner periphery on an aluminium cylinder, which is cooled by forced water flowing in series axially through eight cooling ducts, see Fig. 8. The modelling of the cooling performance is done in two steps. The first step is to model the mechanical fluid flow and heat transfer process taking place inside of the cooling ducts and in the aluminium cylinder. Only one duct is modelled, since the other ducts experience almost similar thermal conditions.

![Fig. 12. Velocity profile of a fluid at an entrance region of a tube (a), at fully developed laminar flow (b) and at fully developed turbulent flow (c).](image-url)
The heat is transferred by conduction inside the solid cylinder, while convection occurs inside the ducts. The heat flux, produced by the heated parts of the machine, is assumed to be homogeneously distributed along the inner surface of the stator. The aim of the first step is to obtain the temperature boundary condition on the surface of the cooling cylinder. Due to axial symmetry, only half of the cooling duct in the aluminium frame is modelled, as illustrated in Fig. 13. Mechanical fluid flow appears in cooling duct (2). The velocity profile is depicted with arrows. Fluid flow is coupled with the heat transfer problem to obtain the temperature distribution across the upper surface (1), which is in physical contact with the inner surface of the stator yoke.

The second step is to model the axial cross-section of the rest of the machine, in which all the heat is transferred by means of heat conduction. It is assumed that the heat from the heated parts of the machine is transferred mostly in the radial direction. Due to thermal symmetry, only 1/15-part of the machine is modelled. The thermal conductivities for the different machine parts are shown in Table IV.

The measurement results from the previous chapter are used to define the boundary conditions in the model. The thermal performance of the machine is simulated, disregarding some imperfections with the manufactured laboratory prototype. The following assumptions are made:

- The rotor losses are neglected,
- The DC-resistance is used to calculate the winding losses (AC-resistance is neglected),
- The contribution to the resistance from the terminal cables is neglected,
- The originally computed electromotive force is used instead of the measured value,
- The teeth are assumed to be of the same non-oriented material as the yoke.

The simulated motoring operation working points are shown in Table V. The average temperature of the winding is assumed to be 5 K less than the end-winding temperature, which is regarded as a hot spot temperature.

### C. Simulation results

The temperature distributions for the four different simulated working points are shown in Fig. 14. The highest temperature is located in the conductor part for all four simulated cases. The highest temperature in Fig. 14 from top is 139 °C (412 K), 157 °C (430 K), 86 °C (359 K) and 98 °C (371 K), respectively. The inlet water temperature is 22 °C in all simulations. The machine seems to be capable of operating continuously for all four simulated conditions, without being overheated. The winding insulation is able to withstand heat shock treatments up to 220 °C [10] and the permanent magnets are estimated to withstand temperatures up to 180 °C during normal working conditions. However, the lifetime of the winding insulation is very much dependent on the operating temperature.

Industrial applications have an expected high utilization and a desired long lifetime. The maximum winding temperature is therefore preferably 145 °C under such operation. With a maximum ambient temperature of 40 °C, the allowed temperature rise is 105 K. If these constraints are used in the simulations at base speed (2000 rpm), the base torque becomes 82 Nm. I.e. the motor has a torque density of 5.0 Nm/kg based on the weight of the active materials.

### Table IV

<table>
<thead>
<tr>
<th>Machine part</th>
<th>Material</th>
<th>Thermal conductivity W/(m·K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductors</td>
<td>Copper</td>
<td>401</td>
</tr>
<tr>
<td>Cooling cylinder</td>
<td>Aluminium</td>
<td>255</td>
</tr>
<tr>
<td>Core</td>
<td>Iron</td>
<td>40</td>
</tr>
<tr>
<td>Permanent magnet</td>
<td>Neodymium-Iron-Boron</td>
<td>9</td>
</tr>
<tr>
<td>Cooling ducts</td>
<td>Water (20°C)</td>
<td>0.58</td>
</tr>
<tr>
<td>Conductor insulation</td>
<td>Varnish (Isonol 51)</td>
<td>0.44</td>
</tr>
<tr>
<td>Slot insulation</td>
<td>Nomex</td>
<td>0.143</td>
</tr>
<tr>
<td>Airgap</td>
<td>Air (70°C)</td>
<td>0.0314</td>
</tr>
</tbody>
</table>

### Table V

<table>
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<tr>
<td>1</td>
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<tr>
<td>6000</td>
<td>32</td>
<td>20</td>
<td>255</td>
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VI. CONCLUSIONS

This paper has analysed a novel PM machine with concentrated fractional pitch windings. The machine has very high performance regarding low torque ripple and high torque density, but it also has some parasitic effects in the form of unbalanced radial forces and alternating magnetic fields in the rotor. The rotor losses can be reduced taking precautions in the early design stage, but the machine is likely to be noisy if the machine structures are not mechanically very rigid and is therefore found to be an undesired weakness with the design. Concentrated three-phase winding machines with an odd number of slots and a number of poles almost equal to the number of slots are sensitive to the discussed parasitic effects (e.g. the pole-slot combinations: 8/9, 14/15 and 20/21). Measurements on a laboratory prototype have been performed and the results compared with finite element method computation results. The thermal performance of the machine has been evaluated.

ACKNOWLEDGMENTS

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REFERENCES

Testing of Silver-, Copper- and Electro-Graphite Brush Materials for Slip Ring Units

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Testing of Silver-, Copper- and Electro-Graphite Brush Materials for Slip Ring Units

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Abstract—In a novel hybrid propulsion system, the Four Quadrant Transducer (4QT), a slip ring unit is required to transfer alternating current to a rotating three-phase winding. This is an automotive application with high demands on cost effectiveness and reliability and the slip ring unit is one of the potentially weak links in the system that has to be investigated carefully. This paper investigates the impact on the performance of a slip ring unit equipped with different brush materials. The investigation focuses on the losses and the temperature rises in the slip ring unit.

I. INTRODUCTION

This paper describes the measurements carried out on slip ring units equipped with brushes of different material compounds. The set-up of the components, the testing methods and the results are explained and analysed. The slip rings are driven by a variable speed drive and the currents through them are fed from a rotating three-phase converter. The performances are monitored for various current and speed conditions. The voltage, the current, the speed, the torque, the temperature and the total distance for the brushes are the measured parameters. The temperatures are logged until a stable slip ring temperature has been reached and the procedure is repeated for a large number of operating points.

The current source is a 50 Hz rotating generator which gives currents up to 200 A. The measurements are limited to 50 Hz in frequency since the frequency has low impact on the slip ring performance according to [1]. The current level is tested in steps from a low value up to about three times the rated value for the slip ring unit. The speed of the slip rings is varied in steps from 0-5000 rpm. During these conditions three different brush qualities are tested.

A slip ring unit is required in the novel hybrid propulsion system called the Four Quadrant Transducer (4QT) [2]-[5] to supply a rotating winding with a three-phase current.

The 4QT system is shown in Fig. 1. The 4QT is mounted between the internal combustion engine (ICE) and the final gear of a hybrid vehicle. The purpose is to keep the operation of the ICE at maximum efficiency during varying driving conditions. The inner rotor in Fig. 1. is equipped with a slip ring unit.

II. LOSSES OF SLIP RINGS

The two dominating losses of a slip ring unit are the friction and electrical losses. This paper focuses on these two losses in the investigation. The electrical losses consist of the resistive losses in the brushes and in the slip ring and the loss from the contact voltage drop between the slip ring and the brush.

The total friction losses are dependent of the brush area, number of brushes, friction coefficient, spring force and the speed of the slip ring. Both the friction coefficient and the voltage drop are dependent of the complex nature of two surfaces forced together. This process is described in [6]-[7] and occurs when the slip ring has the right operating temperature and enough current is passing trough it. When this process starts it lowers the contact voltage drop and the friction coefficient.

The friction and electrical losses are also dependent of the brush material. In this paper three different materials of brushes shown in Table I have been investigated. To optimize a brush for a certain slip ring unit and operation, considerations of both the friction and resistive losses have to be done. This is done by choosing the appropriate brush material and the optimal brush area for the specific slip ring and operation.
III. MEASUREMENTS

A. Laboratory set-up

A principle sketch of the laboratory test bench is shown in Fig. 3. The power slip ring and one signal slip ring are mounted on the same shaft, which is driven by a variable speed ABB integral motor 112 with an integrated inverter and 4 kW mechanical output power. Between the drive motor and the slip ring unit a torque meter was mounted to measure the friction of the slip ring. In addition to the bearings of the electric motor, the shaft is supported by two bearings on each side of the slip ring unit. The purpose of the signal slip ring is to transfer the information from the temperature sensors glued on the power slip rings to the data acquisition unit.

The power slip rings are fed from a three-phase rotating converter. The power slip rings are Y-connected as shown in Fig. 3. This means that the AC source feeds a short circuit only with the resistance of the slip ring unit and the cables. This requires current control of the AC source.

The impedances of the slip ring units are modeled as pure resistances. The friction losses of the brushes can also be estimated pretty well by measuring the difference in losses with and without the brushes in contact with the slip rings. The instruments used during the measurements are given in Table II.

B. Test objects

BGB Engineering Ltd. of England has manufactured the tested slip ring and brush holder, while Carbex AB of Sweden has manufactured the three different brushes. A principle sketch of the slip ring unit with three brushes per phase is shown in Fig. 4. The measurements described in Section IV.A are performed with three brushes per phase, but only one brush per phase in Section IV.B in order to increase the current density capability for test purposes.

The brushes are manufactured with powder technology. This gives the opportunity to change the content of the mixture as desired. The brushes are usually a mixture between graphite and a metal. Typically, more metal content gives a lower resistivity, but higher friction, meanwhile more graphite content gives lower friction, but higher resistivity.

The brush cross-sectional area is $16 \times 32$ mm$^2$. The width of one slip ring is 20 mm, which is 4 mm longer than the brush, while the slip ring diameter is 100 mm.

Each of the slip ring units is equipped with PT-100 temperature elements. These are glued or attached to the cooling element of the brush holder, between the spring bracket and the brush and on the edge of the slip ring, respectively. Because of the rotating slip ring the PT-100 element on the slip ring has to be measured through another slip ring made for signal transmission mounted on the same shaft.

C. Test methods

A comprehensive description of test methods for the measurement of brush characteristics is published by IEC [8].

The temperature rise is in this paper defined as

$$\Delta T = T_{ring} - T_{ambient}$$ \hspace{1cm} (1)

where $T_{ring}$ is the slip ring temperature and $T_{ambient}$ is the ambient temperature. The ring temperature is not necessarily the hottest spot of the slip ring unit. For very high current densities the maximum temperature is expected to be in the middle of the brush. If the friction losses are more dominating than the electrical losses, the maximum temperature will be located closer to the contact surface.

The definition of a stable temperature is that it does not change more than one degree Kelvin per hour. This condition has not been achieved for all performed tests, but the temperatures have been reasonably stable. The reason for the discrepancy is due to the complex nature of brush
performance, which for instance can be expressed in a temperature drop after extended operation with increasing temperature.

D. Measurements

To measure the friction power of the slip ring a torque meter is mounted between the drive motor and the slip ring. To avoid the influence of friction from the bearings (support and motor) and the signal slip ring, the torque of the system without the power brushes is first measured. This value is then subtracted from the measured value of the friction torque of the power slip ring.

The temperatures of the slip ring, brushes and the holder are measured with PT-100 sensors and then saved to a computer with a Hewlett Packard logging instrument.

The current through and the voltage over the slip ring is measured continuously throughout the test.

The ambient temperature and air humidity is measured with a separate meter.

To measure the distance of slip ring that passed the brush a normal bicycle computer is used. A small sensor is attached to the rotating shaft and the radius of the slip ring is fed into the bicycle computer. The computer then records the number of revolutions and calculates the distance covered by the brushes.

IV. Results

The measured results show the losses and temperature rises of the different brush materials during different conditions. It is also shown how the speed and current affect the operation of the slip ring unit.

A. Current density and speed

If only looking at the current density and speed for the brush material copper-graphite, it can be seen that the total power losses shown in Fig. 5 is dependent on both the rotational speed of the slip ring and the current density in the brushes.

The friction losses from measurements with copper-graphite are shown in Fig. 6. In Fig. 6 an almost linear dependency of the speed can be seen for the friction losses. According to Fig. 6 the friction losses have a deviation from the linear behavior between 2000 and 3000 rpm, it is best shown for the low current densities. This could be due to a chemical process that occurs in the sliding contact between the brush and the slip ring. This process takes place in the contact surface which is only 0.1% [6] of the total brush area at every moment in time. These contacting areas are highly conductive bridges through an insulating film of patina. This film comes from oxides that were formed on the slip ring when it was exposed to air before it was brought into contact. When this chemical process begins it lowers the friction and the contact voltage drops between the brush and the slip ring.

The electrical losses shown in Fig. 7 are not significantly dependent on the rotational speed of the slip ring. For the lowest current density the electrical losses are almost constant, but a slight increase in power losses at higher speeds can be observed, mostly due to the higher temperature that implies a higher resistivity of the materials. Looking more carefully at Fig. 7, it shows a drop in the losses at the second measurement point (2000 rpm), due to a reduced voltage drop over the slip ring. The temperature rise in the copper-graphite brush is shown in Fig. 8. If comparing Fig. 5 and Fig. 8 one can, as expected, see a correlation between the temperature rise and the total losses.

<table>
<thead>
<tr>
<th>Component</th>
<th>Manufacturer</th>
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<tbody>
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<td>Temperature logging instrument</td>
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</tr>
<tr>
<td>Temperature sensors</td>
<td>Heraeus M-FK 422 (PT-100)</td>
</tr>
<tr>
<td>Torque meter</td>
<td>ONO SOKKI SS-200</td>
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<td>Torque meter decoder</td>
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</tr>
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<td>Distance meter</td>
<td>TARGA Sigma Sport</td>
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<td>Current and voltage meter</td>
<td>Fluke 123 Scope Meter</td>
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<tr>
<td>Current shunt</td>
<td>Prova CM-05</td>
</tr>
</tbody>
</table>

TABLE II

INSTRUMENTS USED DURING THE MEASUREMENTS

![Fig. 5: Total losses in the slip ring unit equipped with copper-graphite brushes.](image1)

![Fig. 6: Friction losses in the slip ring unit equipped with copper-graphite brushes.](image2)
electrical losses meanwhile Brush 2 (Cu-C) has the lowest friction losses. But when looking at Fig. 9 it can be seen that Brush 2 (Cu-C) has the lowest total losses. The contact voltage drop for the different brushes are shown in Fig. 12, there it can be seen that Brush 1 (Ag-C) has the lowest voltage drop, followed by Brush 2 (Cu-C) and Brush 3 (C) with the highest contact voltage drop. The voltage drop for the three brushes increases with the speed except for the Brush 2 (Cu-C) at the highest speed. This does not seem to be totally realistic, but due to test procedures it can’t be verified with a second test. This voltage drop also has impacts on the electrical losses and then on the total losses. One way to confirm this could be to look at the brush temperature during this test and notice the correlation between the brush temperature and the total losses. The brush temperature rises are shown in Fig. 13 for the case with 13 A/cm² and speeds from 1000 to 5000 rpm. There it can be seen that the temperature drops at the second point at 3000 rpm for Brush 1 (Ag-C) and Brush 2 (Cu-C) even though the total loss is increased.

2) Current dependency

The losses are not only dependent on the speed of the slip rings, the losses are of course highly dependent on the current through the slip rings. The slip rings have been tested with the different brushes at different current levels. The electrical losses at 3000 rpm and at different current densities in the brushes are shown in Fig. 14. There it can be seen that the electrical losses increases almost linearly with the current density. In Fig. 15 the friction losses are shown. The friction losses of electro-graphite brushes show an inverse dependence on the current density, while the silver- and copper-graphite brushes are almost constant. The total losses are shown in Fig. 16. The increase of electrical losses is larger than the decrease of friction losses. If looking at the temperature change at different current densities as shown in Fig. 17, it is clear that the temperatures are increased in the brushes with increased current density.

Fig. 9: Total losses at 13 A/cm² for different brush materials.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)
Fig. 10: Electrical losses at 13 A/cm² for different brush materials.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)

Fig. 11: Friction losses at 13 A/cm² for different brush materials.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)

Fig. 12: Contact voltage drop at 13 A/cm² for different brush materials.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)

Fig. 13: Temperature rise for the brushes at 13 A/cm².
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)

Fig. 14: Electrical losses at 3000 rpm and different current densities.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)

Fig. 15: Friction losses at 3000 rpm and different current densities.
(1: Silver-graphite, 2: Copper-graphite and 3: Electro-graphite.)
V. CONCLUSION

In this paper it has been shown that the copper-graphite brush material is the best choice regarding total losses and temperature rises in the brushes for the considered application. A strong dependency is shown between the current density and the temperature rise due to the increased losses.

The correlation between the temperature rise and the speed of the slip ring is smaller than between the temperature and the current density. The silver-graphite brush has the lowest electrical losses of the three different brush materials, mostly due to lower resistivity.

The electro-graphite brush material performed inefficiently for this type of operation.

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Iron Losses in Salient Permanent Magnet Machines at Field-weakening Operation


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Iron Losses in Salient Permanent Magnet Machines at Field-weakening Operation

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Abstract—Permanent magnet machines for electric vehicles are often designed with a magnetic saliency to improve the torque generation as a result of the reluctance torque at field-weakening operation. The armature reaction is pronounced in such operation, leading to extensive magnetic field harmonics in the teeth and consequently iron losses. This paper analyses the iron losses in two designed machine topologies, inset and interior permanent magnet rotor structures, by measurements and finite element method computations. The measurements and calculations are performed on three prototypes - one with an inset design and two with an equal interior design, but different laminated stator core sheet thickness of 0,35 and 0,20 mm, respectively. The three machines have the same stator designs and water-cooled housings.

Index Terms—Permanent magnet machines, iron losses, inset, interior, field-weakening, electric vehicles.

I. INTRODUCTION

The magnetic field in the stator core of synchronous permanent magnet (PM) machines are often assumed to be sinusoidal. The iron losses can therefore be found by simple analytical calculations. In normal electric machine operation this is a simplified, but still a relatively good estimation. In field-weakening operation this estimation is inadequate, because the magnetic field is no longer sinusoidal due to a high armature reaction. Especially the teeth will experience fast and irregular field variations leading to increased iron losses. This paper describes the cause of the stray losses, where the losses are generally located, which loss components that dominate and methods to reduce these iron losses. The paper presents analytical models based on a literature survey. The analytical models are compared with finite element method (FEM) computations and measurements on three electric machines designed for vehicle propulsion - one inset PM machine and two interior PM machines.

II. SALIENT PERMANENT MAGNET MACHINE DESIGNS

A. Specifications

The studied PM machines are designed for a battery-powered electric vehicle with a single-step gearbox. The lack of a normal multi-stage gearbox means that the electric machines must have a wide field-weakening range in order to give acceptable acceleration performance and top speed to the vehicle for a limited output power of 50 kW. The machine specification in motoring operation is shown in Fig. 1 (the generating operation is simpler). The maximum speed to base speed ratio is four, setting high demands on the machine performance. The nominal battery voltage is 300 V. The DC-voltage is dependent on the power output and the state of charge (SOC) level. The minimum DC-voltage for which the performance specification in Fig. 1 applies is 200 V. All the FEM-computations and measurements described in this paper have been performed with AC-voltages corresponding to a DC-voltage of 200 V, if not clearly indicated otherwise.

B. Magnetic layouts

Two magnetic machine layouts are designed – one with an inset design and the other with an interior design as shown in Fig. 2. Both designs have eight poles and the stator designs are equal with 48 slots and two slots per pole per phase with full pitch windings. The fundamental electrical frequency at the maximum speed of 10000 rpm is 667 Hz. One manufactured prototype is based on the inset PM machine design with a laminated iron sheet thickness of 0,35 mm. Two manufactured prototypes are based on the interior PM machine design with different sheet thickness of 0,35 mm and 0,20 mm, respectively. These three prototypes are analysed in this paper. The machines with 0,35 mm sheet thickness share the same unique stator and housing, i.e. only their individual rotors differ. The inset machine design has an airgap length of 1,2 mm, while it is 0,6 mm for the interior design. The q-to-d-inductance ratio is around 1,7 for the inset machine, while it is around 3,4 for the interior design. The interior design therefore has a substantial reluctance torque component at field-weakening operation. The active machine dimensions and the masses of the different stator parts are shown in Table I. The two machine designs have approximately equal friction losses. The rotor of the inset PM machine is bandaged, which means that the physical airgap has roughly the same length and the rotor surface has roughly the same structure as for the interior PM machines. The bearings used in the study are of identical type. The rotor cores are laminated and the permanent magnets divided into several pieces even axially, i.e. the rotor losses are negligible. The interior permanent magnet machine shown in Fig. 2 is one part of an integrated (with an inverter) electric drive system specially designed for battery-powered electric vehicles. The drive system, which is developed by ABB for an...
automotive manufacturer, and a zero emission vehicle (ZEV) equipped with the same drive system are shown in Fig. 3. This vehicle has a maximum speed of 145 km/h (90 miles/h) with its fixed gear ratio.

![Graph showing machine specifications in motoring mode showing maximum torque, continuous torque, max. power and continuous power versus speed.]

**Fig. 1.** The machine specifications in motoring mode showing maximum torque, continuous torque, max. power and continuous power versus speed.

![Cross-sectional views of one pole of (a) the inset PM machine and (b) the interior PM machine.]

**Fig. 2.** The cross-sectional views of one pole of (a) the inset PM machine and (b) the interior PM machine.

**TABLE I**

<table>
<thead>
<tr>
<th>Design</th>
<th>Inset</th>
<th>Interior</th>
</tr>
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<tbody>
<tr>
<td>Outer stator diameter</td>
<td>232</td>
<td>232</td>
</tr>
<tr>
<td>Inner stator diameter</td>
<td>160</td>
<td>160</td>
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<tr>
<td>Active length</td>
<td>125</td>
<td>125</td>
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<td>Air gap length</td>
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<td>Mass of stator regions [kg]</td>
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<td></td>
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<tr>
<td>Teeth, region I</td>
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<td>0.4</td>
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<tr>
<td>Teeth, region II</td>
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<tr>
<td>Teeth, region III</td>
<td>1.6</td>
<td>1.6</td>
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<tr>
<td>Yoke, region IV</td>
<td>7.6</td>
<td>7.6</td>
</tr>
<tr>
<td>Total stator iron, Σ</td>
<td>14.6</td>
<td>14.6</td>
</tr>
</tbody>
</table>

**III. ANALYTICAL CALCULATIONS**

**A. Modelling of iron losses**

The iron loss model includes the hysteresis, eddy-current and excess loss components [1]. For a sinusoidal magnetic field in sheet geometry, the iron losses per mass unit can be expressed as

\[
p_{\text{iron}} = k_f k_h B_m^2 f + k_f \frac{\sigma \pi^2 d^2}{6 \rho} B_m^2 f^2 + k_e B_m^{3/2} f^{3/2},
\]

where \( k_f \) is the lamination stacking factor, \( k_h \) is the hysteresis material constant, \( B_m \) is the maximum flux density, \( f \) is the electrical frequency, \( \sigma \) is the conductivity of the iron material, \( \rho \) is the density of the iron material, \( d \) is the lamination thickness or gauge and \( k_e \) is the excess material constant. If the magnetic field is not sinusoidal, (1) must be refined. Several authors have made modified models [1]-[6]. The analytical models in (1) and in the references assume that skin effect in the iron sheet is negligible. This is true for ordinary frequencies and gauges, but this assumption is incorrect for higher frequencies [7]. The skin effect causes the flux density to change across the thickness of the sheet. The iron losses will therefore have a frequency dependent behaviour, which is not according to the theoretical models.

The frequency limit for which (1) becomes inaccurate depends on the gauge and the permeability, which varies with the induction level. The frequency limit is extended for decreasing gauge. The models are valid when \( \gamma << 1 \), which is defined by [8]

\[
\gamma = \sqrt{\frac{\pi \sigma \mu f \cdot d}{\delta_p}},
\]

where \( \mu \) is the permeability of the iron material and \( \delta_p \) is the penetration depth. When the penetration depth becomes smaller than the gauge, the eddy-current loss model must be refined. The following expression is proposed by Bertotti for the eddy-current loss component in (1) when \( \gamma >> 1 \), [8]

\[
p_{\text{iron}} = k_f \frac{\pi^2 d \sqrt{\sigma B_m^2 f^2}}{2 \rho \sqrt{\mu}}.
\]

The model is only valid under linear magnetic conditions. The frequency dependence in the two models differs. The eddy-current loss component in (3) is larger than in (1) at first, but at increasing frequencies a breaking point is reached and it becomes vice versa. Since the penetration depth is often close to the gauge in the studied application, none of the analytical models above are completely correct. As can be understood by the discussion above, the modelling of iron losses is quite complex and therefore the most reliable analytical method of today is based on empirical equations from measurements.
performed on similar machine types. An alternative analytical approach that is based on the flux density waveforms for iron loss estimation can be found in [13]. This model can be applied under any specified load condition, including the field-weakening operation region, but it does not consider skin effect.

B. Material properties

The material properties used in this study are defined in Table II. The material constants are found by a curve fitting approach on the material data from the manufacturer of the iron sheets based on measurements according to the Epstein frame method [10]. The material used in the inset and interior PM machines with a sheet thickness of 0.35 mm is the M235-35A. In the interior PM machine with a sheet thickness of 0.20 mm, NO20 is used. Both materials are non-oriented silicon iron sheets from the manufacturer Cogent Surahammars Bruk.

C. Field-weakening performance

The airgap torque versus the current angle for a phase current of 200 A (RMS) is shown in Fig. 4. It is clearly shown that the interior PM machine has a substantial reluctance torque as mentioned earlier. This is an advantage as lower current loading is required for a given torque. The copper losses of the interior PM machines are therefore normally lower than for the inset PM machine. Furthermore, the current rating of the inverter can be reduced compared with the inset machine design, which is dimensioned by the peak torque at base speed requirement.

IV. FINITE ELEMENT COMPUTATIONS

A. Introduction

The iron losses are calculated using the finite element method with time stepping, hence the actual flux density variation over one electrical period can be found. The iron losses per mass unit are then calculated by [11], [12]

\[
p_{\text{iron}} = k_0 B_0^2 f + \frac{1}{T_p} \int k_1 \left[ \frac{\sigma d^2}{\rho} \left( \frac{dB(t)}{dt} \right)^2 + k_2 \left( \frac{dB(t)}{dt} \right)^{3/2} \right] dt. \tag{4}
\]

The iron loss components are individually computed in four different stator core regions. The regions are divided into tooth tips, middle region of teeth, bottom of teeth and stator back, as illustrated in Fig. 5. The separation of the core into these regions makes it possible to find the iron losses in the different parts. The BH-curves for M235-35A and NO20 are almost similar and the curve for the first is used in the computations. The FEM-computations are performed with sinusoidal phase currents, i.e. the influence of the inverter and possible current harmonics is not taken into account. Equation (1) is a special case of (4) for a sinusoidal magnetic field and (4) is also only valid when the skin effect is negligible. The computations are therefore not assumed to give absolutely reliable iron loss figures in this particular study. An ideal numerical method needs to take the flux density variation across the thickness of the iron sheet into account. Despite its weakness, which of course can be improved for high frequency applications, FEM is still a very useful tool for characterization of the iron losses.

B. Computational results

The results of the FEM-computed flux densities in the tooth and yoke at several working points for the inset and interior PM machines are shown in Fig. 6, Fig. 7, Fig. 8, Fig. 9 and Fig. 10. The flux distributions in the inset and interior PM machines at 32 kW motoring operation and the speed 10000 rpm are shown in Fig. 11. The flux density harmonics in the tooth and in the yoke for the same working point are shown in Fig. 12 and Fig. 13, respectively. The FEM-computed iron losses for the three studied machines are given in Table III, Table IV and Table V.

<table>
<thead>
<tr>
<th>Material</th>
<th>Gauge, d [mm]</th>
<th>(k_0) [W/(T^2kg)]</th>
<th>(\sigma) [1/(Ωm)]</th>
<th>(\rho) [kg/m^3]</th>
<th>(k_1) [W/(T^3/2s^1/2kg)]</th>
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<tr>
<td>Cogent Power Ltd.</td>
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Fig. 4. The torque versus current angle for a phase current of 200 A (RMS): (a) Inset machine; (b) Interior machines.

Fig. 5. The stator core is divided in four regions according to the illustration.
C. Computational results

The results of the FEM-computed flux densities in the tooth and yoke at several working points for the inset and interior PM machines are shown in Fig. 6, Fig. 7, Fig. 8, Fig. 9 and Fig. 10. The flux distributions in the inset and interior PM machines at 32 kW motoring operation and the speed 10000 rpm are shown in Fig. 11. The flux density harmonics in the tooth and in the yoke for the same working point are shown in Fig. 12 and Fig. 13, respectively. The FEM-computed iron losses for the three studied machines are given in Table III, Table IV and Table V.

Fig. 6. The FEM-computed normal component of the flux density in the tooth (point A in Fig. 5) versus rotor position at 32 kW motoring operation for the (a) inset machine and (b) interior machines.

Fig. 7. The FEM-computed tangential component of the flux density in the yoke (point B in Fig. 5) versus rotor position at 32 kW motoring operation for the (a) inset machine and (b) interior machines.

Fig. 8. The FEM-computed normal component of the flux density in the tooth (point A in Fig. 5) versus time at a speed of 10000 rpm for (a) the inset machine and (b) the interior machines.

Fig. 9. The FEM-computed BH-curve during one electrical cycle at 32 kW motoring operation and a speed of 10000 rpm, in the middle of the tooth, for (a) the inset machine and (b) the interior machines, respectively.

Fig. 10. The FEM-computed tangential component of the flux density in the yoke (point B in Fig. 5) versus time at a speed of 10000 rpm for (a) the inset machine and (b) the interior machines, respectively.

Fig. 11. Flux distribution in the (a) inset and (b) interior machines at 32 kW motoring operation and the speed 10000 rpm.

The harmonic content of the flux density in the tooth at nominal motoring operation is drastically increasing for increasing speed, especially for the interior PM machines, as illustrated in the figures. The maximum value of the tooth flux density at nominal power and maximum speed for the interior machine is 1.21 teslas (which is even higher than the maximum value of the no-load flux density) with very fast rise times, as shown in Fig. 8. This is not what is commonly assumed for field-weakening operation, namely that the magnitude of the flux density in the overall core is reduced. This assumption is only valid for the yoke, as shown in Fig. 7. The waveform of the flux density in the yoke is preserved for increasing speed (field-weakening) and the magnitude is at the same time reduced. Only a small 3rd harmonic is present, as shown in Fig. 13. This implies that the simplified analytical model in (1) might give satisfactory accuracy for calculating the iron losses in the yoke.
The interior machines have substantial 3rd, 5th, 7th, and 9th tooth flux density harmonics, as shown in Fig. 12, and they are 72%, 67%, 59% and 24% of the fundamental component of 0.51 teslas, respectively. The same harmonics for the inset machine are 51%, 59%, 18% and 8% of the fundamental component of 0.57 teslas. The frequency of these flux density harmonics is 2000, 3333, 4667 and 6000 hertz. The calculated teeth iron losses for different flux density harmonics at tooth flux density 0.0, 0.1, 0.2, 0.3, 0.4, 0.5 teslas, respectively. The same harmonics for the inset machine have therefore much lower iron losses at high speed for the airgap in the region of the permanent magnets. The inset machine is more homogenously distributed, because of the larger effective lines in the teeth for the inset machine (shown in Fig. 11) are irregular magnetic field variations at high speeds. The flux lines in the teeth for the inset machine (shown in Fig. 11) are more homogenously distributed, because of the larger effective airgap in the region of the permanent magnets. The inset machine has therefore much lower iron losses at high speed for equal gauge compared with the interior machine.

### Table III

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### Table VI

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</tbody>
</table>
V. MEASUREMENTS

A. Experimental set-up and methodology

The layout of the experimental set-up is shown in Fig. 14. A power meter measures the electrical power and a torque transducer measures the mechanical power. The core and the windings are equipped with PT-100 sensors and their temperatures are monitored during the tests. The total motor losses are found by subtracting the mechanical output power from the electrical input power. The friction losses are estimated by no-load measurements on a similarly sized machine with an unmagnetized PM rotor. The copper losses are found by DC-resistance tests and scaled to give correct resistance values by the help of the winding temperature sensors. The iron losses are finally found from the total losses minus the friction and the copper losses. All the stray losses are therefore assumed to be located in the stator core. A scheme of the field-weakening control system is shown in Fig. 15.

It is well accepted that iron losses in electric machines with laminated cores are higher than results from Epstein tests, because punching produces unrelieved stress in the laminated sheets, burrs on the edges form inter-laminar contacts and the insulation between sheets is also less than perfect [13]. It is therefore expected that the measured iron losses are higher than for the analytically calculated and for the FEM-computed. In addition, the losses generated by the PWM-modulation are not considered in the analytical calculations and FEM-computations.

B. Test results

The FEM-computed and measured iron losses at the output power 32 kW in the interval 5000-10000 rpm are shown in Fig. 16. The measured losses are higher than the FEM-computed iron losses and the difference is smaller at high speed. This might be due to the loss separation method, which is estimated to have better accuracy for increasing speed. The FEM-computed iron losses are between 22-25% lower than the measured losses at the output power 32 kW and the maximum speed of 10000 rpm.

C. The effect of different PWM-modulation strategies

Different strategies for the PWM-modulation have very high impact on the iron losses in PM machines operating at extreme field-weakening conditions. For the same DC input voltage to the inverter, the fundamental line-to-line AC voltages on the machine terminals differ substantially depending on the chosen control method. Two strategies have been evaluated. The first

### Table VI

<table>
<thead>
<tr>
<th>Harmonic order</th>
<th>Iron losses [W]</th>
</tr>
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<td>Test</td>
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<td>Interior (0.35 mm)</td>
<td>785</td>
</tr>
<tr>
<td>Interior (0.20 mm)</td>
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<tr>
<td>Measured, inset</td>
<td>38</td>
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<tr>
<td>Measured, interior (0.35 mm)</td>
<td>219</td>
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<td>Measured, interior (0.20 mm)</td>
<td>220</td>
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</tbody>
</table>

Fig. 14. The layout of the experimental set-up.

Fig. 15. Scheme of the field-weakening control system.

Fig. 16. FEM-computed and measured iron losses at the output power 32 kW for the inset machine, interior machine with 0.35 mm sheets and interior machine with 0.20 mm sheets.
approach is to use six-step switching or so-called over-modulation in the upper constant power region. The second approach is to use sinusoidal PWM modulation all the time with so-called limited or linear modulation. The fundamental RMS line-to-line voltage using six-step switching is [14]

\[ U_{RMS(1)} = \frac{\sqrt{6}}{\pi} \cdot U_{DC} = 0.78 \cdot U_{DC}, \]

where \( U_{DC} \) is the DC voltage. The fundamental RMS line-to-line voltage using sinusoidal PWM is [14]

\[ U_{RMS(1)} = \frac{\sqrt{3}}{\sqrt{2}} \cdot M \cdot U_{DC} = 0.61 \cdot M \cdot U_{DC}, \]

where \( M \) is the modulation index, which is \( M \leq 1 \) for linear sinusoidal modulation. Measurements of the machine losses at nominal power and 10000 rpm for DC input voltages in the range 150-300 V are performed using two different inverters. The first inverter with a switching frequency of 10 kHz is used for the overmodulation evaluation, while the second inverter with a switching frequency of 8 kHz is used for the limited modulation evaluation. Measurements show the fundamental RMS line-to-line voltage is 77% and 55% of the DC-voltage for the overmodulation and sinusoidal modulation methods, respectively. This means that the fundamental voltage from the inverter using six-step switching is 40% higher than for the sinusoidal inverter in the tests. Measurements show that the measured fundamental line-to-line voltage is 77% and 55% of the DC-voltage for the overmodulation and sinusoidal modulation methods, respectively.

VI. DISCUSSIONS

A. Analytical, FEM and test result comparisons

A comparison of analytically calculated, FEM-computed and measured results of the total iron losses of the three studied electric machines at 32 kW and 10000 rpm is shown in Fig. 18. The analytical calculations are based on (1) and the maximum flux density values in the teeth and yoke (shown in Fig. 8 and Fig. 10). This is a popular method in an early design stage, when the waveforms of the flux density in the teeth and the yoke for whatever reason are not known in detail. The sinusoidal field assumption is obviously not correct at field-weakening operation, but the results of the analytical calculations are nevertheless given in Fig. 18 to give an estimate of the method’s accuracy, which as seen is very poor. The analytically calculated iron losses are between 73-83% lower than the measured iron losses. The FEM-computed iron losses show much better agreement, as mentioned earlier, and are between 22-25% lower than the measured losses.

B. Stray loss reduction methods

One method to reduce the stray iron (and copper) losses in an electrical machine at field-weakening operation is to equip the drive system with a DC-DC-converter between the battery and the inverter. The mission of the DC-DC-converter is to stabilize the DC-voltage at a constant level independent of the magnitude of the load. The machine can therefore utilize the full voltage potential at all times and the field-weakening operation is simpler, resulting in an increased power capability. A higher voltage level also means that thinner cables are required, which can be very beneficial in a vehicle for saving space, weight and cost. The car manufacturer Toyota has adopted this method for the drive system in its second-generation hybrid electric vehicle, Prius, which was introduced in 2003 [15]. Another method for loss reduction in extreme field-weakening operation is to use overmodulation or six-step modulation. As demonstrated in the previous section, this reduces both the copper and the iron losses in the machine. Not only that, the method also minimizes the inverter losses, since the switching frequency becomes lower. The efficiency in the total drive system is therefore optimized.

Fig. 17. Measured losses versus DC-voltage for the interior PM machine with 0.35 mm sheets at 32 kW and 10000 rpm.

Fig. 18. Comparison of analytically calculated (sinusoidal magnetic field assumption), FEM-computed and measured results of the total iron losses in the three synchronous permanent magnet machines at 32 kW and 10000 rpm.
The use of thinner lamination sheets is a very effective way of reducing stray load iron losses, but it also results in a higher cost in both the purchase of the silicon iron material as well as in the machine production (increased punching numbers and more wear on tooling). Another way is to reduce the flux harmonics in the teeth, which can be achieved by increasing the airgap length and by reducing the slot opening width or to introduce magnetic wedges. This method causes poorer torque capability below base speed, so it should be incorporated in the early design stage.

VII. CONCLUSIONS

The iron loss behaviour in salient synchronous PM machines at field-weakening operation has been presented. A comparative study of the results from analytical calculations, finite element method computations and practical measurements has been made. It is shown that the stray losses in field-weakening operation are mainly caused by fast and irregular magnetic field variations in the stator teeth. It is demonstrated that it is very advantageous to implement a six-step PWM-modulation method in field-weakening operation to optimize the efficiency of the drive system.

REFERENCES


Analysis of a PM Machine with Soft Magnetic Composites Core

Analysis of a PM Machine with Soft Magnetic Composites Core

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Department of Electrical Engineering
KTH – Royal Institute of Technology
Stockholm, Sweden

Abstract—Soft Magnetic Composites or iron powder offers an interesting alternative to the traditional laminated silicon iron sheets as core material in electrical machines. This paper deals with analysis of a novel synchronous permanent magnet machine topology using iron powder as the core material. Measurements on a laboratory prototype are performed and the results compared with analytical calculation and finite element method computation results. The machine is found to have a high torque density of 5.1 Nm/kg. The typically high power factor of the presented machine concept saves cost on the required inverter. The machine is most suitable in low-speed applications, since only high pole numbers are practically feasible.

Keywords—Soft Magnetic Composites; PM machines; finite element analysis; thermal calculations; measurements.

I. INTRODUCTION

Soft Magnetic Composites (SMC, i.e. iron powder) offer an interesting alternative to the traditional laminated silicon iron sheets as core material in electrical machines. Iron powder has lower permeability and higher hysteresis losses than laminated iron, so a direct comparison between the two using a 2D machine design would not be fair [1]. However, SMC has equal magnetic properties in three dimensions and exploiting this unique advantage in the machine design can result in innovative concepts with high performance. Iron powder can also be pressed into complex parts and allow the cores and coils to be fitted closely together, which can result in cost effective production methods with low material waste and improved heat transfer capabilities. A novel machine concept using SMC for the core analysed in this paper was earlier presented in [2] and is an evolution of the machine that was first described in [3]. Measurements on a laboratory prototype are performed and the results compared with analytical calculation and finite element method (FEM) computation results.

II. ELECTRICAL MACHINE WITH SMC CORE

A. Machine design

The studied machine is a synchronous permanent magnet (PM) machine with both axial and radial airgap flux. The core consists of 48 segments made of the iron powder SOMALOY™ 500 with 0.5% of the lubricant Kenolube from the manufacturer Höganäs AB. The coils are made of rectangular shaped conductors, which has resulted in a very high slot fill factor of 0.78. The novel PM machine is shown in Fig. 1. The winding is a so-called Gramme-winding, which is wound around the yoke and is a full-pitch winding. Since the yoke is shorter than the active axial length and even the axial surfaces are active in the torque production, this design has short end-windings. The length of the end-windings is approximately 20% of the active length in the presented pancake design. The machine has 16 poles. The topology is not suitable for low pole numbers (<10), since the yoke area is getting larger with decreasing pole numbers, i.e. smaller inner diameter and longer end-windings. The machine is one of two machines designed for a novel hybrid electric propulsion system called the Four Quadrant Transducer (4QT) [2]. The core and coils of Fig. 1.c are rotating and so are the permanent magnets in Fig. 1.a and Fig. 1.b. The machine is therefore cooled by forced air flowing axially through the machine and the six holes, shown in Fig. 1.d. The housing is water-cooled, but that is mainly for cooling the second machine, which is not described in this paper (see [2],[4]).

![Fig. 1. a) A radial cross-sectional cut of one pole of the PM machine. b) An axial cross-sectional cut of one pole of the PM machine. c) The assembled coils and core segments shown in perspective. d) An axial cross-sectional cut of the laboratory prototype including the water-cooled housing.](image-url)
B. Material properties

The B-H-curve for the core material is shown in Fig. 2 and the losses per weight unit versus frequency is shown in Fig. 3. The relative permeability of SMC is much lower than in a silicon iron laminated core material as shown in Fig. 2. SMC is consequently not suitable in induction machines, but permanent magnet machines are not so sensitive to this parameter, even if it of course is a drawback. The losses in a SMC core are also quite high compared to silicon iron, which is illustrated by Fig. 3 and Fig. 4. The losses at a frequency of 400 hertz and an induction of 1,5 teslas are for instance three times higher for SMC than for a 0,35 mm sheet laminated iron core measured according to the Epstein frame method [5]. It is well accepted that iron losses in machines with laminated cores are higher than results from Epstein tests. The reasons for additional iron losses due to the manufacturing process are that punching produces unrelieved strains, burs on the edges form contacts between sheets and the insulations between the sheets are not perfect [1]. However, the expected increase due to production imperfections is less than 20% [6], so a laminated core has still a great advantage. The permanent magnets used are the rare-earth material NdFeB with a nominal remanence of 1,08 teslas at 20 °C.

Fig. 3. Core loss versus frequency for the iron powder material at the sinusoidal peak flux densities of 0,5 T, 1,0 T and 1,5 T, respectively (from bottom to top).

III. ELECTROMAGNETIC CALCULATIONS

A. Torque and machine parameters

The maximum average electromagnetic torque of a non-salient radial flux synchronous PM machine can be expressed by

\[ T_{em} = 2\pi R^2 L B_{rad} K_1 = R A_{rad} B_{00} K_1, \]  

(1)

where \( R \) is the airgap radius, \( L \) is the length of the rotor, \( A_{rad} \) is the radial airgap area, \( B_{00} \) is the fundamental air gap flux density and \( K_1 \) is the fundamental current loading (RMS-values). The maximum average electromagnetic torque of a non-salient single-sided axial flux machine is given by

\[ T_{em} = \pi R \left( R_e^2 - R_i^2 \right) B_{ axial} K_1 = R_e A_{ax} B_{o1} K_1, \]  

(2)

where \( R_e \) and \( R_i \) is the outer and inner radii, respectively, \( A_{ax} \) is the axial airgap area and \( K_{o1} \) is the fundamental current loading at the outer radius. The outer radius and the radius of the radial airgap are approximately the same for the presented SMC-machine (\( R_o \approx R \)). This means that the maximum average electromagnetic torque of the SMC-machine, with one radial airgap and two axial airgaps is given by

\[ T_{em} = 2\pi R \left[ R L + \left( R^2 - R_i^2 \right) \right] B_{01} K_1 = R \left[ A_{rad} + 2 A_{ax} \right] B_{o1} K_1. \]  

(3)

The torque production advantage of the SMC-machine compared with a radial or an axial flux machine is accordingly due to its larger airgap area. This statement cannot be given without a discussion, since some questionable assumptions are made. The magnetic and the electric loadings must be equal for the machines. This is an overestimation for the presented machine, since cores based on iron laminations have better magnetic properties, as described above. The electric loading of the SMC-machine will therefore also be punished, since more iron is required, leaving less space for copper. This is not completely true, because the segmented SMC-core enables a
very high slot fill factor and a very high thermal conductivity between the copper and the core, which means that the SMC-machine may actually have almost the same electric loading as a conventionally designed machine. Undoubtedly, the high magnetic loading is a problem, since the iron losses are much higher in SMC-cores. The radial airgap radius, the active length, the outer and inner radii of the axial airgaps of the presented machine is 105, 66, 104 and 73 mm, respectively. The calculated line-to-line no-load voltage at 1000 rpm is 48 volts (at 120 °C). The phase resistance and inductance are both very low of 10 mΩ (at 120 °C) and 8,3 mH, respectively. The low inductance results in a power factor close to unity. This means that a relatively small inverter is required, which is cost saving for the drive system. The finite element method computed electromagnetic torque versus the current loading at the radial airgap is given in Fig. 5. The electromagnetic FEM-computations are performed using Flux2D [7].

Different torque ripple reduction techniques are applied in the electrical machine design. The permanent magnet arc of the radial part is equal to two slot pitches. The permanent magnets for the radial part are furthermore divided in three and each magnet is stepwise skewed by 1/3 slot pitch in the laboratory prototype (not implemented in the FEM-model). The permanent magnets for the axial part has an arc length of 2,5 slot pitches at their outer radii, while they have an arc length of 1,5 slot pitches at their inner radii, i.e. they have a shape with a built-in continuous skew of one slot pitch.

B. Losses

1) Core losses

The teeth iron losses are calculated by

\[ P_{\text{m,SMC}} = (k_{\lambda,SMC} f B_{m,tool}^2 + k_{\text{eddy,SMC}} f^2 B_{m,tool}^2) m_{\text{tool}}, \quad (4) \]

and the yoke iron losses are calculated by

\[ P_{\text{y,SMC}} = (k_{\lambda,SMC} f B_{m,yoke}^2 + k_{\text{eddy,SMC}} f^2 B_{m,yoke}^2) m_{\text{yoke}}, \quad (5) \]

where \( k_{\lambda,SMC} \) (12,5⋅10^{-2} Ws/T²/kg) is the hysteresis coefficient and \( k_{\text{eddy,SMC}} \) (2⋅10^{-5} Ws²/T²/kg) is the eddy current coefficient of the Soft Magnetic Composites material, \( f \) is the electrical frequency, \( B_m \) is the maximum flux density and \( m \) is the mass. The iron coefficients are found by a curve-fitting approach of the material data from the manufacturer [8]. The mass of the teeth is 6,1 kg, while the mass of the yoke is 2,9 kg. The armature reaction is small due to the high number of poles, as seen in Fig. 9. The maximum no-load flux density and the maximum flux density at load in the different core regions are therefore almost of the same magnitudes. The maximum flux density is consequently assumed constant independently of load and sinusoidal variation is also assumed. This assumption is roughly valid, since the machine is not field-weakened.

2) Winding losses

The copper losses in a three-phase winding fed with direct current is given by

\[ P_{\text{copper,DC}} = 3 R_{\text{DC}} I^2 \quad (6) \]

where \( R_{\text{DC}} \) is the DC-resistance for one phase winding and \( I \) is the phase current (RMS). When the current frequency is high and/or the conductor cross-section is large, as in the presented machine, equation (6) underestimates the real winding losses, since eddy-current losses are induced in the copper due to skin and proximity effects. The resistance in a conductor carrying alternating current must be compensated by an AC-coefficient, \( k_{AC} \), that takes the additional losses into account. The resulting copper losses are given by

\[ P_{\text{copper}} = 3 k_{AC} R_{\text{DC}} I^2. \quad (7) \]

The calculation of the AC-coefficient for a rectangular conductor, which is frequency dependent, is found in [2].
3) Mechanical losses

The mechanical losses consist of windage losses created by air friction and bearing losses created by friction between balls, raceways and lubricant in the bearings. The windage losses increase cubically and the bearing losses increase linearly with the speed, respectively [9]. The speed of the presented machine is relatively low, which means that the mechanical losses are almost negligible. The losses are estimated based on measurements on a similarly sized motor. The total mechanical losses are estimated to be 45 watts and 150 watts at a speed of 2000 rpm and 4000 rpm, respectively.

4) Stray losses

The stray losses are additional losses caused by distorted or non-sinusoidal flux density variations in the stator core, AC losses in the winding due to stray alternating magnetic fields (i.e. from permanent magnets), permanent magnet losses due to large permeance variations or high armature reactions, and rotor back losses for the same reasons. These losses are simplistically estimated to be given by

\[ P_{\text{stray}} = \frac{1}{4} \left( P_{\text{copper}} + P_{\text{back,SMC}} + P_{\text{back,SMC}} \right). \]  \hspace{1cm} (8)

5) Total losses

The total losses are the sum of the loss components described above, as given by

\[ P_{\text{total}} = P_{\text{copper}} + P_{\text{back,SMC}} + P_{\text{back,SMC}} + P_{\text{mechanical}} + P_{\text{stray}}. \]  \hspace{1cm} (9)

The losses in the PM rotor are assumed to be negligible, because the rotor mainly experiences a magnetic DC-field.

6) Efficiency

The machine efficiency is calculated as (generator mode)

\[ \eta = \frac{\omega_m T_{\text{shaft}} - P_{\text{total}}}{\omega_m T_{\text{shaft}}} \cdot 100\%, \]  \hspace{1cm} (10)

where \( \omega_m \) is the mechanical angular speed and \( T_{\text{shaft}} \) is the shaft torque. The efficiency map of the machine based on the calculated losses is given in Fig. 7.

C. Flux densities

The flux density in the different machine regions are shown in Fig. 8 for the working point 128 Nm. The bottom picture in Fig. 8 shows that the intersection between the teeth and the core back is saturated. This region of the machine is a weak part of the magnetic design. It is possible to increase the area of this intersection by decreasing the slot depth or the slot width, but at the same time the available space for the copper winding is also decreased.

IV. THERMAL CALCULATIONS

The machine is cooled by means of forced air convection. The thermal analysis of it is performed with the use of the finite element software FEMLAB [10]. The analysis is done in two steps. In the first step the velocity distribution in the region where the forced air convection occurs is obtained.
Fig. 9. The FEM-computed (a) no-load flux density and (b) flux density at 128 Nm in the middle of the radial airgap showing the radial component (x-direction for the global coordinate system).

Fig. 10. The definition of the global coordinate system in polar cylindrical coordinates.
The velocity distribution is calculated by solving the Incompressible-Navier-Stokes equations as given by [11]-[12]

\[
\begin{align*}
\frac{\partial \mathbf{u}}{\partial t} + \mathbf{u} \cdot \nabla \mathbf{u} &= -\frac{1}{\rho} \nabla p + \nu \nabla^2 \mathbf{u} \\
\nabla \cdot \mathbf{u} &= 0
\end{align*}
\]  

(11)

where \( \mathbf{u} \) is the fluid velocity vector, \( \nu \) is the kinematic viscosity, \( p \) is the pressure and \( \rho \) is the mass density, where the first line represents the conservation of momentum, while the second line represents the conservation of mass. In the second step the temperature distribution is obtained by applying the conservation of energy equation as given by [13]

\[
\rho c_p \frac{\partial T}{\partial t} + \nabla \cdot \left( -k \nabla T + \rho c_v T \mathbf{u} \right) = q_c ,
\]

(12)

where \( c_p \) is the thermal capacity of the fluid at constant pressure and \( T \) is the temperature. The expression within the brackets is the heat flux vector and \( q_c \) is the heat source. The mechanical fluid flow is assumed not to affect the heat transfer, since the temperature gradient is insignificant. Only one slot pitch of the machine is modelled due to thermal symmetry and the analysed model is shown in Fig. 12.

Some typical thermal material properties of electrical machine parts are shown in TABLE I. The results from the computations in Fig. 13 are showing the temperature distributions in the model for different power losses of 442, 769 and 1060 watts, respectively. The losses are taken from no-load measurements as described in the next section, and the thermal coefficients used in the computations are calibrated against measurement results. The thermal contact between the coil and yoke is assumed to be perfect, which means that the loss distribution has small influence on the temperature rise. The core and the winding in the laboratory prototype is vacuum pressure impregnated (VPI), which ensures a high conductivity between these parts. The calculated average temperature in the core versus the volume flow rate of forced air convection at a machine power loss of 1060 watts is shown in Fig. 14. The value of the air velocity of 13 m/s on the inlet that corresponds to approximately 8.7 m\(^3\)/min (with an inlet temperature of 25\(^\circ\)C), gave a good agreement between FEM-calculations and measured temperatures [14]. The calculated average temperatures of the core at steady-state are 69, 101 and 129\(^\circ\)C, respectively, for the three loss conditions.

V. MEASUREMENTS

The measurement set-up is shown in Fig. 15. A torque transducer measures the mechanical power and a power meter coupled to an oscilloscope measures the voltages. A fan of 185 watts provides the forced air-convection. The water-cooling system is inactive during these tests. The measurement results from the no-load tests are given in TABLE II.

<table>
<thead>
<tr>
<th>Machine part</th>
<th>Material</th>
<th>( \rho ) [kg/m(^3)]</th>
<th>( c_p ) [J/(kg K)]</th>
<th>( k ) [W/(m K)]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductors</td>
<td>Copper</td>
<td>8900</td>
<td>390</td>
<td>401</td>
</tr>
<tr>
<td>Cooling cylinder</td>
<td>Aluminium</td>
<td>2702</td>
<td>896</td>
<td>255</td>
</tr>
<tr>
<td>Core</td>
<td>Iron/steel</td>
<td>7600</td>
<td>745</td>
<td>40</td>
</tr>
<tr>
<td>Core</td>
<td>SMC</td>
<td>7400</td>
<td>450</td>
<td>20</td>
</tr>
<tr>
<td>Permanent magnet</td>
<td>NdFeB</td>
<td>7800</td>
<td>440</td>
<td>9</td>
</tr>
<tr>
<td>Cooling ducts</td>
<td>Water (20(^\circ)C)</td>
<td>998</td>
<td>4190</td>
<td>0.58</td>
</tr>
<tr>
<td>Conductor insulation</td>
<td>Varnish (Isilon 51)</td>
<td>1100</td>
<td>1000</td>
<td>0.44</td>
</tr>
<tr>
<td>Slot insulation</td>
<td>Nomex</td>
<td>950</td>
<td>1170</td>
<td>0.143</td>
</tr>
<tr>
<td>Airgap</td>
<td>Air (40(^\circ)C)</td>
<td>1.09</td>
<td>1014</td>
<td>0.0265</td>
</tr>
</tbody>
</table>
Fig. 13. FEM-computed temperature distribution in one tooth and coil segment for the power loss (a) 442 watts, (b) 769 watts and (c) 1060 watts (in kelvins).

Fig. 14. FEM-calculated average temperature in the core versus the volume flow rate of forced air (of 25°C inlet temperature) at a machine power loss of 1060 W.

Fig. 15. Measurement set-up for the laboratory prototype, which is cooled by forced air from a fan. The built-in water-cooling system is not utilized in these measurements.

Unfortunately, the winding is manufactured with a fault in one of the phases (discovered after the VPI-process), which creates circulating currents and extra losses. The fault is illustrated by the measured unsymmetrical no-load phase back-EMFs in Fig. 16. This means that the machine is only tested at no-load generating condition, in order to avoid further damage. The electromagnetic performance of the machine is therefore not verified. However, the thermal performance of the machine is verified. The good thermal contact between the core and the winding makes the temperature rise in the different parts almost independent of the loss distribution, as explained earlier. This means that the maximum power loss for a chosen maximum temperature rise can be defined. The maximum power loss is found to be 1060 watts for a temperature rise of 105 kelvins. It is now possible to estimate the rated performance of the machine based on the combined measured and calculated data. The machine is therefore estimated to be able to operate continuously in the region as defined in Fig. 17. The rated speed is assumed to be 2000 rpm, where the rated torque is 106 Nm. This gives a torque density in generating operation of 5,4 Nm/kg (5,1 Nm/kg in motoring operation), which is in class with water-cooled high performance concentrated winding machines with silicon iron laminated cores [4].

TABLE II. Measurement results at thermal steady-state from no-load tests.

<table>
<thead>
<tr>
<th>Speed [rpm]</th>
<th>1005</th>
<th>1499</th>
<th>1903</th>
</tr>
</thead>
<tbody>
<tr>
<td>Torque [Nm]</td>
<td>5,1</td>
<td>5,9</td>
<td>6,4</td>
</tr>
<tr>
<td>Power losses [W]</td>
<td>442</td>
<td>769</td>
<td>1060</td>
</tr>
<tr>
<td>Phase voltage, A, RMS [V]</td>
<td>22,5</td>
<td>33,8</td>
<td>43,1</td>
</tr>
<tr>
<td>Phase voltage, B, RMS [V]</td>
<td>10,1</td>
<td>16,3</td>
<td>21,9</td>
</tr>
<tr>
<td>Phase voltage, C, RMS [V]</td>
<td>22,8</td>
<td>34</td>
<td>41,5</td>
</tr>
</tbody>
</table>
| Temperatures
| Permanent magnets [°C] | 47  | 63  | 77  |
| Soft Magnetic Composites [°C] | 70  | 102 | 130 |
| Copper conductors [°C] | 71  | 100 | 126 |
| Ambient, air inlet [°C] | 25  | 26  | 25  |
| Air outlet [°C] | 35  | 42  | 51  |
| Temperature rise [K] | 45  | 76  | 105 |

Fig. 16. The measured no-load phase back-EMFs. (The winding is manufactured with a fault in one of the phases, which creates circulating currents and extra losses.)

Fig. 17. The estimated continuously operating region of the machine is illustrated by the shaded area (generating mode).
VI. LABORATORY PROTOTYPE

Some pictures illustrating the manufacturing process of the laboratory prototype are shown in Fig. 18 and the assembly of the machine is shown in Fig. 19. The bobbin-wound copper coils are almost completely filling up the slot space, especially along the slot width of 4.4 mm, which is filled up by a copper conductor width of 4.0 mm. Each slot consist of 11 conductors with a height of 1.06 mm. Leaving a 1 mm margin of air in the slot top, the slot fill factor is incredible 0.78, for a slot with a total height of 14 mm. The active parts of the machine consisting of permanent magnets, copper, SMC and steel (for PM rotor back) have a weight of 1.3, 3.1, 9.0 and 6.3 kg, respectively, and a total weight of 19.7 kg. The cogging torque is low due to the shape of the permanent magnets in the axial airgaps and the stepwise skewing of the permanent magnets in the radial airgap.

VII. CONCLUSIONS

This paper has analysed a novel machine concept using SMC for the core. Measurements on a laboratory prototype have been performed and the results combined with analytical calculations and finite element method computations are used to evaluate the performance of the machine. The topology looks promising and a high torque density of 5.1 Nm/kg is estimated for forced air-cooling. The torque production capability is not verified in practice due to a winding fault. The typically high power factor of the machine concept saves cost on the required inverter. The concept is most suitable in low-speed applications, since only high pole numbers are practically feasible.

Fig. 18. The Soft Magnetic Composites core and the rectangular copper conductors shown in parts and assembled.

Fig. 19. The assembly of the machine with both PM rotor and SMC rotor.

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REFERENCES

Curriculum Vitae

The author was born in 1971 in Sarpsborg, Norway, where he also attended fundamental school and high school. From 1990 he studied Electrical Engineering at Chalmers University of Technology in Gothenburg, Sweden, from where he received his M.Sc. degree in 1994. In 1995 he did his military service for Norway working as a clerk at the NATO Programming Centre in Belgium. Since 1996 he has been working with electrical machine design at ABB Corporate Research in Sweden. He joined the division of Electrical Machines and Power Electronics at the Royal Institute of Technology in the end of 2000. In the spring of 2001 he performed guest studies at the Renault research centre in Paris, France. He is inventor of a patent on electrical machine design for an application with extreme overload and fast acceleration operation. He is author of eleven technical papers.