



The University of
Nottingham

UNITED KINGDOM • CHINA • MALAYSIA

University of Nottingham

Faculty of Engineering

Department of Electrical and Electronic Engineering

**Advancing the Design and Modelling Processes
for High Performance Electrical Machines**

Ahmed Al-Timimy, B.Sc, M.Sc.

Thesis submitted to the University of Nottingham
for the degree of Doctor of Philosophy

May 2019

ABSTRACT

The implementation of an integrated oil pump and drive motor (where no mechanical coupling between the motor and pump is required) is a very attractive solution for various industries, especially where high performance requirements are important, such as for the aerospace industry. As in many applications, high power density and high efficiency are very important design objectives. However, to achieve such objectives while keeping the overall weight of the electrical machine to minimum values is very challenging and requires an optimal machine design with a detailed knowledge of power losses. To address these aspects, this thesis aims to improve the machine performance through accurate modelling by considering often overlooked aspects during the development process of electrical machines.

To achieve this goal, the first stage of the project involved the analysis, design and optimisation of a high performance, high power density electrical machine for directly-coupled pump application. The second stage has shown how accurate machine models that include the effects of machining processes required to produce the finished product (e.g. the stator core) and the effects of inactive component (e.g. rotor endcaps), can be achieved and used to provide a more accurate estimation for the machine performance than with the traditional methodologies.

A prototype machine based on the first stage of this project is manufactured and experimentally tested. This machine was then used as a vessel for the second stage of the work. The main contribution of this work is the investigation and verification of the effects on the accuracy of design models due to 1) the detrimental effects of machining processes of the stator core and 2) the detrimental effects of leakage fluxes flowing in the inactive materials/components of the machine. This knowledge is then used to inform the future machine designer on how to take into consideration these aspects during the design of electrical machines.

ACKNOWLEDGMENT

First and foremost, I would like to express my deepest appreciation to my sponsor, the Higher Committee for Education Development in Iraq, for receiving the opportunity to study in the United Kingdom.

I would also like to thank my supervisors. Professor Michael Galea, Professor Chris Gerada, Dr. Paolo Giangrande and Dr. Michele Degano for their guidance, encouragement and support throughout this PhD study.

Special thanks to all technicians in the university workshop for their support and effort to make this work success and providing an excellent work environment.

My thanks go to all my colleagues in the Department of Electrical Engineering, Power Electronics, Machine and Control (PEMC) Group where I spent four years of my life's best time.

Finally, I would like to thank my family for their continued support over the duration of the project. With much love, I want to thank my wife, Marwah and my children, Hasan and Yousif for their love, patience and support during all these years.

TABLE OF CONTENTS

ABSTRACT	II
ACKNOWLEDGMENT	III
TABLE OF CONTENTS	IV
LIST OF FIGURES	VIII
LIST OF TABLES	XIII
LIST OF ABBREVIATION	XV
LIST OF SYMBOLS	XVII
CHAPTER 1. INTRODUCTION	20
1.1 PROJECT DESCRIPTION	22
1.2 AIM AND OBJECTIVES OF THE THESIS	23
1.3 THESIS OUTLINE	24
1.4 SCIENTIFIC CONTRIBUTION AND LIST OF RELATED PUBLICATIONS	26
CHAPTER 2. PMSM AND LOSSES MODELLING	29
2.1 INTRODUCTION	29
2.2 ELECTRICAL ACTUATORS IN AEROSPACE	29
2.3 ELECTRICAL MOTORS FOR AEROSPACE ACTUATORS.....	31
2.4 PMSMs	32
2.5 LOSSES IN THE PMSM	33
2.5.1 Copper losses.....	33
2.5.2 Iron losses.....	35
2.5.3 Eddy current losses in the rotor	37
2.5.4 Mechanical losses.....	38
2.6 CONCLUSION.....	43
CHAPTER 3. INITIAL DESIGN OF THE INDUSTRIAL MACHINE	44
3.1 INTRODUCTION	44
3.2 DESIGN SPECIFICATIONS AND CONSTRAINTS	44
3.3 DESIGN PROCESS	47
3.3.1 Winding configuration.....	47
3.3.2 Slot-pole combination	49
3.3.3 Material selection.....	51
3.4 INITIAL DESIGN CONSIDERATIONS	55
3.5 PRELIMINARY MACHINE DESIGN	57

3.5.1	<i>Initial comparisons</i>	57
3.6	THERMAL MANAGEMENT	60
3.6.1	<i>Cooling concepts considered</i>	61
3.6.2	<i>Deciding of cooling options</i>	63
3.7	INITIAL VERIFICATION	66
3.8	MECHANICAL-ROTOR MODEL	69
3.8.1	<i>Design results</i>	69
3.9	MOTOR SIZING	72
3.9.1	<i>Rotor dimensioning</i>	72
3.9.2	<i>Airgap and PM thickness</i>	73
3.9.3	<i>Relative magnet span</i>	76
3.9.4	<i>Relative tooth width and stator back iron</i>	76
3.9.5	<i>Relative slot opening</i>	78
3.9.6	<i>Relative rotor back iron (hollow shaft)</i>	79
3.10	INITIAL COMPARISON TO THE MACHINE TARGETS	80
3.11	HIGHLIGHTS AND CONTRIBUTIONS TO RESEARCH	82
3.12	CONCLUSION	83
CHAPTER 4. FINAL DESIGN OF THE INDUSTRIAL MACHINE AND PERFORMANCE ANALYSIS		84
4.1	INTRODUCTION	84
4.2	IMPROVED MACHINE DESIGN	84
4.2.1	<i>Changing the stator lamination material</i>	84
4.2.2	<i>Changing the magnetization structure of the PM</i>	88
4.3	DEMAGNETIZATION ANALYSIS	90
4.4	UPDATED COMPARISON TO THE MACHINE TARGETS	91
4.5	LOSSES MODELLING	93
4.5.1	<i>Copper losses</i>	93
4.5.2	<i>Stator core losses</i>	96
4.5.3	<i>Rotor eddy-current losses</i>	97
4.5.4	<i>Mechanical losses</i>	100
4.6	PERFORMANCE EVALUATION	102
4.7	CONCLUSION	112
CHAPTER 5. MANUFACTURING AND EXPERIMENTAL VALIDATION		113
5.1	INTRODUCTION	113
5.2	MANUFACTURING AND CONSTRUCTION	113
5.2.1	<i>Stator construction</i>	113
5.2.2	<i>Coils, cooling channels and oil compatibility</i>	114
5.2.3	<i>Rotor construction</i>	116
5.2.4	<i>Housing construction</i>	117
5.3	OVERALL STRUCTURE AND ASSEMBLY	118
5.4	EXPERIMENTAL MEASUREMENTS AND VALIDATION	121
5.4.1	<i>Resistance measurements</i>	121

5.4.2	<i>Inductance measurements</i>	127
5.4.3	<i>Back-EMF measurements</i>	131
5.4.4	<i>Windage losses measurements</i>	135
5.4.5	<i>Characteristics torque vs current measurements</i>	139
5.4.6	<i>Motor efficiency measurements</i>	140
5.4.7	<i>Harmonic losses</i>	142
5.4.8	<i>Thermal measurements</i>	145
5.5	CONCLUSION	146
CHAPTER 6. INFLUENCE OF STATOR CORE MANUFACTURING ON MACHINE PERFORMANCE		147
6.1	INTRODUCTION	147
6.2	MATERIAL AND MANUFACTURING EFFECTS ON MACHINE PERFORMANCE	149
6.3	CHARACTERISATION OF MAGNETIC MATERIAL PROPERTIES	151
6.3.1	<i>Description of the test lamination sheet and the manufactured stator core</i>	151
6.3.2	<i>Description of the measuring setup</i>	153
6.3.3	<i>Investigated samples</i>	154
6.3.4	<i>Magnetic measurements</i>	154
6.4	MEASUREMENT RESULTS	156
6.4.1	<i>Basic magnetic properties</i>	156
6.4.2	<i>Core losses</i>	157
6.4.3	<i>Validating the negligible leakage assumption</i>	159
6.5	THE INFLUENCE OF MATERIAL PROPERTIES ON MACHINE EFFICIENCY PERFORMANCE	161
6.6	HIGHLIGHTS OF THE CHAPTER AND CONTRIBUTION	165
6.7	CONCLUSION	166
CHAPTER 7. INFLUENCE OF ROTOR ENDCAPS ON MACHINE PERFORMANCE		168
7.1	INTRODUCTION	168
7.2	DESIGN CONSIDERATION	171
7.2.1	<i>The use of a non-permeable rotor shaft</i>	171
7.2.2	<i>Mechanical benefits of the end-caps</i>	172
7.2.3	<i>Electro-magnetic aspects of the end-caps</i>	173
7.3	INFLUENCE OF ENDCAPS ON ELECTROMAGNETIC PERFORMANCE	174
7.3.1	<i>Open circuit</i>	175
7.3.2	<i>Electromagnetic torque</i>	177
7.4	INFLUENCE OF ENDCAPS ON TORQUE-SPEED CHARACTERISTICS	178
7.4.1	<i>PM flux linkage</i>	178
7.4.2	<i>Inductance</i>	179
7.4.3	<i>Torque and power speed curves</i>	179
7.5	INFLUENCE OF ENDCAPS ON CORE AND ROTOR EDDY-CURRENT LOSSES AND EFFICIENCY	180
7.5.1	<i>Core losses</i>	180

7.5.2	<i>Rotor losses</i>	180
7.5.3	<i>SPM efficiency</i>	182
7.6	PERFORMANCE EVALUATION	183
7.7	HIGHLIGHTS OF THE CHAPTER AND CONTRIBUTION.....	184
7.8	CONCLUSION.....	184
CHAPTER 8. IMPROVING THE PERFORMANCE OF THE CASE STUDY MACHINE THROUGH MORE ACCURATE MODELLING		186
8.1	INTRODUCTION	186
8.2	IMPLEMENTATION OF THE ACCURATE MODELLING.....	187
8.3	RESULTS AND DISCUSSION	188
8.4	CONSIDERATIONS ON THE UPDATED MODEL.....	192
8.5	FINAL RECOMMENDATIONS FOR THE NEXT GENERATION DESIGN OF CASE STUDY MACHINE	193
8.6	HIGHLIGHTS OF THE CHAPTER AND CONTRIBUTION.....	194
8.7	CONCLUSION.....	195
CHAPTER 9. CONCLUSION AND FUTURE WORK.....		196
9.1	GENERAL CONCLUSION	196
9.1.1	<i>Machine design and optimization</i>	196
9.1.2	<i>Core material machining</i>	197
9.1.3	<i>Permeable rotor endcaps</i>	198
9.2	FUTURE WORK	199
9.2.1	<i>Experimental tests</i>	199
9.2.2	<i>Machine design</i>	200
9.2.3	<i>Modelling processes</i>	201
APPENDIX A. TECHNICAL SOFTWARE		202
APPENDIX B. STATOR CORE MATERIAL CHARACTERIZATION .		203
APPENDIX C. THERMAL MODELLING		208
APPENDIX D. MECHANICAL STRESS MODELLING.....		215
REFERENCES		218

LIST OF FIGURES

Figure 2-1: Different types of actuators.....	29
Figure 2-2: PMSM structure	32
Figure 2-3: Power losses in PMSM	33
Figure 3-1: Required torque-speed characteristic	45
Figure 3-2: Winding configurations of different topologies.....	47
Figure 3-3: Winding layout.....	48
Figure 3-4: Magnetization BH curve and power loss for the M270-35A [102]	52
Figure 3-5: Sm2Co17 demagnetization curves [104]	53
Figure 3-6: Magnetization BH curve for 17-4PH Stainless Steel material.....	54
Figure 3-7: Winding distribution with different fill factor	56
Figure 3-8: FE models of the updated slot-pole combinations.....	58
Figure 3-9: Flux density map of the updated slot-pole combinations	58
Figure 3-10: Schematic of a PMSM thermal network in MotorCAD	60
Figure 3-11: Semi flooded cooling arrangement	62
Figure 3-12: Flooded cooling arrangement.....	62
Figure 3-13: Windage losses versus speed	63
Figure 3-14: Windage losses, temperature and torque with different cooling system.....	65
Figure 3-15: Cross-section area of the PMSMs with cooling channel	66
Figure 3-16: Temperature distribution in the machines.....	67
Figure 3-17: Stresses in permanent magnet and sleeve	70
Figure 3-18: Stresses in permanent magnet and sleeve at maximum speed	71
Figure 3-19: Cross-section of the 8p/9s PMSM	72
Figure 3-20: Rotor moment of inertia versus rotor diameter	73
Figure 3-21: Influence of airgap and magnet thickness on.....	75
Figure 3-22: Influence of relative magnet span on	76
Figure 3-23: Variation of the losses with tooth width and back iron at rated speed....	77
Figure 3-24: Influence of relative slot opening span on	78
Figure 3-25: Influence of rotor back iron on	79
Figure 3-26: Cross-section of 8p/9s PMSM and the flux density map.....	80

Figure 4-1: BH curve for the stator lamination material.....	85
Figure 4-2: Flux density map.....	85
Figure 4-3: Cooling channel in a slot of dummy stator	86
Figure 4-4: Cross-section of the PMSM with different rotor PM magnetization	88
Figure 4-5: Performance parameters versus PM structures	89
Figure 4-6: Demagnetization prediction at $T_{pm}= 350\text{ }^{\circ}\text{C}$ and 5 Nm @ 19000 rpm...91	
Figure 4-7: Cross-section of 8p/9s PMSM and the flux density map at rated torque..92	
Figure 4-8: Cross-section of stator core with different number of strands per turn 93	
Figure 4-9: Ratio of ac to dc copper losses with different number of strands	94
Figure 4-10: Cross-section of 8p/9s PMSM with 12 turns per coil and 4 strands	94
Figure 4-11: Ratio of ac to dc copper losses.....	95
Figure 4-12: Flux density distribution in the stator winding using 3D FE model	95
Figure 4-13: Stator core losses of 8p/9s PMSM	96
Figure 4-14: Flux density distribution in the stator core.....	97
Figure 4-15: Current density distribution in solid rotor shaft at 19000 rpm.....	97
Figure 4-16: Magnet, sleeve and rotor shaft losses of 8p/9s PMSM.....	98
Figure 4-17: Effect of segmentation of eddy current losses	99
Figure 4-18: Kinematic viscosity vs. temperature for the adopted cooling oil.....	100
Figure 4-19: The effect of temperature and speed on	101
Figure 4-20: Bearing losses at different speeds using SKF model	102
Figure 4-21: No load flux density	103
Figure 4-22: Back-EMF of the 8p/9s PMSM	104
Figure 4-23: Torque at full load condition @ 8700 rpm and 80 A.....	105
Figure 4-24: Full load flux density map and flux line distribution.....	105
Figure 4-25: PM flux linkage at different q-axis currents	106
Figure 4-26: Inductances calculations at different currents.....	107
Figure 4-27: Torque-speed, Power-speed characteristics	107
Figure 4-28: Efficiency characteristics	108
Figure 4-29: Cross-section area of the 8p/9s PMSM.....	109
Figure 4-30: Temperature distribution in the machine for a condition 19000 rpm ...	109
Figure 4-31: An axisymmetric structural FEA analysis model of the rotor	110

Figure 4-32: An equivalent (von-Mises) stress in the rotor at 20 krpm.....	110
Figure 4-33: Shaft deformation due to unbalanced force	111
Figure 4-34: The first bending mode of rotor assembly	111
Figure 5-1: Stator core of PMSM	114
Figure 5-2: Stator core with winding and details of the cooling channel	115
Figure 5-3: Rotor shaft construction of PMSM	117
Figure 5-4: Housing construction of PMSM	118
Figure 5-5: PMSM final motor assembly	118
Figure 5-6: PMSM: final prototype	119
Figure 5-7: Experimental test bench.....	119
Figure 5-8: Test rig set-up high	120
Figure 5-9: Phase resistance as function of frequency (LCR bridge measurements)	123
Figure 5-10: Analysed stator / winding assembly.....	124
Figure 5-11: Experimental setup.....	124
Figure 5-12: Winding temperature under $I_{rms}=25$ A with different frequencies.....	125
Figure 5-13: DC phase resistance versus temperature	126
Figure 5-14: Phase resistance versus frequency	126
Figure 5-15: Oscilloscope and AC power supply connection for L_d inductance measurement.	127
Figure 5-16: DC power supply connection for d-axis alignment	128
Figure 5-17: DC power supply connection for q-axis alignment	128
Figure 5-18: Experimental setup for the d and q-axis alignment.....	129
Figure 5-19: Measured voltage and current waveforms	130
Figure 5-20: Measured flux linkage waveform	130
Figure 5-21: Measured and predicted d-axis inductance	130
Figure 5-22: Measured and predicted q-axis inductance	131
Figure 5-23: No load line to line Back-EMF waveform at 8700 rpm	132
Figure 5-24: Comparison of the no load line to line back-EMF waveform.....	132
Figure 5-25: FFT of the line-to-line back EMF (between phase A and B).....	132
Figure 5-26: No load line to line Back-EMF waveform at 19000 rpm	133
Figure 5-27: Comparision of the no load line to line Back-EMF waveform.....	134
Figure 5-28: FFT of the line-to-line back EMF (between phase A and B).....	134

Figure 5-29: Amplitude of the line-to-line back EMF as function of the speed.....	135
Figure 5-30: Dummy motor components for windage loss measurement.....	136
Figure 5-31: Windage losses experimental setup	137
Figure 5-32: Measured windage losses versus speed	138
Figure 5-33: Measured torque versus current at different speeds.....	139
Figure 5-34: The measured and calculated torque versus current	140
Figure 5-35: Schematic of the efficiency measurements set-up	140
Figure 5-36: Comparison between the measured and calculated losses	141
Figure 5-37: Efficiency map of the machine at different speeds and loads.....	142
Figure 5-38: Efficiency map of the machine at base speed and maximum load	142
Figure 5-39: Phase current waveforms and harmonic content at $F_o=580$ Hz and $F_c=16$ kHz.....	144
Figure 5-40: Eddy current density distribution in the rotor at 8700 rpm.....	144
Figure 5-41: Comparison of the losses at $F_o=580$ Hz and $F_c=16$ KHz.....	145
Figure 5-42: Measured temperatures under rated load condition at 8700 rpm.....	145
Figure 6-1: Prepared material sample for testing.....	152
Figure 6-2: CoFe material stator core (left) and wound core under test (right).....	152
Figure 6-3: Magnetic measurements system, MPG 200 Brockhaus setup bench.....	153
Figure 6-4: Comparison among BH magnetization curves.....	157
Figure 6-5: Comparison among DC relative permeability	157
Figure 6-6: AC hysteresis loop at 2 T and 1000 Hz	158
Figure 6-7: Specific core loss comparisons utilizing the three DSs	159
Figure 6-8: Field flux lines in the stator core.....	160
Figure 6-9: BH magnetization curves for the measured stator core	160
Figure 6-10: Stator core losses and efficiency at rated torque and different speeds..	162
Figure 6-11: Efficiency map of the PMSM	163
Figure 6-12: Stator core losses map of the PMSM.....	164
Figure 6-13: Torque reduction map of the PMSM	164
Figure 7-1: Models of the PMSM.....	170
Figure 7-2: Flux density distribution in the M3 model (80A, 8700rpm).....	172
Figure 7-3: Axial deformation (mm) in the PMSM rotor with and without endcaps	173
Figure 7-4: The influence of different endcap material	174

Figure 7-5: No load performance of the PMSM with / without endcaps.....	177
Figure 7-6: Torque waveform of the PMSM with and without endcaps at 8700 rpm	177
Figure 7-7: PM flux-linkage at different q-axis current levels	178
Figure 7-8: D-axis and q-axis inductances.....	179
Figure 7-9: Torque-speed characteristics.....	179
Figure 7-10: Core loss at different q-axis currents	181
Figure 7-11: Rotor loss at different q-axis currents	181
Figure 7-12: Eddy-current density distribution in rotor components of M3 model ..	182
Figure 8-1: PMSM: final FE models	187
Figure 8-2: Permeable rotor shaft (from Chapter 7)	187
Figure 8-3: Comparison among the material properties (from Chapter 6)	188
Figure 8-4: Comparison of the back-EMF at 8700 rpm	189
Figure 8-5: Comparison of the back-EMF at 19000 rpm	190
Figure 8-6: The measured and calculated torque versus current	190
Figure 8-7: Losses at different operating conditions	191
Figure 8-8: Efficiency at different operating conditions.....	191
Figure B-1: Specific core losses	203
Figure B-2: The relation between the P/f with different values of B.....	204
Figure B-3: AC hysteresis at different for the characterized core	207
Figure C-1: Schematic of a PMSM thermal network in MotorCAD.....	208
Figure C-2: Thermal nodes location for the PMSM	210
Figure D-1: Rotor model.....	215

LIST OF TABLES

Table 3-1: Machine design requirements and performance indicators	45
Table 3-2: Fundamental winding factor and harmonic leakage factor for concentrated windings with different slot/pole combinations.....	49
Table 3-3: Magnet properties - Recoma 33E [104]	53
Table 3-4: Material properties	55
Table 3-5: Initial design parameters	57
Table 3-6: Initial performance comparison.....	59
Table 3-7: Losses distribution of PMSMs at 5 Nm and 19000 rpm	66
Table 3-8: Thermal comparison between the 8p/9s and 8p/12s PMSMs	68
Table 3-9: Machine performance for different airgap thicknesses	74
Table 3-10: Machine dimensions for preliminary design	80
Table 3-11: Machine specifications for preliminary design	81
Table 3-12: Comparison of PMSM design to machine targets.....	82
Table 4-1: SiFe vs. CoFe performance comparisons @ 10.5 Nm	85
Table 4-2: General Machine Dimension.....	87
Table 4-3: General Machine Specifications.....	87
Table 4-4: SMPM vs. Quasi-Halbach performance comparisons for a fixed envelop size	89
Table 4-5: SMPM vs. Quasi-Halbach performance comparisons for a fixed output torque	90
Table 4-6: Comparison of PMSM design to machine targets.....	92
Table 4-7: Maximum allowable stress limit for rotor components.....	110
Table 5-1: DC resistance measurments using the Valhalla Scientific 4300B and FEA	121
Table 5-2: AC resistance measurments using the LCR Bridge HM8118 and FEA...	122
Table 5-3: No-load results at 8700 rpm	133
Table 5-4: No-load results at 19000 rpm	134
Table 5-5: Rrotor shaft dimenssions	137
Table 5-6: Physical properties of the coolant oil	138
Table 6-1: Material properties of the investigated CoFe material	152

Table 6-2: Specification and capabilities for the measurement system	154
Table 6-3: Comparison of the Back-EMF for different data sheets.....	161
Table 7-1: Average torque of the investigated machine	172
Table 7-2: Material properties	174
Table 7-3: Efficiency at different input conditions.....	183
Table 7-4: Performance comparison of PMSM with and without rotor endcaps	183
Table 8-1: Performance comparison of PMSM with different models.....	192
Table 8-2: Performance of PMSM with new torque limit (10.5 Nm)	193
Table B-1: The coefficients of iron losses calculation.....	205
Table C-1: Symbols description for the schematic of thermal network	208
Table C-2: Colour description for the thermal resistances of the LPN (MotorCAD)	209
Table C-3: Mass density, Heat capacity and thermal conductivity for the different parts of the machine	211
Table D-1: Material properties.....	215

LIST OF ABBREVIATION

2D	Two Dimensions
3D	Three Dimensions
AC	Alternating Current
AWG	American Wire Gage
BEMF	Back Electromotive-Force
BLPM	Brushless Permanent Magnet
BLAC	Brushless AC
BLDC	Brushless DC
CoFe	Cobalt Iron
DC	Direct Current
DS	Data Sheet
EHA	Electro Hydraulic Actuator
EMA	Electro Mechanical Actuator
EDM	Electro Discharge Machining
FE	Finite Element
FEA	Finite Element Analysis
GCD	Greatest common divider
HA	Hydraulic Actuator
IM	Induction Motor
IPM	Interior Permanent Magnet
LP	Lumped Parameter
LCM	Least Common Multiplier
MMF	Magnet Motive Force
MEA	More Electric Aircraft
NdFeB	Neodymium Iron Boron
PBW	Power by Wire
PMSM	Permanent Magnet Synchronous Motors
PEC	Power Electronic Converter
PM	Permanent Magnet
PWM	Pulse Width Modulation

SBI	Stator Back Iron
SR	Switched Reluctance
SMPM	Surface Mounted Permanent Magnet
SiFe	Silicon Steel
SmCo	Samarium Cobalt
THD	Total Harmonic Distortion
TRL	Technology Readiness Level
UMP	Unbalanced Magnetic Pull

LIST OF SYMBOLS

A_{core}	Cross-sectional area of the core (mm ²)
$B_{(t)}$	Flux density (T)
C_f	Friction coefficient
C_a	Cooling area (mm ²)
D_{so}	Stator outer diameter (mm)
D_{si}	Stator inner diameter (mm)
F	Frequency (Hz)
E	Young's modulus (GPa)
FF	Fill factor
$H_{(t)}$	Magnetic field strength (A/m)
I_q	Current along the quadrature q-axis (A)
I_d	Current along the direct d-axis (A)
i_1, i_2	Primary and secondary currents (A)
I_{rms}	Current rms value (A)
J_n	Eddy current density (A/mm ²)
J_r	Moment of inertia (kg.m ²)
K	Roughness coefficient
K_v	Voltage constant (Vs/rad)
K_t	Torque constant (Nm/A)
k_w	Fundamental winding factor
L	Stack length (mm)
L_{ax}	Axial length of the rotor (mm)
L_{core}	Mean length of the magnetic flux path (mm)
L_g	Airgap length (mm)
L_s	Sleeve thickness (mm)
L_q	Inductance along the quadrature q-axis (mH)
L_d	Inductance along the direct d-axis (mH)
M_s	Magnet span (mm)
M_t	Magnet thickness (mm)

M	Mass (Kg)
N_1, N_2	Primary and secondary number of turns
N_{base}	Base speed (rpm)
N_{max}	Maximum speed (rpm)
N_s	Number of slots
P_{in}	Input power (W)
P_p	Pole pitch (mm)
p	Number of poles
P_{cu}	Copper losses (W)
$P_{specific}$	Stator core loss density (W/kg)
P_{core}	Stator core losses (W)
P_{eddy}	Rotor eddy current losses (W)
P_{mech}	Mechanical power (W)
P_w	Windage losses (W)
q	Number of slot per pole per phase
r	Rotor radius (mm)
Re_δ	Reynold number
R_{ac}	AC resistance of phase A (Ω)
R_{dc}	DC resistance of phase A (Ω)
$R_{act}(T, f)$	Resistance of phase A, active length (Ω)
$R_{end}(T, f)$	Resistance of phase A, end-winding length (Ω)
R_{ph}	Resistance of phase A, measured value (Ω)
R_{OD}	Outer radius of the stator yoke (mm)
R_{iD}	Inner radius of the stator yoke (mm)
R	Winding phase resistance (Ω)
R_{bi}	Rotor back iron (mm)
S_{OD}	Stator outer diameter (mm)
S_r	Split ratio
S_o	Slot opening (mm)
S_p	Slot pitch (mm)
S_{bi}	Stator back iron (mm)

S_l	Stator core length (mm)
T_{em}	Electromagnetic torque (Nm)
T_r	Torque ripple (%)
T_{rated}	Rated torque (Nm)
T	Temperature (°C)
T_t	Tooth tip (mm)
T_w	Tooth width (mm)
T_p	Slot pitch (mm)
ν	Poisson ratio
v_q	Voltage along the quadrature q-axis (V)
v_d	Voltage along the direct d-axis (V)
v_1, v_2	Primary and secondary voltages (V)
V_{l-l}	Voltage, terminal value (V)
V_{pk}	Voltage, phase value (V)
ω	Angular speed (rad/s)
ρ	Electrical resistivity ($\Omega \cdot m$)
ρ_m	Mass density of the fluid (kg/m^3)
ρ_{fe}	Mass density of the rotor components (kg/m^3)
ρ_{co}	Mass density of the CoFe material (kg/m^3)
σ_c	Circumferential stress component (MPa)
σ_r	Radial stress component (MPa)
σ_v	Von Mises (equivalent) stress (MPa)
σ_δ	Harmonic airgap leakage factor
η	Efficiency (%)
ρ_o	Electrical resistivity of the copper at $T_o = 20^\circ C$
α	Temperature coefficient of resistivity
ψ_d	Flux linkage along the quadrature q-axis (Wb)
ψ_q	Flux linkage along the direct d-axis (Wb)
ψ_{PM}	Flux linkage, no load (Wb)
θ	Rotor position (Degree)

Chapter 1. INTRODUCTION

In recent years, demanding regulations on fuel emissions and the need of decreasing both the operating and the maintenance costs have encouraged the aircraft industry to move from conventional hydraulic to electrical actuation systems [1, 2]. This shift towards the More Electric Aircraft (MEA) concept via the use of Electro Hydraulic Actuators (EHAs) and/or Electro Mechanical Actuators (EMAs) can reap excellent benefits, for example in terms total system weight reduction due to the removal of the hydraulic network. Other advantages include the possibility of self-contained distributed actuation and the high dynamic performance which can be ensured by both low inertia motors and high-bandwidth controllers [3]. This is particularly important for aerospace applications, where the high performance actuation system should meet strict requirements related to the safety, reliability, efficiency and weight challenges compared to Hydraulic Actuators (HAs), which suffer from the heavy hydraulic plant, extensive mechanical linkage and maintenance [4, 5].

It is well known that a more compact EHA can be achieved by reducing the size, i.e. increasing the power density of one its main components, namely the electrical machine [6]. Today, there is a wealth of literature on the design and development of high performance machines [2]. Several examples of high performance electrical machines for aerospace actuators exist [7-12]. Most of these usually focus on the design process of electrical machines and the use of high performance materials and configurations in order to improve the performance of the aerospace actuators. All this indicates that for aerospace applications, the need for improvement in terms of size, efficiency and power density is a main issue which the electrical machine designer has to consider through all the design stages.

However, the extreme demands in terms of power density and efficiency are also resulting in an increasing trend that aims to consider effects that traditionally have been ignored in the design of electrical machines. These include considerations of losses due to the machining of the stator core and other aspects such as the losses due to leakage

fluxes in the supporting structures of the electrical machine such as rotor endcaps, which are usually placed at the end region of the rotor shaft and do not contribute to the torque production (i.e. inactive components).

This thesis therefore aims to fill the gap in knowledge related to these traditionally-ignored aspects in the design processes of electrical machines and to give a clear indication of how much of an effect on the overall performance such aspects have. To do this, a complete understanding of the power losses and their influence on the whole machine is required.

Therefore, a detailed analysis of the copper, iron and mechanical losses of electrical machines is required. Iron losses mainly depend on the material properties which can be affected by the manufacturing processes on the stator and rotor cores [13-15]. The uncertainties related to the prediction of the losses and efficiency when considering these core machining activities need to be examined, since the actual machine material properties might be changed [15, 16].

It is also recognized that an accurate machine model, including the end effects (e.g. end-windings and permeable machine structure), is important to the machine designers when the purpose of designing the machine with the required torque and estimate the efficiency correctly [17]. It is shown that neglecting the end effects could lead to a significant error in the main flux calculation and inductance [18]. Therefore, when a high confidence in the results is needed, the use of 3D-FEA becomes mandatory despite the well-known drawback of higher computational time [18].

All this indicates that with a complete knowledge about power losses and 3D end effects, an accurate estimation for the machine performance can be achieved and can then be used to improve the performance value of any machine, but most importantly those machines that are applied in applications where mass and efficiency are of the utmost priority such as in aerospace actuation applications.

1.1 Project description

A recent project within the Power Electronics, Machines and Control (PEMC) research group at the University of Nottingham (UoN) was focused on the development of a high performance, electrical machine for an aircraft EHA application. The project's aim is to develop an electrical machine with high power density and a high efficiency as required by the partner's specifications. The project involved the analysis, design and optimisation of a high performance, high power density electrical machine for a directly coupled pump application.

This thesis is therefore aligned with the above-mentioned project, where this machine is taken as a case study to investigate the usually-ignored considerations mentioned in the above introduction. In this thesis, the design process of this electrical machine is presented and explained. This thesis shows that an electrical machine can go beyond the current state of art expectations, in terms of efficiency, mass, inertia and VA (volt-ampere) capabilities. However, it was perceived that extra performance could be gained by considering the usually-ignored effects mentioned above. To achieve this extra improvement in the performance of electrical machines, the rest of the thesis focused on the losses and manufacturing aspects. Very accurate analysis and prediction tools were developed to fully characterise the prototype machine's losses and performance estimation requirements.

1.2 Aim and objectives of the thesis

The main aim of this work is to contribute to knowledge about the design of high performance, permanent magnet (PM) machines by considering certain losses that are usually ignored. The basic concept is that performance can be improved in terms of power density and efficiency if the machine designer considers traditionally-ignored aspects such as core machining losses and the effects of other losses due to for example rotor end-caps.

In order to reach this aim, a number of objectives were identified, as follows

- 1) An appropriate literature survey that covers the current knowledge in terms of
 - a. high performance machine design
 - b. modelling of losses and characterisation of PM machines and their magnetic circuits
 - c. considerations on the traditional and not so traditional, loss-inducing effects in electrical machines.
- 2) The design and development of the high performance machine as required by the industrial partner and the experimental validation of all the models compiled to achieve this design.
- 3) To investigate and analyse the effects that core machining (with focus on the stator laminations) has on the losses of electrical machines and therefore on their performance. The machine developed in Point 2) above is considered as a case study.
- 4) To investigate and analyse the effects that leakage flux in the machine has on its performance. This focuses mainly on the role of the rotor end-caps of the particular machine developed for the industrial project.
- 5) To gather and compile the learning from points 3) and 4) above and to show how the knowledge of these effects can be utilised enhance machine performance through better modelling of machines. An updated version of the original, industrial machine was proposed.

1.3 Thesis outline

In the following, the general structure of the PhD thesis is outlined.

1) Chapter 1 - Introduction

Acts as a general introduction to the project where the application is presented and its objectives and requirements, including the thesis outline, scientific contributions and a list of related publications.

2) Chapter 2 – PMSM and losses modelling

This chapter serves firstly as an introduction to electrical machines for aerospace applications and secondly as general literature review of power losses in electrical machines. The literature review includes details on the theory and calculation methods for power losses.

3) Chapter 3 – Initial design of the industrial machine

This chapter looks into the initial design, focusing on a permanent magnet synchronous motor (PMSM) for a direct coupled pump application. As well as the design procedures and the initial considerations associated with the application, this chapter shows the mechanical and thermal analysis of the designed machine.

4) Chapter 4 – Final design of the industrial machine and performance analysis

This chapter focuses on the optimization of PMSM and details the final design. Optimization procedures that consider high power density and high efficiency as the main objectives are included. A special focus is also given to the power losses analysis of the designed machine. Finally, the overall performance of the machine is analyzed and discussed, highlighting the achieved design targets.

5) Chapter 5 – Manufacturing and experimental validation

In order to validate the design procedures presented in Chapter 4, a prototype of the machine is built and tested in Chapter 5. Experimental results are presented and compared with the simulated FE results, all of which are complemented by discussions and analyses in this chapter.

6) Chapter 6 – Influence of stator core manufacturing on machine performance

This chapter covers the effect of manufacturing process on the magnetic properties of the hard-magnetic material (e.g. stator core). Detailed characterization processes are carried out in order to achieve a full understanding of this effect for the application at hand. Then the overall influence on the machine performance at different operating points is investigated and the results are provided at the end of the chapter. Special attention is focused on the efficiency maps of the machine, as this aspect is highly dependent on the material properties that are the most sensitive to manufacturing processes such as the material's anhysteretic BH curve and its specific core loss. One of the significant outcomes of this work has been the accurate estimation of the machine's performance.

7) Chapter 7 – Influence of rotor endcaps on machine performance

This chapter will investigate the influence of the leakage flux on the performance of the developed machine in Chapter 4, which is due to the rotor permeable endcaps. Accurate FE model with rotor endcaps is developed and validated in this chapter. Also, different endcap thickness and different rotor shaft materials are investigated and compared in this work. Finally, accurate results for torque, losses and efficiency (relative to machine at hand) are provided at the end of this chapter.

8) Chapter 8 – Improving the machine performance of the case study machine through more accurate modelling

Based on the material characterization of Chapter 6 and the rotor endcaps of Chapter 7, an updated, reliable FE model for the machine performance estimation is presented in Chapter 8. The accuracy of the updated model is investigated and verified with experimental measurements. Finally, recommendations for better design are proposed at the end of this chapter.

9) Chapter 9 - Conclusions and future works

This chapter summarizes the conclusions of the work and gives some recommendations and suggestions for future work.

1.4 Scientific Contribution and list of related publications

The contribution of this thesis can be summarised as:

- Development of a method for the design of high power density PMSM using a fully flooded cooling system, for EHA applications.
- Defining accurately the variations in the magnetic material properties and iron losses of a manufactured Cobalt-Iron stator core. The knowledge on the exact material properties allow us then to re-evaluate the machine performance at different operating points considering all the discrepancies due to the machining processes of the stator core. Therefore, accurate estimation for the efficiency, torque and stator core losses can be achieved.
- Predict and verify by experiments the influence of a permeable rotor endcaps (3D end effect) on the electromagnetic performance of the PMSM. Unlike the 2D-FE model, the existence of 3D-FE model without rotor endcaps showed inaccuracies in the computation of the flux linkages, back-EMF, inductances and then on the predicted torque-speed curve. Obviously, for the design of PMSM with permeable rotor shaft, the rotor endcaps must be taken into account in order to have an accurate estimation of the machine capabilities.

Several parts of this thesis have been presented in international conferences and journals by the author during his Ph.D. program. Hereafter the publications are listed:

International conferences

- Paolo Giangrande, **Ahmed Al-Timimy**, Alessandro Galassini, Savvas Papadopoulos, Michele Degano and Michael Galea, "Design of PMSM for EMA Employed in Secondary Flight Control System," *2018 International Conference on Electrical Systems for Aircraft, Railway, Ship Propulsion and Road Vehicles & International Transportation Electrification Conference (ESARS-ITEC)*, Nottingham, UK, 2018, pp. 1-6.

- **Ahmed Al-Timimy**, Paolo Giangrande, Michele Degano, Michael Galea and Chris Gerada, "Investigation of AC Copper and Iron Losses in High-Speed High-Power Density PMSM," 2018 *XXIII^d International Conference on Electrical Machines (ICEM)*, Alexandroupoli, Greece, 2018., 2018, pp. 263-269.
- H. Mahmoud, **A. Al-Timimy**, M. Degano, M. Di Nardo, N. Bianchi and C. Gerada, "Axial eccentric SynRel and SPM motors analytical models validation using 3D finite element," 2017 IEEE 11th International Symposium on Diagnostics for Electrical Machines, Power Electronics and Drives (SDEMPED), Tinos, 2017, pp. 38-44.
- **A. Al-Timimy**, P. Giangrande, M. Degano, M. Galea and C. Gerada, "Comparative study of permanent magnet-synchronous and permanent magnet-flux switching machines for high torque to inertia applications," 2017 IEEE Workshop on Electrical Machines Design, Control and Diagnosis (WEMDCD), Nottingham, United Kingdom, 2017, pp. 45-51.
- **A. Al-Timimy**, M. Degano, Z. Xu, G. Lo Calzo, P. Giangrande, M. Galea, C. Gerada, He. Zhang, L. Xia, "Trade-off analysis and design of a high-power density PM machine for flooded industrial pump," *IECON 2016 - 42nd Annual Conference of the IEEE Industrial Electronics Society*, Florence, Italy, 2016, pp. 1749-1754.
- Z. Xu, **A. Al-Timimy**, M. Degano, P. Giangrande, G. Lo Calzo, H. Zhang, M. Galea, C. Gerada, S. Pickering, L. Xia, "Thermal management of a permanent magnet motor for a directly coupled pump," 2016 XXII International Conference on Electrical Machines (ICEM), Lausanne, 2016, pp. 2738-2744.
- **A. Al-Timimy**, M. Degano, P. Giangrande, G. Lo Calzo, Z. Xu, M. Galea, C. Gerada, He. Zhang, L. Xia, "Design and optimization of a high-power density machine for flooded industrial pump," 2016 XXII International Conference on Electrical Machines (ICEM), Lausanne, 2016, pp. 1480-1486.

Journals

- Paolo Giangrande, Alessandro Galassini, Savvas Papadopoulos, **Ahmed Al-Timimy**, Giovanni Lo Calzo, Michele Degano, Michael Galea and Chris Gerada, "Considerations on the Development of an Electric Drive for a Secondary Flight Control Electromechanical Actuator," in (2019 Early Access) *IEEE Transactions on Industry Applications*, doi : 10.1109/TIA.2019.2907231.
- **Ahmed Al-Timimy**, Mahir Al-Ani, Michele Degano, Paolo Giangrande, Chris Gerada and Michael Galea, "Influence of Rotor Endcaps on the Electromagnetic Performance of High Speed PM Machine," in *IET Electric Power Applications*, vol. 12, no. 8, pp. 1142-1149, 9 2018.
- **Ahmed Al-Timimy**, Paolo Giangrande, Michele Degano, Zeyuan Xu, Michael Galea, Chris Gerada, Giovanni Lo Calzo, He Zhang and Linqun Xia., "Design and Losses Analysis of a High Power Density Machine for Flooded Pump Applications," in *IEEE Transactions on Industry Applications*, vol. 54, no. 4, pp. 3260-3270, July-Aug. 2018.
- **Ahmed Al-Timimy**, Gaurang Vakil, Michele Degano, Paolo Giangrande, Michael Galea and Chris Gerada, "Considerations on the effects that soft magnetic materials' manufacturing processes have on the electrical machine's performance," in *IEEE Transactions on Energy Conversion*, vol. 33, no. 3, pp. 1154-1163, Sept. 2018.

Chapter 2. PMSM AND LOSSES MODELLING

2.1 Introduction

This chapter presents an extensive literature review of PMSMs for aerospace actuators and their losses. First, an overview of the electrical machines related to the application under study, types and challenges, is given. Then, a description of the main power losses in electrical machines, including copper losses, iron losses and mechanical losses, are presented.

2.2 Electrical actuators in aerospace

With the continued developments of the aircraft industry, the majority of conventional hydraulic actuation systems have or are being replaced by electrically actuated systems, either partially or fully electrical. The main advantage is that this eliminates the needs for complex mechanical pipes and therefore lower overall weight, oil leakage reduction and increased efficiency [19]. Modern actuation systems on aircraft are divided into EHAs and EMAs [20, 21]. General schematics of these technologies are shown in Figure 2-1.

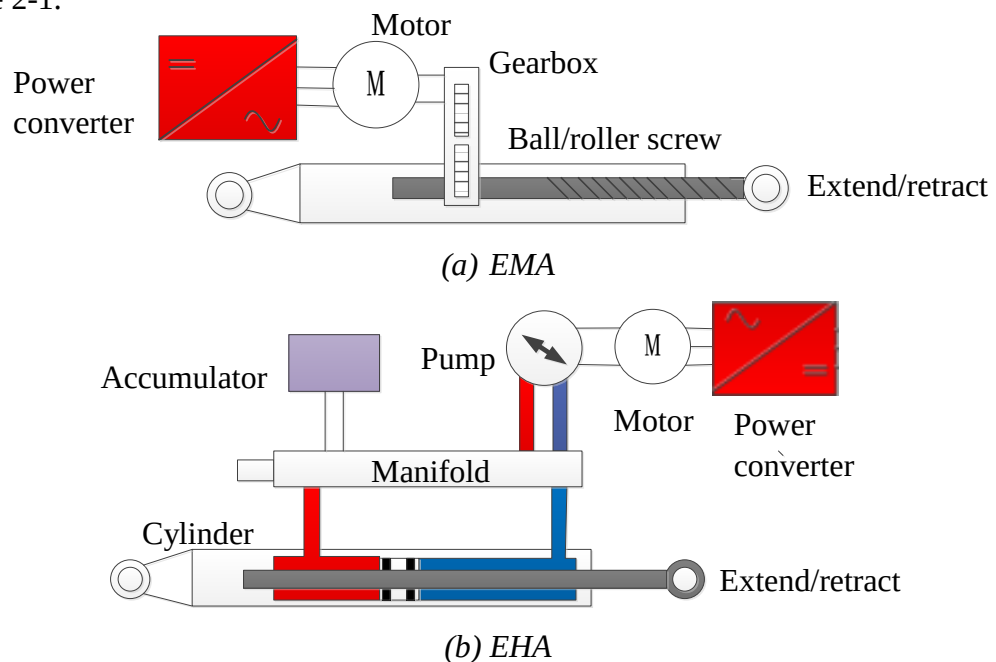


Figure 2-1: Different types of actuators

Today, both EMAs and EHAs are currently employed on large commercial aircraft. The first commercial use of EHA was for an Airbus 380 aircraft (2005) for a primary flight control surface (ailerons and elevators) [4]. The adoption of EHAs led to a 1500 kg weight reduction on Airbus A380 [22]. On the Boeing 787, EMAs are used for landing gear braking system and spoiler surface [23].

An EMA comprising an electrical machine, a gearbox or screw mechanism and usually a remote power electronics converter removes the need for the hydraulic system, resulting in an oil-free solution and typically with a lower total system weight, when compared to HAs [10]. EMAs comprise high force to mass ratio, the removal of potential hazardous hydraulic fluid and therefore are considered to be lower maintenance solutions [20, 24]. One of the downsides of EMAs is the possibility of mechanical jamming between the mechanical components [10]. This is usually taken care of by adding a backup drive (redundancy) at the cost of having higher weight and cost [4]. These cons regarding to the mechanical jamming and gearbox makes the EHAs a preferred solution, especially in those applications where the motor and the pump are embedded together, supplying the hydraulic system [10, 21]. In this type of embedded actuators, the electrical motor drives a hydraulic pump at a variable speed in order to move the hydraulic fluid [4, 20]. The main benefits of having a motor directly coupled to the pump are the elimination of the shaft seals and the potential improvement in the thermal rating of the machine [25-27]. Removing the shaft seals results in a reduced risk of oil leakages, while the thermal improvement comes from the fact that the coolant in the pumping system can also be used to cool the electrical machine itself [25, 27]. Another benefit of this integrated design concept is to simplify the control of the actuator. By changing the motor rotation direction, one can change the direction of the actuator force without having to use a high-pressure manifold. This is particularly useful for a hydraulic actuator system which needs to switch direction frequently [28]. However, EHAs still comprise a hydraulic system, however this is a contained system within the actuator, as opposed to the historical hydraulic system where oil used to be pumped long distances across the aircraft body.

When comparing these two technologies, it is clear that the main drawback of the EHA is the continued use of hydraulic and the main challenge with EMAs is the possibility of mechanical jamming. For safety critical applications such as aircraft, the risk of

mechanical jamming is today still considered as a major bottleneck. Therefore, EHAs are still the preferred solution in the aerospace field, due to the well-established “know-how” regarding hydraulic systems [10, 21].

2.3 Electrical motors for aerospace actuators

The electric motor is the driving element of the EHA or EMA system where a high power density and efficient motor play an important role in the overall performance of the actuation system [5, 29]. Most aerospace actuation applications today comprise either brushless permanent magnet (BLPM) motors or switched reluctance (SR) motors [7-12, 25, 30-37]. This is mainly due to the high power density and efficiency, high torque to inertia ratio and lightweight of the BLPM motors, and the fault tolerant capability and reliability characteristics of the SR motors [38]. Induction motors (IM) are the most widely used and well-known AC motor in general, however they have rarely been adopted in aerospace applications, mainly due to their lower power density and higher loss density when compared to the BLPM motors [2, 38, 39].

Permanent Magnet Synchronous Machine (PMSM) is one type of brushless PM machine. These are well-known to have very positive characteristics in terms of torque and power density, low losses and high efficiency and high design flexibility. Their main disadvantages are related to the high cost of the PMs (and the fluctuations in the market price), the inherent sensitivity to demagnetisation and temperature and the fact that these machines are permanently excited.

SR motors on the other hand are known to be very robust and relatively cheap to construct and therefore extremely appropriate for safety critical applications. However, when compared to PMSMs, their power and torque density performance is much lower [2, 9].

Based on the above, simplified qualitative analysis and based on the requirements of the application described in Chapter 3, it was immediately clear that only a PMSM solution would be able to satisfy these requirements. The rest of this literature review is therefore focused on PMSMs.

2.4 PMSMs

Generally, PMSMs can be categorized into two main groups, based on the position of the PMs on the rotor, namely surface mount PM (SMPM) motors and interior PM (IPM) motors [2]. In literature [40-43], extensive comparisons of these two topologies have been reported. For SMPM motors, the PMs are mounted on the surface of the rotor as shown Figure 2-2 (a). This arrangement provides higher torque density due to high air gap flux density, which potentially reduces the volume and mass of the machine [2, 44]. In addition, a SMPM motor has low rotor inertia and higher dynamic performance [2]. The main drawbacks of this configuration are the risk of PM demagnetization, since the PMs are directly exposed to various stator winding harmonics, as well as the need for retaining sleeves for the PMs especially for high speed machines [45]. IPM motors have their PMs “buried” inside the rotor as shown in Figure 2-2 (b). This configuration typically provide higher per unit inductances and thus field weakening capability [46]. Also, IPM motors are more robust and are often used for high-speed applications. Also, buried PMs provide more protection against demagnetization [47].

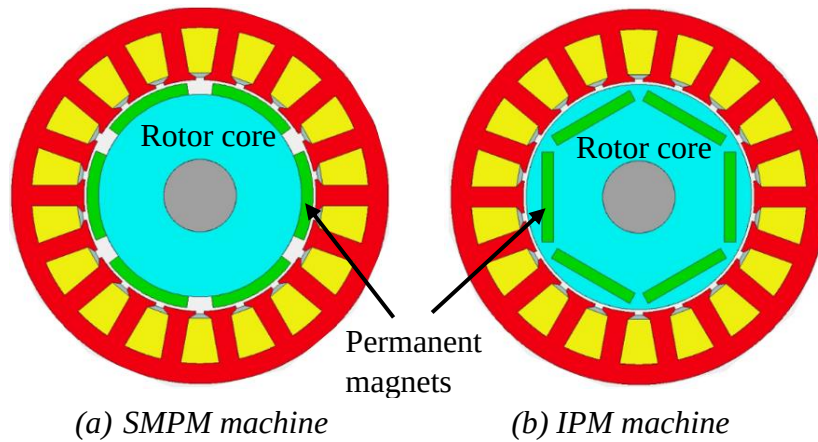


Figure 2-2: PMSM structure

Based on the application described in Chapter 3 in mind, then the SMPM machine retained by a sleeve is considered as an optimal solution due to its higher efficiency, higher torque and power density and simple structure [48]. An extra perceived advantage of this configuration is that lower rotor inertia can be achieved by having a hollow rotor shaft. Due to this, a decision was taken to concentrate the work on the SMPM configuration.

2.5 Losses in the PMSM

The main aim of this thesis is to understand and increase the knowledge related to the design of electrical machines relative to often-overlooked aspects such as losses due to manufacturing effects and leakage fluxes in the inactive material. To be able to do this, a thorough understanding of the losses in a PM machine is required. This Section is therefore a summary of the important points related to losses in electrical machines as known and as characterised today.

Power losses represent the main source of heat in electrical machines. The accurate prediction of the losses and their distribution is important especially for high-speed and high-power density applications.

In general, the power losses in electrical machines can be categorized into copper losses, iron losses, eddy current losses in the rotor (i.e. PM, sleeve, shaft) and mechanical losses as shown in Figure 2-3. The following sections describe these losses in more detail.

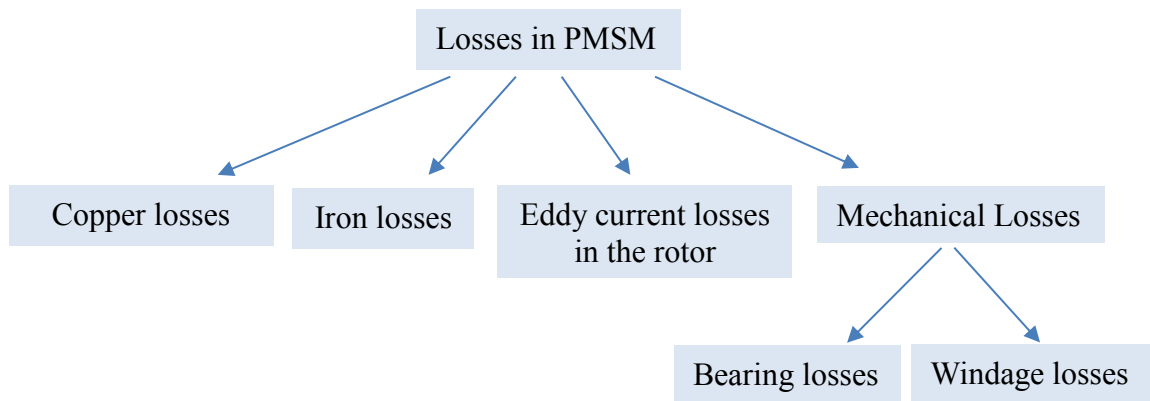


Figure 2-3: Power losses in PMSM

2.5.1 Copper losses

One of the major power losses components, for EMs is the stator copper losses or (I^2R) . For traditional machines, operating in steady state at relatively low speeds, this is considered to be a DC loss. However, as machines tend to become more power dense, the current trend towards higher speeds is resulting in the AC copper losses to become more and more important [49]. In PMSMs, this AC loss, i.e. the increasing copper losses relative to higher frequencies is mainly due to the skin and proximity effects,

which influence the winding resistance value. Both phenomena are caused by eddy currents. In the first case (i.e. skin effect), the eddy currents are self-induced by the same source conductors, while in the second case (i.e. proximity effect), the eddy currents are due to either the current flowing in nearby conductors sharing the same slot or the external magnetic field originating from the moving permanent magnets [50, 51]. The skin effect causes the current density distribution near the conductor surface to be higher than the current density inside the conductor when an AC current passes through the conductor itself. Thus, copper losses become higher. The skin effect depends on the frequency and the dimensions of the conductor and it can be dealt by choosing a conductor with diameter smaller than the skin depth δ [50], which is given by (2-1), where, ρ and μ_r are the electrical resistivity and the relative permeability of the conductor, respectively, f and μ_o are the frequency and the permeability of the free space, respectively.

$$\delta = \sqrt{\rho / \pi f \mu_o \mu_r} \quad 2-1$$

The proximity effect causes a non-uniform current density distribution inside the conductors. Losses due to this effect are dependent on how the conductors are placed inside the slots and it can be mitigated by either using windings with parallel strands (resulting in smaller diameter of the conductor) or using the Litz wire (thinner wire strands connected in parallel) [50, 52]. The main disadvantages rising from the adoption of windings with parallel strands are represented by the manufacturing complexity and the unbalanced current distribution across the strands. The latter results in a circulating current effect [51]. On the other hand, by using windings with parallel stands, a greater fill factor could be achieved with lower cost with respect to the Litz wire.

Therefore, the DC resistance of the phase winding must be computed including a factor that takes into account the AC effect due to the induced eddy currents.

Hence, the total copper losses in a solid conductor can be obtained by (2-2), where σ is the electric conductivity of the conductor, J_e is the induced eddy current density in the conductor and can be calculated based on the magnetic vector potential in the slot sub-domain [53]. J_c is the exciting current density.

$$P_{cu} = \frac{1}{\sigma} \iiint_V (J_e + J_c)^2 dV \quad 2-2$$

In general, both finite element (FE) methods and analytical models were widely used in the analysis of the AC copper losses in electrical machines [51, 53-62]. The FE method is accepted to be the most accurate method for electromagnetic field analysis. However, the application of FE method in the prediction of AC copper losses is very time consuming and requires significant computational resources. On the other hand, analytical models have been shown to be an alternative option for the analysis of AC copper losses with a faster solution time and acceptable results despite their limitations [53, 58, 59].

2.5.2 Iron losses

Machine iron losses can become quite cumbersome especially for high speed applications and therefore for appropriate designs, they need to be estimated accurately [15, 63]. However, accurate estimation of core losses is today still considered to be an extremely difficult task. Even FE methods still fail to achieve very accurate representations of these losses. This is mainly due to 1) the flexible nature of the loss coefficients of the materials used, 2) the dependence of these losses on various phenomena such as saturation, temperature and stresses, 3) the effects of hysteresis and various other factors.

Chapter 5 of this thesis will focus on effects that are usually ignored during the design processes of EMs, such as the effects due to the influence of manufacturing process on the stator core which could affect the material properties such as (BH curve and hysteresis losses) and thus affect the stator core losses [15]. In this section, however the common principles behind the monitoring of these losses are presented.

The total iron losses per mass unit (W/kg) can be calculated by (2-3) using the Modified Steinmetz method [64, 65], where the calculated losses include the hysteresis component P_h and the eddy current components P_e , while the excess losses are combined with the classical eddy current losses [66].

$$P_{core} = P_h + P_e \quad 2-3$$

In ferromagnetic material, the deflection and friction of magnetic domains under magnetization process might cause a loss. The losses caused by this effect is called

hysteresis loss and it's proportional to the area encircled by the hysteresis loop in the BH curve of the material. The magnetic energy dissipated is converted into heat and is permanently lost [64, 66]. However, the hysteresis losses can be reduced by selecting material with a narrow hysteresis loop. The hysteresis loss can be described as shown in (2-4), where, K_h is the hysteresis constant, f is the frequency, B is the peak value of the flux density and β is the Steinmetz constant.

$$P_h = K_h f B^\beta \quad 2-4$$

The eddy current losses are caused by the electric currents induced in a lamination sheet of conductive material when it is subjected to alternating magnetic field. These eddy currents will circulate locally within the material and produce a power that is dissipated as heat due to the conductivity of the material [67]. Eddy current losses can be minimized with stacking sheets by thin slices [64]. However, the use of thinner laminations can increase the cost. The eddy current loss is described by (2-5), where, K_f is the stacking factor and K_e is the eddy current constant.

$$P_e = K_f K_e f^2 B^2 \quad 2-5$$

Based on the curve fitting techniques on the material data provided by the manufacturer or material characterization, the constants such as K_h , K_e and β can be calculated [64].

The estimation of the machine iron losses using equation (2-3) has a limited accuracy [66]. This is due to the assumption that the magnetic flux density in the machine iron core varies sinusoidally [65]. However, when the electrical machine is fed by pulse width modulation (PWM) inverter, the variation in flux density in the stator core is far from sinusoidal. In this case, minor hysteresis loops in the BH curve occur and the iron losses is significantly influenced [65]. Therefore, the iron losses model must include the summation of every hysteresis loop, the majors and the minors. The hysteresis losses equation (2-3) which is based on peak value of the magnetic flux density, is valid only if there is on minor hysteresis loops. To verify the effect of these variations in the flux density distribution on the iron losses. Different iron losses models which include the minor hysteresis loops have been investigated in [65]. In addition, the magnetic flux density in the stator and rotor of the electrical machines can be rotational. This might lead to increase the iron losses when compared to the case of alternating magnetic flux.

Therefore, it is necessary to consider the effect of the rotational magnetic field on the iron losses calculation. This can be achieved by decomposed the rotational flux density at each point into the radial and tangential components. Then the total iron losses are obtained by the sum of the harmonics of the magnetic flux density using the Steinmentz equation [65].

2.5.3 Eddy current losses in the rotor

In high speed applications, the eddy current loss generated in the rotor of PM machines can become really dominant, resulting in very low machine efficiency and also much higher temperatures seen by the rotor components. For PM machine this is a serious challenge as PM material is very sensitive to its operating temperature and can result in deteriorated operation and in partial (or full) demagnetisation in case of sustained operation at high temperature. Therefore, accurate prediction and characterisation of the rotor eddy current losses is imperative for machines in harsh environment, such as that described in Chapter 3. The main causes for these eddy currents in the rotor components are [68]:

- 1) Stator slotting effects where the no load losses are generated due to the airgap reluctance variations caused by the slot opening. This effect could be generally reduced by increasing the effective airgap or reducing the slot opening.
- 2) Space harmonics of the MMF stator due to the winding distribution, which is not synchronized with the rotor and is the major cause of the losses at load condition. This type of losses is dependent of the winding layout selection, where fractional slot machines with concentrated windings have more losses compared to distribute windings in which the loss is almost neglected.
- 3) Stator current harmonics caused by the PWM.

Generally, the eddy current losses can be expressed as given in (2-6), where J_e is the eddy-current density in a conductive volume, σ is the material conductivity and V is the volume of the material.

$$P_{eddy} = \frac{1}{\sigma} \iiint_V J_e^2 dV \quad 2-6$$

Accurate estimation of rotor eddy current losses can be achieved by analytical methods [69-73] or FE analysis [74-81]. For an initial assessment and to provide quick results at the early design stage, rotor losses estimation based on analytical methods are preferred. However, the analytical approaches are based on a number of assumptions and simplifications which may affect the accuracy of the results. For example, ignoring the stator slotting effect, rotor end effects and reaction field of eddy currents on the magnetic field will impose a limitation on the application of the analytical model. On the contrary, more accurate results can be achieved by using 2D and 3D-FE models, which allow to analyse the rotor eddy current losses in PM machines with complex geometry.

2.5.4 Mechanical losses

In general, the most significant mechanical losses in an electrical machine are the windage losses of the rotating parts, in air/fluids, and the contact friction losses in the bearings [82]. In conventional machines these are often smaller with respect to the electrical losses. However, in high speed machines with high power density and forced cooling system, the mechanical losses can become a significant part of the overall power losses.

2.5.4.1 Windage losses

When two surfaces are moving in relation to each other, a phenomenon known as windage loss is created [83]. For rotating EMs, windage losses depend on the velocity of the rotor, where typically the higher the velocity of the surface of the rotor, the higher the amount of air being disturbed by its passing. For most machines, this happens in the air gap. Windage losses are however also dependent on various other factors, including the smoothness (or roughness) of the rotor surface, the air-gap thickness, the slotting configuration of the stator bore and the temperature, pressure and density of the air being effected [84].

Considering a cylindrical rotor rotate in an enclosed stator with an air gap length and laminar flow (i.e. when the fluid flows in the same direction with no disruption) [84], then a shear stress on the rotor surface will be generated. The medium in which the rotor is rotating, which can be liquid (e.g. Water or Oil) or gas (e.g. Air, Nitrogen, and

Helium), will then automatically present a drag force in opposition to that rotation. This drag force produces additional friction torque acting on the rotor for the rotating cylinder. In addition to these forces, heat is generated in the medium due to the relative motion of the fluid that flows between the rotor and stator and as a result, the temperature would be higher at the rotor surface. Generally, the flow distribution inside the air gap of high speed electrical machines is turbulent. Even at low speed operations if there is a roughness on the surface of the rotor or the stator, a measure of turbulence will exist.

As demonstrated in [84], windage losses in the airgap of the rotating machines can simplistically be described by (2-7) where K is the roughness coefficient, ρ_m is the mass density of the fluid, ω is the angular velocity, L_{ax} is the axial length of the rotor, r is the rotor radius, The friction coefficient C_f depends on many factors and it is often empirically determined [84]. For rotating bodies with smooth surfaces K is 1, while for axially slotted surfaces K can be considered to be 2.5 [85].

$$P_w = KC_f \pi \rho_m \omega^3 r^4 L_{ax} \quad 2-7$$

Usually, the nature of the flow inside the air gap is determined by Reynolds number $R_{e\delta}$, which is defined as the ratio between the inertia and viscous forces. Depending on the geometry of the machine, the Reynolds number can be calculated in different ways. For the rotating cylinder enclosed in a stator, the turbulence is described by the Couette Reynolds number [84] as described in (2-8), where L_g is airgap length in the radial direction, μ is the dynamic viscosity of the fluid and u_1 is the peripheral speed of the rotor.

$$R_{e\delta} = \frac{\rho_m u_1 L_g}{\mu} \quad 2-8$$

Most of the parameters for the expression given in (2-7) are known and can be estimated directly. However, the two unknown variables are the friction coefficient C_f and roughness coefficient K , which should be determined accurately in order to estimate the windage losses. Assuming the roughness coefficient of the surface is smooth, the only unknown variable is the friction coefficient C_f . Since the nature of the turbulent flow inside the air gap of high speed machines is complex, the frictional shear stress is

difficult to analyse. This is usually done using an empirical dimensionless friction coefficient C_f which depends on many factors, such as the nature of the flow and quality of the surface. For a rotating cylinder, the friction coefficient can be expressed as described in (2-9), where, τ_1 is the shear stress on the rotor surface (N/m^2) [84].

$$C_f = \frac{\tau_1}{\frac{1}{2} \rho_m u_1^2} \quad 2-9$$

Various literature describes the complexity to calculate analytically the friction coefficient presented in (2-9). Several studies have investigated the friction coefficients, in a cylindrical rotating system in order to provide an accurate model. It is commonly accepted that the most representative modelling can be achieved when using (2-7).

The influence of the windage losses generated inside the airgap have been studied extensively [83, 84, 86-89]. Thus, a lot of empirically-derive or experimentally-derived coefficients for the windage losses calculations are available today. One of the best studies regarding to the windage losses analysis and calculation in recent years is reported in [84]. A qualitative analysis on friction coefficients is presented. This work considers various experimental tests under varying experimental conditions such as different fluid properties, air gap length, speed operations and the roughness on both stator and rotor. Based on the measured losses, a friction coefficient recommendation was developed. This is reported in (2-10) – (2-13).

$$C_f = 5 [(L_g/r)^{0.3}/R_{e\delta}] \quad (R_{e\delta} < 64) \quad 2-10$$

$$C_f = (L_g/r)^{0.3}/R_{e\delta}^{0.6} \quad (64 < R_{e\delta} < 500) \quad 2-11$$

$$C_f = 0.515 [(L_g/r)^{0.3}/R_{e\delta}^{0.5}] \quad (500 < R_{e\delta} < 10^4) \quad 2-12$$

$$C_f = 0.0325 [(L_g/r)^{0.3}/R_{e\delta}^{0.2}] \quad (10^4 < R_{e\delta}) \quad 2-13$$

The above expressions for the windage losses are written under the assumption of no axial flow. When the flow has both tangential and axial flow components, more power is needed for the acceleration of the rotor because, the rotor outer forces the coolant flow into a tangential movement and therefore more power losses [85]. If the air gap length is small compared to the rotor radius, the windage losses due to the axial flow of

the coolant fluid can be determined from (2-14), where q_m is the mass flow rate of the coolant fluid and k_2 is the velocity factor and is equal to 0.48 when both stator and rotor surfaces are rough [84, 85].

$$P_{f_{axial}} = k_2 q_m u_1^2 \quad 2-14$$

However, this factor is expected to be lower than the value of 0.18 if the rotor surface and stator is smooth and to 0.15 in case of smooth rotor surface and rough stator surface (slotted) as the fluid flowing along the rotor is significantly affected by the nature of the surface and rotation of the rotor. Also, the slots of the stator have a positive effect and can decrease the losses associated with the cooling flow through the air gap [85].

The ends of rotor faces also contribute to friction [84, 85]. Considering a cylinder rotating in a free space without a stator, the nature of the tangential flow can be estimated by using the tip Reynolds number, as shown in (2-15).

$$Re_r = \rho_m \omega r^2 / \mu \quad 2-15$$

The friction losses at the end faces of the rotor can then be estimated using (2-16), where $C_{f_{end}}$ is the friction coefficient and r_1 and r_2 are the inner and outer radii of the rotor end [84].

$$P_{f-end} = 0.5 C_{f_{end}} K \rho_m \omega^3 (r_1^5 - r_2^5) \quad 2-16$$

When a rotor end is rotating in free space, the friction coefficient can be calculated as shown in (2-17) – (2-19).

$$C_{f_{end}} = \frac{64}{3 \cdot Re_r} \quad (30 > Re_r) \quad 2-17$$

$$C_{f_{end}} = \frac{3.87}{Re_r^{0.5}} \quad (30 < Re_r < 3 \cdot 10^5) \quad 2-18$$

$$C_{f_{end}} = \frac{0.146}{Re_r^{0.2}} \quad (Re_r > 3 \cdot 10^5) \quad 2-19$$

More accurate results can be achieved by using numerical approaches based on Computational Fluid Dynamics (CFD). These computational techniques can analyse very complex airflow and can therefore provide very accurate results [86]. However, as usual, they require heavy computational resources. The analytical expressions

presented in this section can therefore still be used for the initial estimation of the drag torque and windage losses of EMs.

2.5.4.2 Bearing losses

Bearing losses occur due to the friction that is created between the bearing balls and the internal and the external races of the same bearings. Generally, the bearing losses are depending on many parameters, such as bearing type, bearing diameter, lubrication, load and rotor speed [90]. However, changes in temperature, lubricant viscosity, clearance and/or misalignment in bearings could also influence the friction losses. Therefore, accurate estimation of bearing losses is quite complex, but for an initial estimation it can be normally done via semi-empirical equations, of different complexity [91, 92] or by using specific tools provided by the manufacturers (i.e. SKF) [93].

With the following formula, the bearing losses can be calculated analytically using (2-20), where, k_b is the friction coefficient of the bearing (typically 1-3), m_{rotor} is the mass of the rotor (kg) and n_{rpm} is the rotational speed of the rotor (rpm) [91].

$$P_{bearing} = k_b m_{rotor} n_{rpm} \times 10^{-3} \quad 2-20$$

While the SKF model [93] uses equation (2-21) to estimate the bearing losses, where T_f is total frictional moment (Nm).

$$P_{bearing} = T_f n_{rpm} \times 1.05 \cdot 10^{-7} \quad 2-21$$

In the above expression for the bearing losses, the total frictional moment is separated into four terms as shown in (2-22), where T_{rr} is the rolling frictional moment, T_{sl} is the sliding frictional moment, T_{seal} is the frictional moment of seal, T_{drag} is the frictional moment of the drag losses.

$$T_f = T_{rr} + T_{sl} + T_{seal} + T_{drag} \quad 2-22$$

According to [93], the total frictional moment of the bearing is not constant but depends on the bearing type, speed, temperature and lubrication system. Therefore, more accurate results can be achieved using the SKF model.

2.6 Conclusion

This chapter reviews the different topologies of electrical machines used in aerospace actuators. While each machine has its own advantages and challenges, a PMSM has been selected thanks to its high efficiency and high power density when compared to others. The losses in PMSMs are a very important factor and thus a part of this chapter is dedicated to these losses. Power losses calculation methods were briefly presented, including the theory and calculation methods for determining copper, iron, eddy current and mechanical losses in PMSMs.

The challenges for the high speed and high power density machine design lies in combining the high electrical loading required by the usually limited envelope with the rotor size limited by the inertia specifications. The large electrical loading can lead to considerable amount of losses in the form of heat. High operating speeds cause a significant increase in the machine iron losses, with ac losses are usually more dominant in the stator windings and eddy current losses in the rotor. In addition, the high speeds lead to significant friction loss increase in the bearings and windage losses within the airgap. For this reason, designing an electrical machine towards higher speed, higher power density and lower inertia requires a thorough knowledge of power losses and their influence on the machine's performance.

Chapter 3. INITIAL DESIGN OF THE INDUSTRIAL MACHINE

3.1 Introduction

This chapter covers the design and analysis of a high power density surface mounted PMSM for the direct coupled pump application, as described in Section 1.1. First, the required specifications and design constraints for the application are presented. Using the above information, an initial trade off study on potential machine configurations has been done to identify the most suitable machine topology for the considered application.

The study then started to focus more towards the final design as the target requirements were introduced in a logical sequence. The final design is then geometrically optimised in order to reach the better performing solution. Considering, the high power density and high speed for the required application, thermal and mechanical analyses were used as part of the design stage in order to guarantee a safe machine design and operation.

The overall design process with an effective thermal management was a highlight of this chapter and a key contribution to knowledge.

3.2 Design specifications and constraints

This section presents the main requirements and challenges associated with the PMSM used to drive an EHA for a harsh environment application. The set of requirements is tabulated in Table 3-1. Figure 3-1 illustrates the operating regions of the machine.

Table 3-1: Machine design requirements and performance indicators

Requirements	Values
Torque @ 8700 rpm	10.5 Nm
Torque @ 19000 rpm	5 Nm
Efficiency @ 8700 rpm	$\geq 90\%$
Efficiency @ 19000 rpm	$\geq 85\%$
Dc link voltage	270 dc
Maximum phase current	≤ 85 A
THD of BEMF	$\leq 5\%$
Moment of inertia	$\leq 1 \times 10^{-4} \text{ kg} \cdot \text{m}^2$
Weight	≤ 2.5 kg
Total length	≤ 100 mm
Axial cross section	$\leq 70 \times 70$ mm
Cooling method	Flooded
Coolant flow rate	4.2 L/min @ 90 °C
Maximum winding temperature	210°C
Ambient temperature	90°C

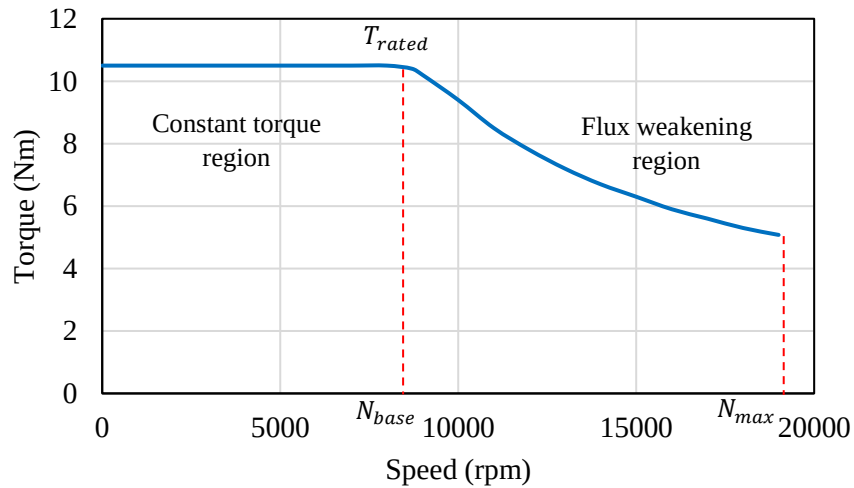


Figure 3-1: Required torque-speed characteristic

From the above, it can be argued that the main design objective related to this particular machine design are as follows.

1) Torque-speed characteristics

A constant torque T_{rated} is required over a speed range from 0 to 8700 rpm. Beyond the base speed N_{base} of 8700 rpm, flux weakening is required until the maximum speed N_{max} of 19000 rpm.

2) High power density

High power to mass and power to volume ratio are key design aspects of the application, due to the space constraints. The total mass of the machine should be less than 2.5 kg, including all active parts: rotor, stator, front/ending bearing, resolver and housing, excluding the cooling fluid mass.

3) High reliability

A directly-coupled pump motor is considered in order to reduce the possibility of the oil leakage and to increase the reliability of the EHA application. These includes minimizing the number of shaft seals between the motor and pump as it is not practical to seal against high hydraulic pressure.

4) Flooded machine operation

For increasing the current density of the motor and due to absence of the shaft seals, the oil of the pumping system can be used to cool the machine. This allows much higher than typical current densities to be used.

5) Windage losses

A flooded cooling system is significantly effective, since the heat can be removed directly from the windings and from the rotor. However, as the rotating velocity of the machine is high, the oil-windage losses at the rotor surface becomes critical in terms of machine efficiency.

6) Fast dynamic response

The motor should have a low moment of inertia to respond quickly to changes in load accelerations.

3.3 Design process

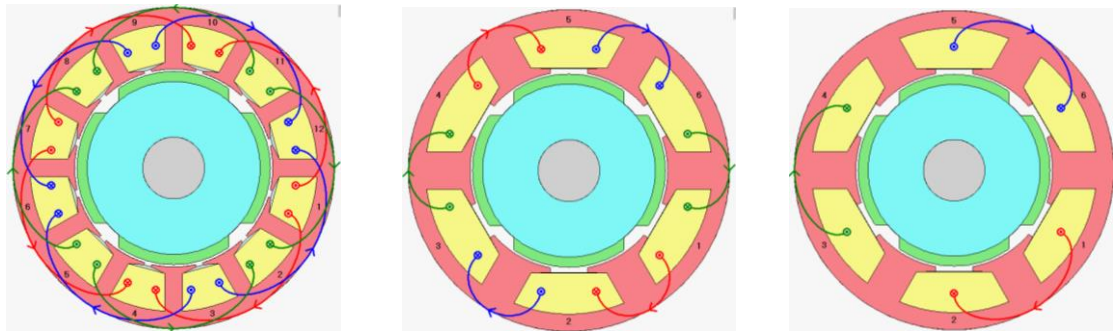
Because of the space and mass constraints of the project, the first step of the design process focuses on the investigation of machines capability in terms of overall the speed range. The main parameters deemed to be important at an initial selection stage are the following:

- 1) Winding configurations
- 2) Slot-pole combination
- 3) Material selection

These points are thoroughly discussed in the following subsections.

3.3.1 Winding configuration

For three-phase electrical machines the winding configuration can be classified into two types, namely distributed (overlapping) and concentrated (non-overlapping) windings as shown in Figure 3-2.



(a) *Distributed winding with $N_s=12$, $p=4$* (b) *Double layer concentrated winding with $N_s=6$, $p=4$* (c) *Single layer concentrated winding with $N_s=6$, $p=4$*

Figure 3-2: Winding configurations of different topologies

Distributed windings such as those with number of slots per pole per phase equal to 1 tend to have larger end windings compared to the machines with fractional slot concentrated windings and hence result in a longer motor for the same stack length. This leads to higher copper losses and manufacturing issues for high pole numbers. On the other hand, they present a good fundamental winding factor and are free of sub-harmonics [44].

Concentrated windings (tooth-wound coils) offer lower motor length for the same core length and lower copper losses due to smaller end windings. Also, the concentrated windings can be built with a segmented stator that provides higher slot fill factors [94]. They also offer a good fault tolerant capability, since the windings are wound around one tooth and therefore the risk of short-circuit between windings of different phases is negligible. However, due to the large MMF harmonic content they can produce a large amount of parasitic losses at high speeds and high electrical frequencies [95].

For this specific application, a concentrated winding configuration has been preferred over distributed windings for two main reasons:

- 1) Due to the dimensional constraints of the machine (stack length).
- 2) Due to the high performance required, it was seen right from the start that significant thermal management techniques would be required. A main technique being considered was the possibility of cooling channels along the axial length of the slots. However, a cooling channel inside the slot of the machines with distributed winding is quite complex. This concept is shown in Figure 3-3 (b). The end-windings overlap to form the phase coil and thus the end winding covers both ends of the slot creating a barrier to the cooling fluid flowing through the channel, making it physically challenging to create a cooling channel with distributed windings.



(a) Concentrated windings

(b) Distributed windings (single layer)

Figure 3-3: Winding layout

3.3.2 Slot-pole combination

Once the winding configuration is defined, the slot/pole number can be determined. The selection of the slot/pole combination is critical for the machine design performance. High-power density machines can be achieved by using high pole numbers. However, this will increase the electrical frequency for given mechanical speed. Since iron losses depend on the frequency, higher iron losses are expected with larger pole numbers [90]. While, to achieve the same power density but with low pole numbers, higher currents are required. Thus, additional copper losses and thermal problems can occur [96]. Therefore, the number of poles should be chosen based on a compromise between a lower mass (higher torque density) and lower electrical frequency (lower magnetic losses).

Considering all the above and prior to the preliminary design, Table 3-2 shows the fundamental winding factor k_w and the airgap harmonic leakage factor σ_δ for different slot-pole combinations of concentrated coils PMSMs [97].

Table 3-2: Fundamental winding factor and harmonic leakage factor for concentrated windings with different slot/pole combinations

$N_s \backslash p$		4	6	8	10	12	14	16
6	k_w	0.866	-	0.866	0.5	-	0.5	0.866
	σ_δ	0.462		4.85	40.1		79.6	22.4
9	k_w	-	0.866	0.945	0.945	0.866	0.617	0.328
	σ_δ		0.462	2.15	3.92	4.85	21.7	104
12	k_w	-	-	0.866	0.966	-	0.966	0.866
	σ_δ			0.462	1.95		4.79	4.85
15	k_w	-	-	-	0.866	-	0.951	0.951
	σ_δ				0.462		2.1	3.59
18	k_w	-	-	-	-	0.866	0.902	0.945
	σ_δ					0.462	1.75	2.15
21	k_w	-	-	-	-	-	0.866	0.89
	σ_δ						0.462	1.69
24	k_w	-	-	-	-	-	-	0.866
	σ_δ							0.462

As can be seen from Table 3-2, higher winding factor can be achieved when the number of slots is close to the number of poles ($N_s = p \pm 1$ or $p \pm 2$). This can lead to improved machine performance and reduced cogging torque [98]. Indeed, to minimize the cogging torque, the least common multiple (LCM) between slot and pole should be high, giving a higher cogging frequency but with lower amplitude [96]. Also, the greatest common divider (GCD) needs to be high, as this typically means lower unbalanced magnetic force due to the symmetry of the electrical machine [96].

On the other hand, slot/pole combinations with higher airgap harmonic leakage factors σ_δ have higher leakage inductances and usually larger field weakening capability, but lower maximum torque at base speed (assuming the voltage limit is the same). However, a higher airgap harmonic leakage factor may also indicate a higher amount of rotor losses resulting from asynchronous current linkage harmonics. Combinations with a very high σ_δ are not feasible at all, as their torque production is considerably limited by very high leakage inductances [99]. Therefore, from the torque and efficiency point of view, the slot/pole combinations with high fundamental winding factor and low harmonic leakage factor are favourite [97].

However, due to the consideration on the PEC switching frequency (16 kHz), a pole number of 8 or less is preferred. In fact, to have maximum torque, machines with fundamental winding factor lower than 0.866 are ignored. Therefore, the remaining feasible configurations (4p/6s, 6p/9s, 8p/6s, 8p/9s, 8p/12s) will be compared in the next section. The analysis is based on some key performance characteristics across the whole speed range using FEA for a more accurate simulation.

3.3.3 Material selection

Selecting the suitable materials for different parts of the electrical machine plays a critical rule in terms of machine performance and satisfying the requirements. Therefore, this section presented the materials used for the PMSM.

3.3.3.1 Lamination steel material

Iron alloy with Silicon (SiFe) is the most widely used material in electrical machines [100]. Generally, the percentage of Silicon content is about 1 to 3%. The Si makes the iron alloy mechanically harder and increases the electrical resistivity and thus lower the eddy current loss [100]. However, increasing the percentage of Silicon has the advantages of increase the resistivity of the material and therefore more reduction in the eddy current loss. But, this will weaken the material and make it brittle in addition to the reduction in the saturation limit and permeability [100].

In case of high power density applications with limited space, Cobalt Iron (CoFe) material with high saturation limit of 2.3 T could be used. The CoFe material are made of approximately 49% Cobalt, 49% Iron and 2% Vanadium. In addition to the maximum saturation value, this material has a high mechanical strength which make it suitable for high-speed applications [100]. However, the main disadvantages associated with CoFe material is its higher price, availability on market as well as the higher mass density 8120 kg/m^3 compared to 7650 kg/m^3 for SiFe material [101].

For the initial machine design, a standard SiFe material namely (M270-35A) with lamination thickness of 0.35 mm and saturation limit of 1.7 T has been selected as iron material for both the stator and rotor laminations. The M270-35A has a good compromise in terms of availability in market, cost, permeability and specific power loss. The magnetization BH curve and the specific power loss of the lamination steel are presented in Figure 3-4.

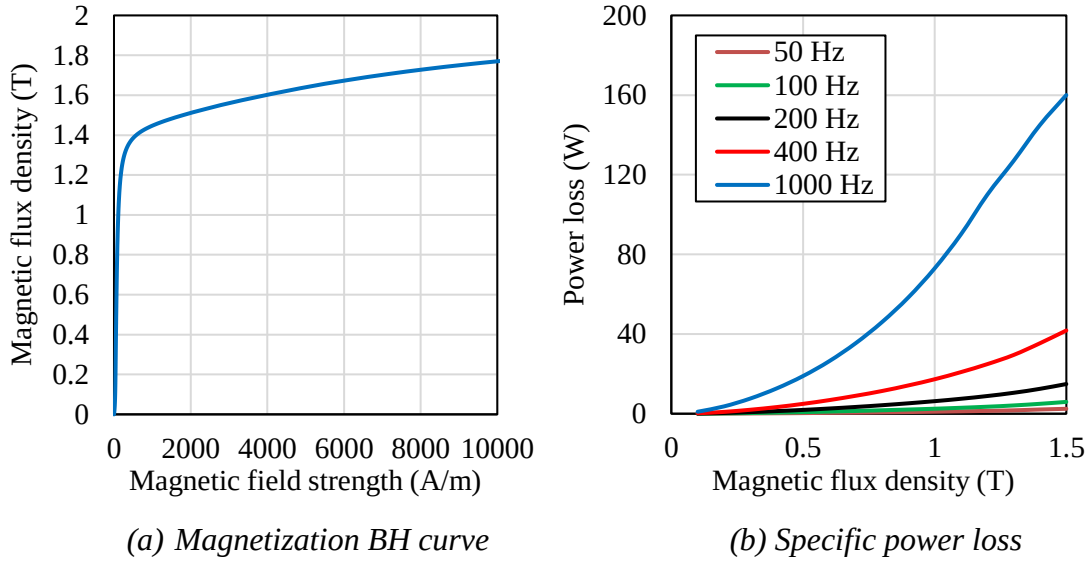


Figure 3-4: Magnetization BH curve and power loss for the M270-35A [102]

3.3.3.2 Permanent magnet material

Selecting an appropriate PM material always requires a careful investigation. The most popular types include Samarium-Cobalt (SmCo), Neodymium Iron Boron (NdFeB), Alnico and Ferrites. Each PM has its own magnetic properties that make it ideal to specific applications. Ferrite magnets have moderate energy density and low cost. From an operating temperature point of view, Alnico provides the highest level among all magnetic materials. However, there are limitation to its performance because it is inherently having a low coercive force H_c [103]. NdFeB PMs have the highest energy density values but their maximum operating temperature is limited to 230°C. From all the PMs available today in the market, Samarium Cobalt material for the PM has been selected due to its lower temperature coefficient (highest stability). The chosen material for this specific application is Recoma 33E grade of ($\text{Sm}_2\text{Co}_{17}$) which has the highest operating temperature and maximum energy product in the market but more expensive [104]. For the application under study, having a flooded airgap machine will lead to higher windage losses and therefore higher temperature at the rotor surface. For this reason, high temperature magnets have been selected to guarantee a safe operation against the PMs demagnetization. The magnet properties are summarised in Table 3-3.

Table 3-3: Magnet properties - Recoma 33E [104]

Characteristics	Values
Magnet remanence B_r	1.17 T
Intrinsic coercivity H_c	1710 kA/m
Coercivity H_c	860 kA/m
Maximum energy product BH_{max}	225 kJ/m ³
Remanence temp. coefficient αB_r	0.035 %/°C
Intrinsic coercivity temp coefficient αi_{BrC}	0.25 %/°C
Magnet relative permeability μ_r	1.068
Max. working temperature T_{max}	350 °C
Density	8300 kg/m ³
Electrical resistivity ρ	90 $\mu\Omega \cdot m$

The demagnetization curves for the selected permanent magnet material at different operating temperatures are shown in Figure 3-5. Assuming that, the temperature of the magnet remains lower than 350 °C and as long as the magnet flux density is higher than 0.3 T the magnets will be safe and will not demagnetize.

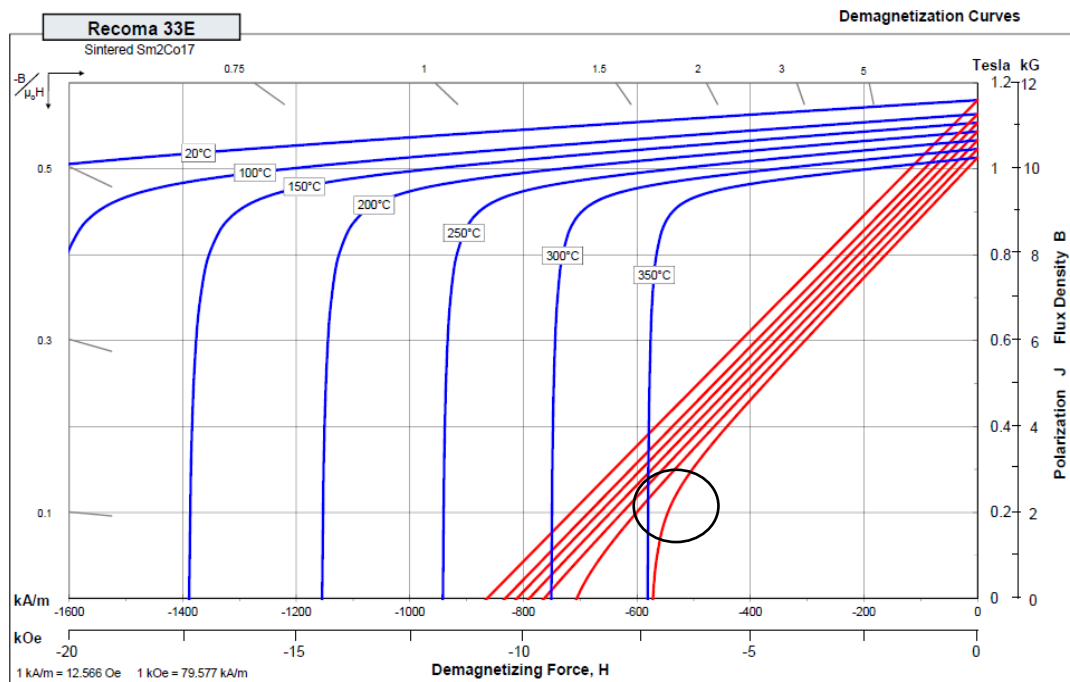


Figure 3-5: Sm2Co17 demagnetization curves [104]

3.3.3.3 Stainless steel material

In general, the stainless-steel materials, which are ferromagnetic, are used for the rotor shaft of the electrical machines. Steel presents high Yield strength, good corrosion resistance and good mechanical properties [47]. As a downside it has a relatively high electrical conductivity and thus eddy currents are free to circulate in the rotor shaft, generating additional losses. The selected material for this application is 17-4PH Stainless-steel which has some magnetic properties such as low remanent flux density and high magnetic field strength. The magnetization BH curve of 17-4PH Stainless-steel is shown in Figure 3-6, while the material properties are summarized in Table 3-4.

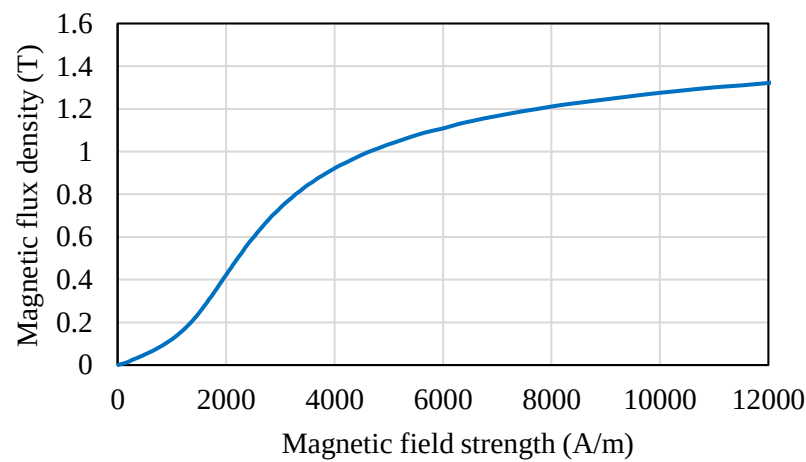


Figure 3-6: Magnetization BH curve for 17-4PH Stainless Steel material

3.3.3.4 Sleeve material

For high speed applications, a retaining sleeve is often adopted to contain the permanent magnets against the centrifugal force. Table 3-4 present the sleeve material properties. Generally, three types of high strength materials are used namely, Carbon fiber, Inconel 718 and Titanium alloy Ti6A14V [105]. Carbon fiber has the highest electrical resistivity and the lowest mass density among the three materials and therefore very small eddy current loss and low mass for the sleeve. But the heat transfer coefficient is very small and thus would results in overheating and demagnetization of the permanent magnets [105]. Both of Inconel 718 and Titanium alloy have a better heat transfer coefficient but their eddy current loss is very high compared to the carbon fiber. Titanium has lower eddy current loss and mass density compared to the Inconel material but also has a higher economic cost [105].

Table 3-4: Material properties

Material properties	Carbon Fiber	Inconel 718	Titanium Ti-6AL-4V	17-4 PH Steel
Density (kg/m ³)	1800	8190	4400	7820
Young's Modulus (GPa)	125	205	110	196
Poisson's ratio	0.28	0.28	0.31	0.27
Thermal expansion (1/K)	0.38e-6	1.3e-5	8.45e-6	6e-6
Thermal Conductivity (W/ (m.K))	0.7	14.7	7.5	19.55
Permitted stress (MPa)	1300	735	550	919
Yield /tensile Strength (MPa)	1960	1100	825	1379
Electrical Resistivity (Ω .m)	2e-5	1.25e-6	1.78e-6	7.7e-7

Since the rotor of the designed machine will be flooded with oil, the Carbon Fiber has not been considered due to the unknown behaviour when immersed in oil. Therefore, the optimal option which provides a compromise between the losses in the sleeve and mass density is the Titanium Ti-6AL-4V.

3.3.3.5 Magnet wire thermal class

Regarding to the specification and requirements presented in Section 3.1, no insulation failure should occur at motor winding for a maximum operating temperature of 210 °C. Generally, the magnet wire is classified according to the temperature class and/or insulation class. For that reason, Class C insulation that can withstand a temperature of up 220°C was selected.

3.4 Initial design considerations

In this section, the initial design parameters and assumptions are used as a starting point of the preliminary design, before moving on to the detailed design, where 2D FE optimization will be used.

- Geometrical constraints

Several geometrical dimensions have been fixed to initial values, such as:

1. The stator outer diameter S_{OD} and stack length L are restricted to the volume constraints given (see Table 3-1), where S_{OD} is 70 and L is 85 mm.
 2. The split ratio S_r (which is set to 60%) is the ratio between the stator inner to stator outer diameters and has a direct effect on the machine's performance due to its relationship with the electrical loading.
 3. The airgap thickness L_g is one of the most important parameters in electrical machines, linked to its magnetic loading. While a smaller L_g tends to improve performance, it is however limited by the manufacturing aspects such as mechanical manufacturing tolerances as well as the stiffness and the dynamic performance of the moving rotor [106]. For this application with the possibility of flooded airgap, L_g is set to 1 mm allowing 0.5 mm for the retaining sleeve.
 4. The magnet thickness M_t and span M_s have a direct effect on the airgap flux density (magnetic loading) of the electrical machine. Using a thicker M_t with full span M_s , results in a higher airgap flux density. However, the volume of PM material and cost is also increased [103]. Thus, to provide an acceptable an airgap flux density, a magnet with thickness of 5 mm and span of 80% is selected.
- Fill factor
Defined as the ratio between the copper area to the total slot area, excluding the slot opening section. Generally, this value can be assumed as 55% for concentrated windings. However, for this application the fill factor will be fixed to 45%. This is due to the fact that, 10% of the slot area was already assigned for the cooling channel, as shown in Figure 3-7.

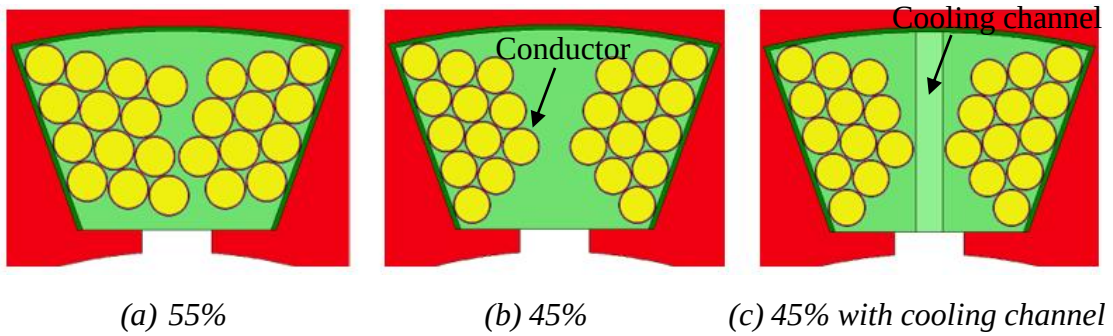


Figure 3-7: Winding distribution with different fill factor

Table 3-5 summarize the main design parameters which have been fixed at the initial design stage.

Table 3-5: Initial design parameters

Parameters	Value	Units
Split ratio S_r	0.6	-
Stator outer diameter S_{OD}	70	mm
Stack length L	85	mm
Magnet span M_s	80	%
Magnet thickness M_t	5	mm
Airgap thickness L_g	1	mm
Sleeve thickness L_s	0.5	mm
Slot fill factor FF	0.45	-
Cooling area C_a	0.1*Slot area	%

3.5 Preliminary machine design

The initial design was based on FE analysis. The FE model of the investigated machines have been created using the commercial package software (Infolytica MagNet). A fully parametrical design tool that is extremely efficient for sensitivity analysis iterations was developed. These tools were then used to investigate various machine configurations in order to find the optimal solution for the application at hand.

3.5.1 Initial comparisons

Based on the design requirements and the initial input parameters presented in Section 3.4, a number of machine designs were compared against each other. A design constraint was set in the form of the output torque, which was kept constant for all designs, so the efficiency was only based on the input power required. The input power is limited by the converter specifications, which are 85 A and 270 V DC. The efficiencies shown comprise the losses of stator and rotor (including end-winding losses). For each design the machine geometrical parameters are optimized to have a maximum average output torque. Figure 3-8 shows the final FE models for the different slot/pole combinations. While, Figure 3-9 presents the flux density distributions of the five studied PM machines, estimated under a full load condition of 10.5 Nm.

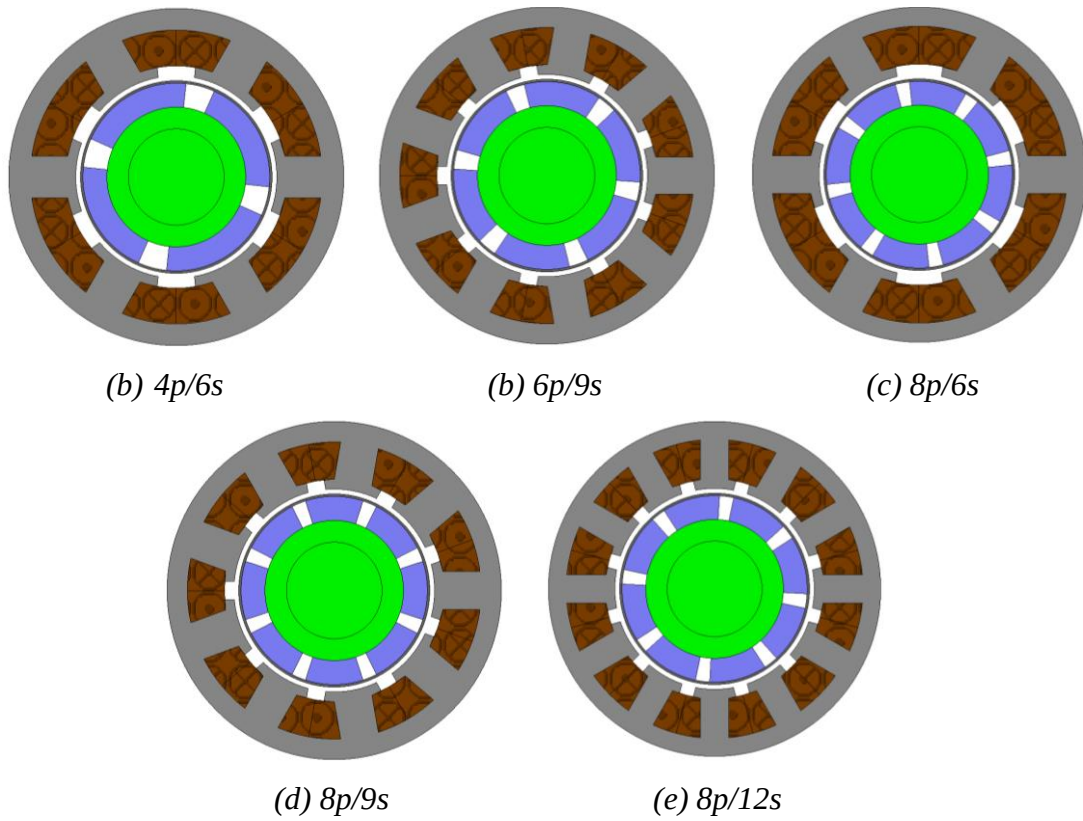


Figure 3-8: FE models of the updated slot-pole combinations

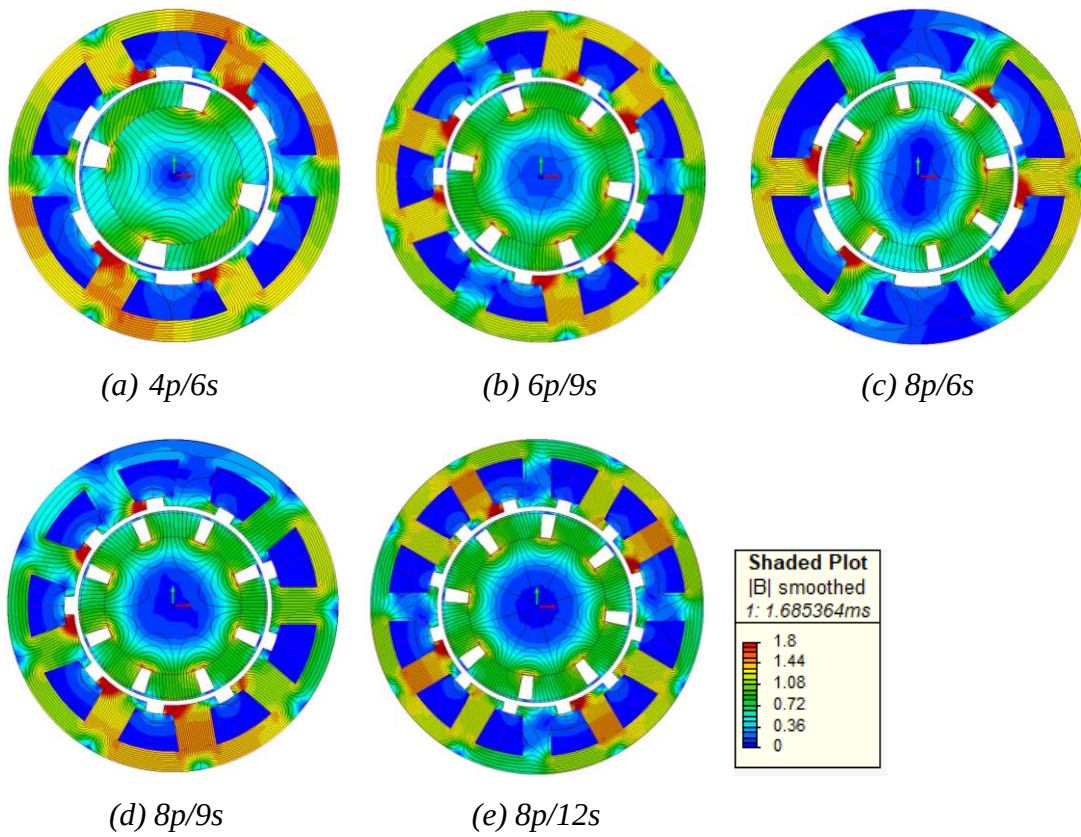


Figure 3-9: Flux density map of the updated slot-pole combinations

As can be observed from Table 3-6, the 4p/6s machine has the highest power losses and the lowest efficiency at the two operating points. While for the 6p/9s design, the efficiency at the rated operating condition (10.5 Nm @ 8700 rpm) is 4.33% lower than the requirements. This configuration also has the highest torque ripple for the maximum speed operation. Among the 8 pole machines, the 8p/9s design has the highest efficiency at the rated operating condition and the lowest torque ripple at the two operating points. While, the 8p/12s design has the lowest THD and the highest efficiency at the maximum speed operation.

As a result, the best overall configurations in terms of efficiency, torque ripple and THD are the 8p/9s and 8p/12s PMSMs. However, it is also true that at this stage, these designs do not meet the efficiency requirements at the rated torque condition (See Table 3-1). Another perceived concern is the rotor moment of inertia which should remain below the limit 10^{-4} (kg.m²), but it can be seen that these machines all exceed this limit. However, it was also perceived that these aspects can be improved during the more detailed design stages. These two configurations are therefore considered for the next design stage, where their thermal management methodology is developed and analysed.

Table 3-6: Initial performance comparison

Parameters @ 8700 rpm	Units	4p/6s	6p/9s	8p/6s	8p/9s	8p/12s
Torque	Nm	10.5	10.5	10.5	10.5	10.5
Current	A	74	69	80.13	74.5	65.5
THD back EMF phase	%	7.06	4.2	3	3.53	1.32
Torque ripple	%	16.37	8.5	4.46	0.89	3.48
Efficiency	%	78.6	86.05	84	88.03	86.83
Total losses	kW	2.6	1.55	1.82	1.3	1.45
UMP	N	0.2	0.15	0.1	493	0.3
Moment of inertia	10 ⁻⁴ (kg. m ²)	1.52				
Parameters @ 19000 rpm	Units	4p/6s	6p/9s	8p/6s	8p/9s	8p/12s
Torque	Nm	6.7	6.19	6.79	6.12	5.96
THD back EMF phase	%	4.8	4	7.42	3.62	1.29
Torque ripple	%	10.9	13.4	7.08	0.74	6.7
Efficiency	%	80.8	87.24	84.2	87.5	88.3
Total losses	kW	3.16	1.8	2.52	1.73	1.57

3.6 Thermal management

Thermal management plays a major role in the design process of electrical machine and needs to be coupled with the electromagnetic design to have an accurate prediction of the temperature and maximize the output power. There are many works which describe various thermal management techniques and thermal modelling of electrical machines [85, 107-112]. The most common methodologies today include analytical lumped parameter analysis or numerical analysis (FEA, CFD). In this work, a commercial thermal software known as MotorCAD software is used. The MotorCAD thermal model is based on the lumped parameter (LP) model. As a compromise between accuracy and the required computational resources, a LP based thermal analysis is an appropriate option for preliminary design purposes. Figure 3-10 shows the thermal network of the investigated PMSM.

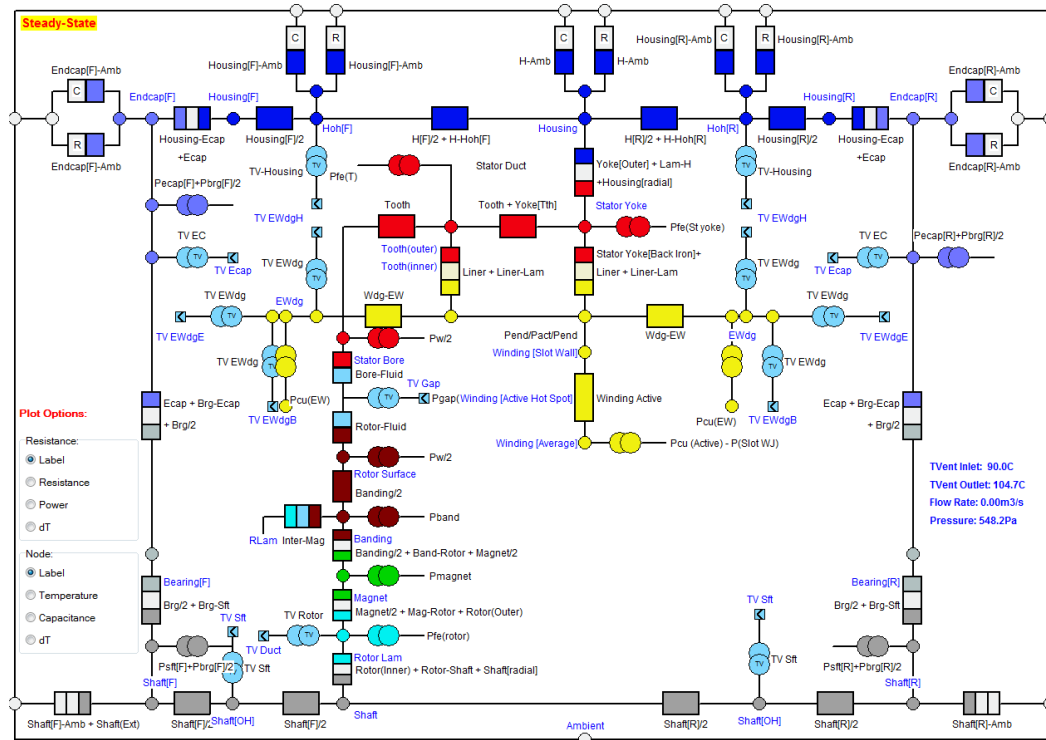


Figure 3-10: Schematic of a PMSM thermal network in MotorCAD

3.6.1 Cooling concepts considered

Various cooling methods can be used to cool electrical motors and the application of the cooling strategy is dependent on the environment and operating conditions of the machine. For a motor-driven pump, the most convenient way to cool the machine is to use the same oil of the pumping system, which can give excellent cooling capabilities without any additional oil supply system and therefore increase the reliability and compactness of the motor [113]. Several cooling methods are considered at the beginning of the thermal management design. These include semi-flooded cooling and completely-flooded cooling [111]. The following sections will give brief introductions about these strategies.

3.6.1.1 Semi-flooded cooling

The main sources of heat in an electrical machine are in general the copper losses and the eddy current losses. Copper loss is generated in the coils, while for a high-speed rotating machine, the main eddy current losses are usually generated in the rotor. A semi-flooded machine cooling scheme, such as shown in Figure 3-11, will enable direct cooling on the stator, i.e. directly on the coils and the end-windings, therefore high current density in the design can be achieved.

Figure 3-11 shows a potential concept of semi-flooded cooling for the machine considered in this study. As can be observed, the motor inner space is divided into two separated chambers by a cylindrical oil sleeve. The space within the sleeve can be called the rotor chamber where all the rotating parts will be located. The rotor chamber is free from oil. While, the space between the motor housing and the sleeve is called the stator chamber, which includes the stator core with machine windings and is completely flooded in oil. Oil from the pump flows from one end to another through cooling channels in the slots between coils and directly removes the heat generated by coils (copper losses) and back iron (iron losses). These cooling channels inside the slots are formed by creating a space between the two coil sides. Although, the implementation of an oil sleeve could add another complexity for manufacturing as well as the possibility of oil leakage, the machine's performance and efficiency are significantly improved by reducing windage losses inside the airgap [113]. However, in case of

significant losses occurring in the rotor, a high temperature will cause a damage to the critical parts of the machine such as the PMs.

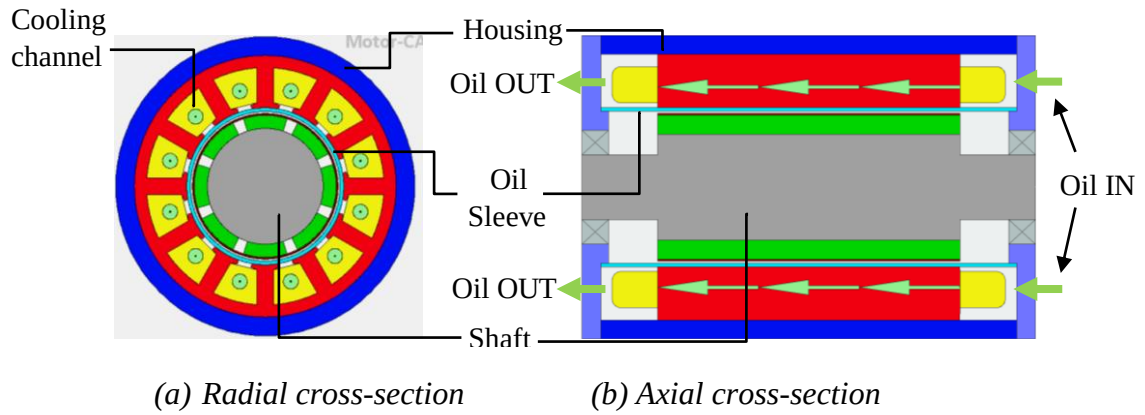


Figure 3-11: Semi flooded cooling arrangement

3.6.1.2 Flooded cooling

Another concept which is investigated in this project is that of having a completely-flooded machine. This is quite similar to the semi-flooded machine; however, it does not separate the machine into two chambers and thus is basically simpler than the semi-flooded machine. However, the main disadvantage related to this configuration is that with a flooded rotor in a liquid, then very high windage losses are generated at the air-gap due to the viscosity of the fluid, which can result in very high temperatures on the rotor surface. Figure 3-12 shows a flooded machine design. The machine is cooled by oil flowing through cooling channels in slots and a gap between stator and rotor.

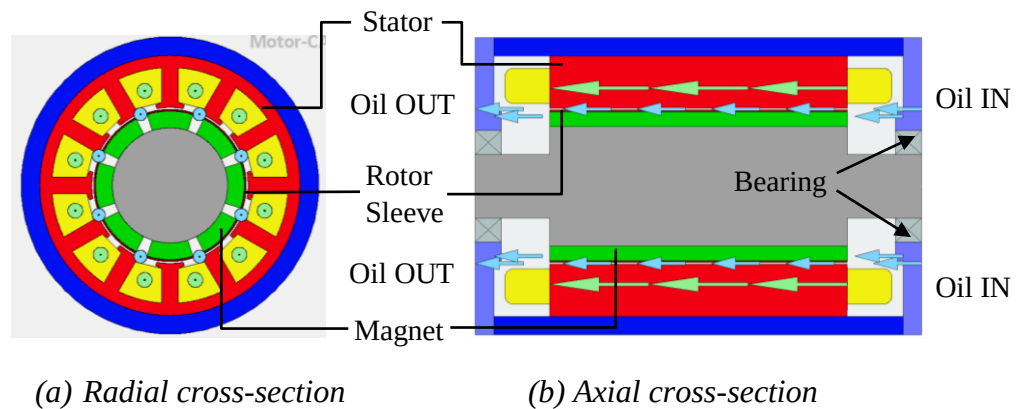


Figure 3-12: Flooded cooling arrangement

3.6.2 Deciding of cooling options

To achieve a better understanding of the two cooling options being considered, a comparative exercise between the two options is done in this section. As the physical dimensions of the 8p/12s and 8p/9s machines described in the previous sections are very similar, then this exercise is done on only one of them, namely the 8p/12s.

The models shown in Figure 3-11 and Figure 3-12, were thus built in MotorCAD. For the non-flooded cooling system, an oil sleeve was implemented to ensure that the rotor rotates in a free-oil chamber and avoid any oil leakage while the stator chamber is flooded with high pressure. A cooling oil flow rate of 4 L/min with an oil inlet temperature of 90 °C, as defined by the industrial project specifications was set. For this analysis, the main result that was considered important was the windage loss and this was determined for several operating speeds.

The obtained results are shown in Figure 3-13. As expected, the windage losses are heavily influenced by the operating speed. For the maximum speed of 19 krpm, the recorded losses are approximately 686.9 W and 1.62 W for the flooded and non-flooded airgap, respectively.

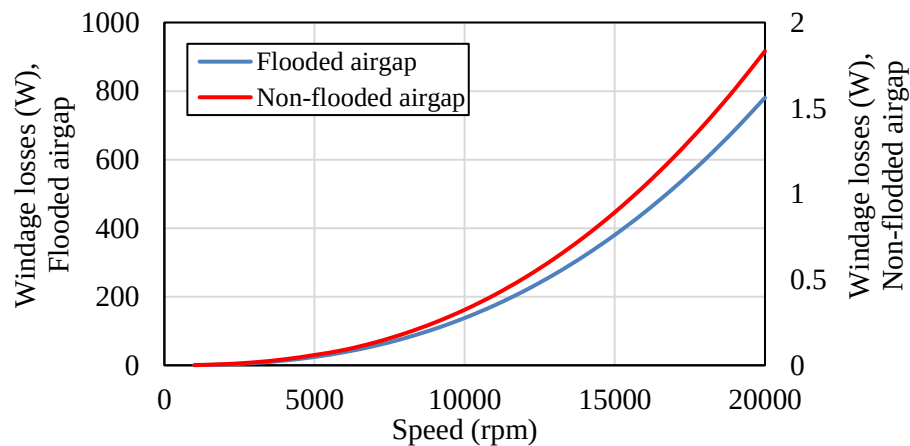


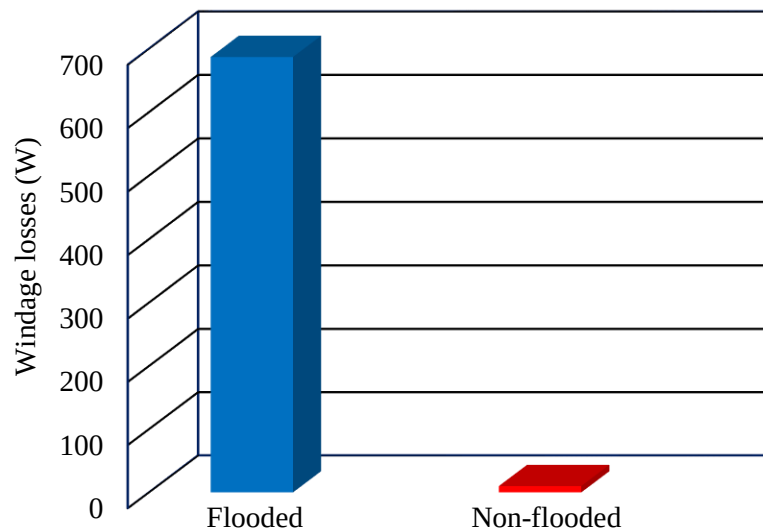
Figure 3-13: Windage losses versus speed

Figure 3-14 compares the two different cooling systems. From these results and those of Figure 3-13, it can be perceived that the semi-flooded cooling option, i.e., when the rotor is rotating in air, is better in terms of total windage loss. On the other hand, this configuration does result in a much higher temperature for the PMs, which could lead to demagnetisation and therefore a deteriorated performance.

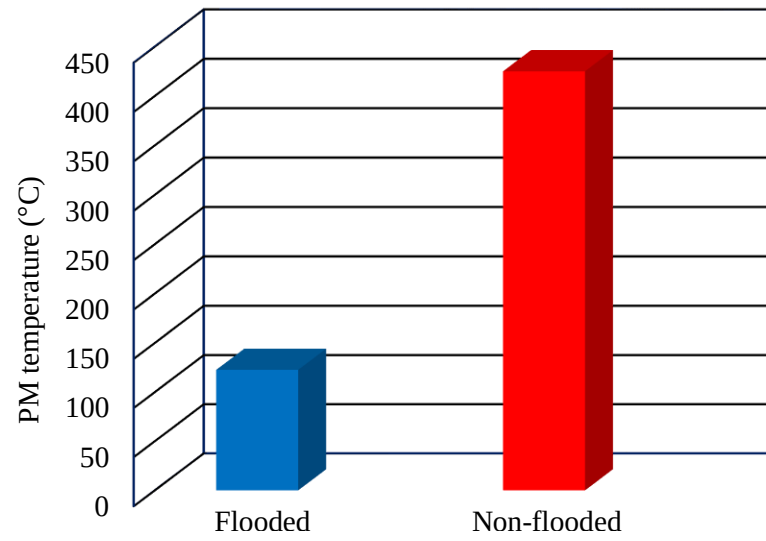
The application of the semi-flooded cooling system with oil sleeve also resulted in a 12.76% reduction in the output torque compared to the flooded system as shown in Figure 3-14 (c). This is mainly due to the 23.3% increase in airgap thickness caused by the additional oil sleeve.

For the relatively high-speed machine being considered here, where rotor losses are expected to be quite high, then a non-flooded airgap has a significant drawback on the overall machine performance due to the high temperature inside the PM. Such a high temperature is beyond the capability of any magnet grade. In addition, more current is required for the non-flooded cooling system to achieve the torque requirements and thus will lead to further reduction in the efficiency and thermal issue.

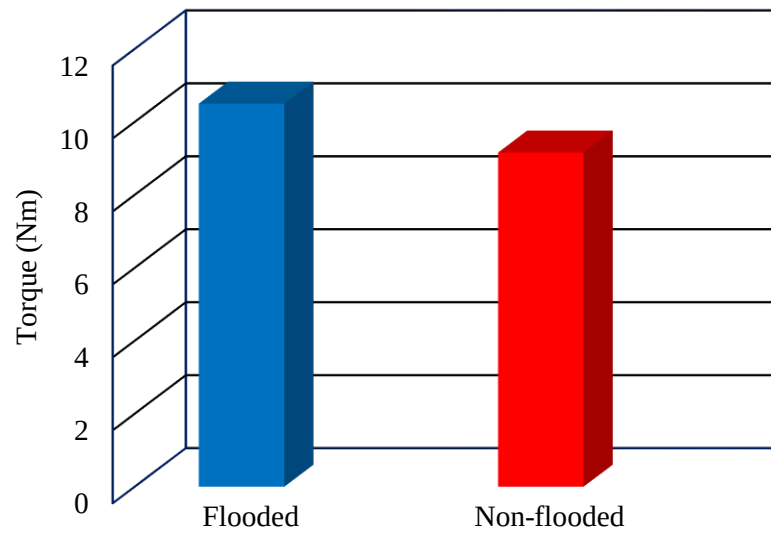
Because of these reasons, it was decided to adopt a flooded cooling system, as shown in Figure 3-12.



(a) Windage losses



(b) PM temperature



(c) Torque

Figure 3-14: Windage losses, temperature and torque with different cooling system

3.7 Initial verification

Once the general cooling system was decided, then a more detailed thermal analysis was conducted for both machine contenders, i.e., for the 8p/9s and 8p/12s PMSMs. Based on simple analysis in terms of liquid flow relative to thickness of the conduit, then the space for the cooling channel was chosen to be 10% of the total slot area. For the initial design, the length of the cooling channel was selected to be equal to the slot depth. Figure 3-15 show the cross-section area of the PMSMs equipped with an ideal cooling channel at the centre of each slot.

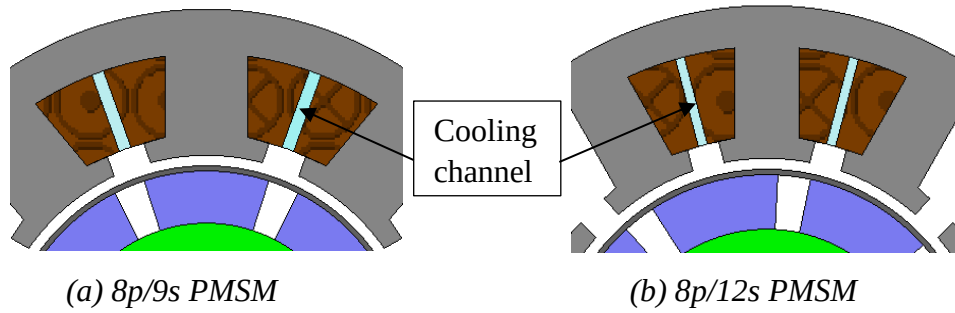


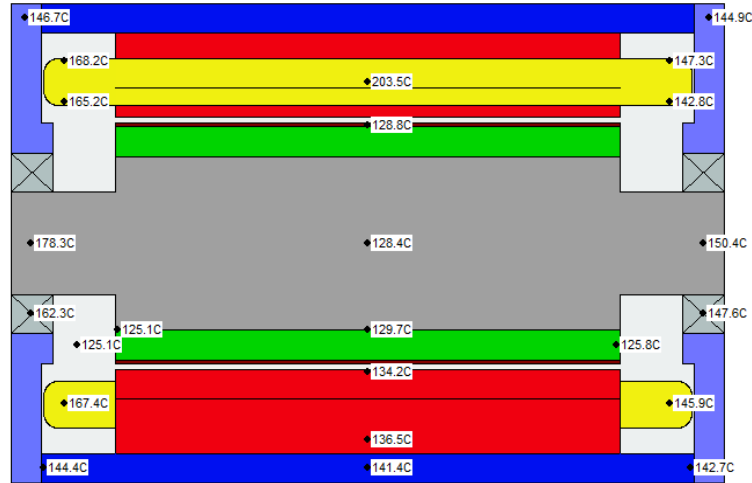
Figure 3-15: Cross-section area of the PMSMs with cooling channel

For the worst operating full load case, at 19000 rpm, the loss distribution is listed in Table 3-7. The copper and windage losses are still the dominant components in terms of the high-speed efficiency, as shown in Table 3-7. However, from these results it can be perceived how the magnetic field of the 8p/9s machine has significant space harmonics, which contribute to higher losses in the rotating components (i.e. magnet, sleeve and solid shaft) [114]. While, higher copper losses are observed in the 8p/12s design, as shown in Table 3-7. This is mainly due to the increase of the current density to produce the desired torque.

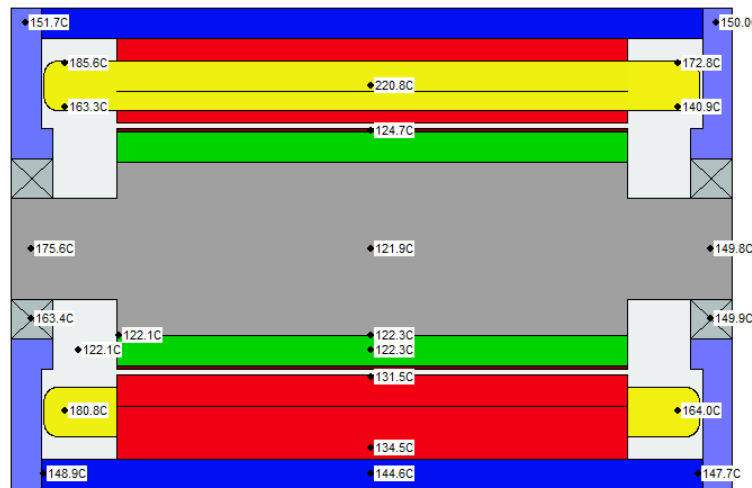
Table 3-7: Losses distribution of PMSMs at 5 Nm and 19000 rpm

Parameters	8p/9s	8p/12s
Stator iron losses	73.1	68
Copper losses	1190	1375
Rotor losses	472.6	130.9
Windage losses	686.7	
Total (kW)	2.42	2.26
Efficiency (%)	83.41	83.9

From the above comparison, the temperatures inside the machine, predicted by the lumped parameter thermal model, are shown in Figure 3-16. As can be seen, for a limited oil flow rate of 4 L/min at a temperature of 90°C, the winding average temperature for the 8p/9s and 8p/12s PMSMs are 194.8°C and 219.2°C, respectively. The PM temperature is under 130°C which is well below the magnet temperature limit (350°C) for both designs.



(a) 8p/9s PMSM



(b) 8p/12s PMSM

Figure 3-16: Temperature distribution in the machines

Table 3-8: Thermal comparison between the 8p/9s and 8p/12s PMSMs

Machine topology		8p/9s	8p/12s
Housing		141.2	144.6
Magnet		129.9	122.7
Stator tooth		136.45	134.9
Copper	Maximum	208.4	259.3
	Minimum	141.9	136.18
	Average	194.8	219.2

Table 3-8 compiles the results from Figure 3-16 in one table for better visual comparisons. Even though the total losses were higher in the 8p/9s machine, the higher copper losses of the 8p/12s machine were considered to be more critical. The ensuing heat from these losses (i.e. copper losses) is more difficult to remove from the windings of the 8p/12s machine, even with a flooded cooling system. When considering all these aspects, it was decided that the 8p/9s machine is the better performing option for the application at hand, even though this comes with the known challenge of the relatively high unbalanced magnetic pull (UMP).

3.8 Mechanical-rotor model

As the rotational speed increases, the mechanical stresses acting on the rotor structure, due to the centrifugal force and modes of vibration, must be considered as part of the initial design stage in order to guarantee a safe machine design and operation [115]. Indeed, the PMs are mechanically weak and brittle [116], hence a retaining sleeve is required for maintaining the circumferential stress below the yield strength of the PM material at maximum operating speed. The sleeve is usually fitted (or wound, if carbon fibre is used) over the PMs [115]. In practice, sleeves are normally applied with large interference fit to achieve the required retention. However, when the rotor is in standstill or at low speeds, the high interference fitted sleeve results in high compression force, which axially stresses the PMs. Therefore, the retention sleeve thickness and rotor stress must be calculated precisely. Otherwise the excessive stress can cause serious damage to the rotor [117].

Generally, the mechanical stresses are highest at the contact region between the PM and the sleeve and between the PM and the shaft [115]. These stresses are affected by the centrifugal force due to the rotation of the rotor and the compressive force due to the interference fit and can be calculated by using the equivalent (von Mises) stress as [118].

$$\sigma_v = \sqrt{\sigma_c^2 + \sigma_c\sigma_r - \sigma_r^2} \quad 3-1$$

It is worth noting that both circumferential (σ_c) and radial (σ_r) components contribute to the equivalent (von Mises) stress and based on (3-1), a mechanical stress analysis can be carried out on the PM and the sleeve. The mathematical formulation of the σ_c and σ_r are explained in detail in Appendix D.

3.8.1 Design results

A study to investigate the stresses in both of PM and sleeve is given in this section. An analytical model was developed and used for the mechanical stress evaluation as to the recommendations in [115, 119, 120]. Based on recommendations from rotor assembly manufacturers, the sleeve thickness was set to minimum thickness of 0.4 mm for manufacturability reasons. The circumferential and radial stresses have been calculated,

and the results are presented in Figure 3-17 for different rotor radii and magnet thicknesses considering both standstill and maximum rotational speed.

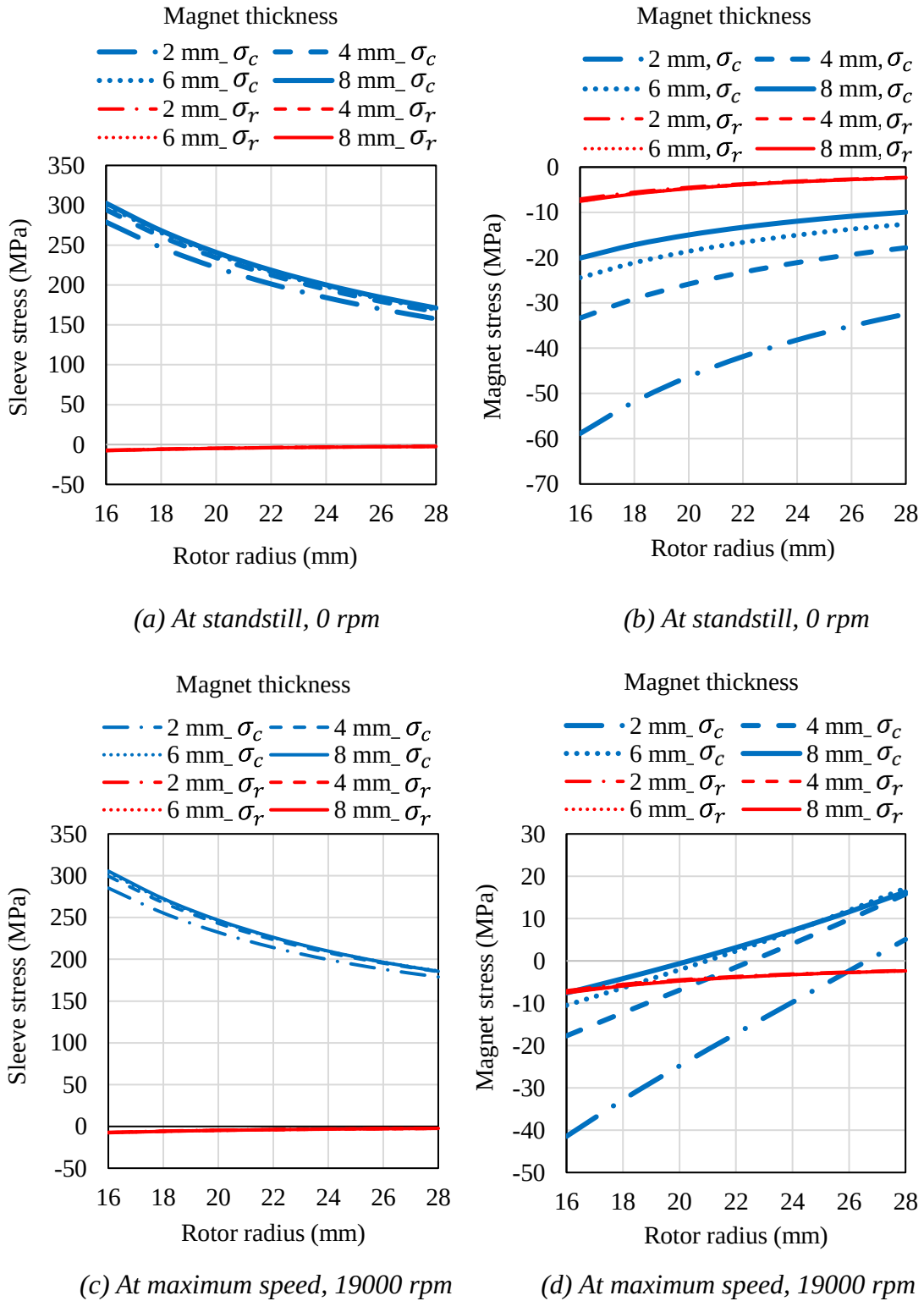


Figure 3-17: Stresses in permanent magnet and sleeve

At first, focusing on the outer rotor radius, it can be seen that larger outer radii lead to lower mechanical stresses in the sleeve, as shown in Figure 3-17 (a). This is mainly due to the reduction of the compression force, which could also reduce the PM contact pressure. Furthermore, although a PM contact pressure benefits from a high interference fit to ensure that the PMs remains in the compression, it can be clearly seen that increasing the outer radius of the rotor leads to lower compression force in the circumferential direction of the PM, as highlighted in Figure 3-17 (b).

For the titanium sleeve, the maximum tangential stress is 300 MPa at standstill and 305.5 MPa at maximum speed, as shown in Figure 3-17 (a) and (c). This is mainly due to the compressive force caused by the interference fit. In contrast, only 5.5 MPa is added to the maximum stress of the sleeve when the rotor rotates at maximum speed. Figure 3-17 (b) and Figure 3-17 (d) show the mechanical stress of the PM at standstill and maximum rotational speed, respectively. At a standstill, the PMs are substantially under compressive stress only. The higher compressive stress is below the permissible compressive strength of the PM, which is 800 MPa. On the other hand, the mechanical stress in the PM increases significantly at maximum speed. This is expected because of the increase in the centrifugal force caused by the rotation.

Therefore, in order to ensure a safe operation, the Von Mises stress σ_v calculated from (3-1) must be less than the total yield strength of the selected materials for maximum speed. Figure 3-18 shows the equivalent (Von mises) stress in the PM and the sleeve. With the information provided in Figure 3-18, an accurate estimate of the rotor's design parameters to satisfy the stress requirements and avoid any mechanical failure.

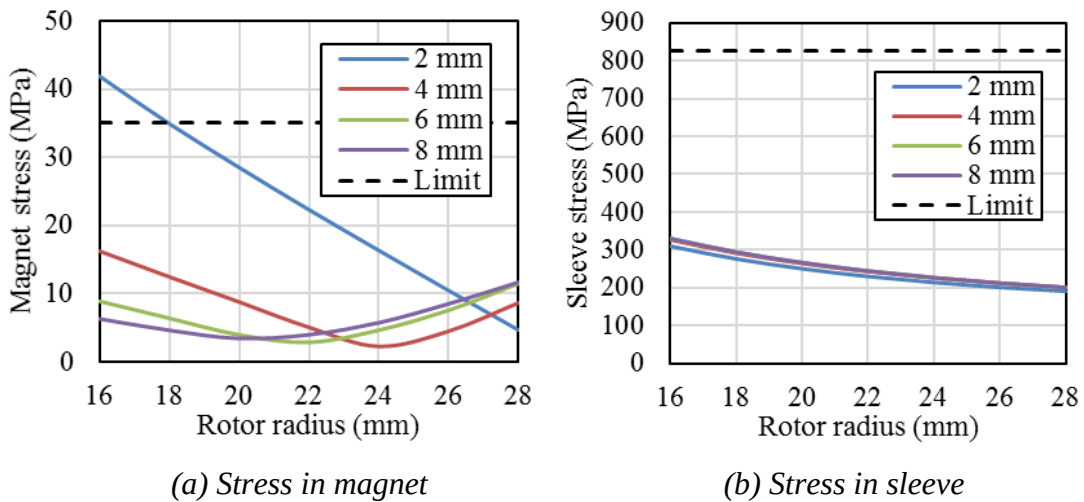


Figure 3-18: Stresses in permanent magnet and sleeve at maximum speed

3.9 Motor sizing

Having decided the optimal slot-pole combination (chosen from Section 3.7) and cooling system, the next step of the design process focuses on the fulfilment of the torque for the specific space available while meeting all the other requirements and constraints. For this reason, an initial sensitivity analysis on the main geometrical parameters of the 8p/9s machine has been carried out with the aim of defining a preliminary machine design. The considered parameters are shown in Figure 3-19.

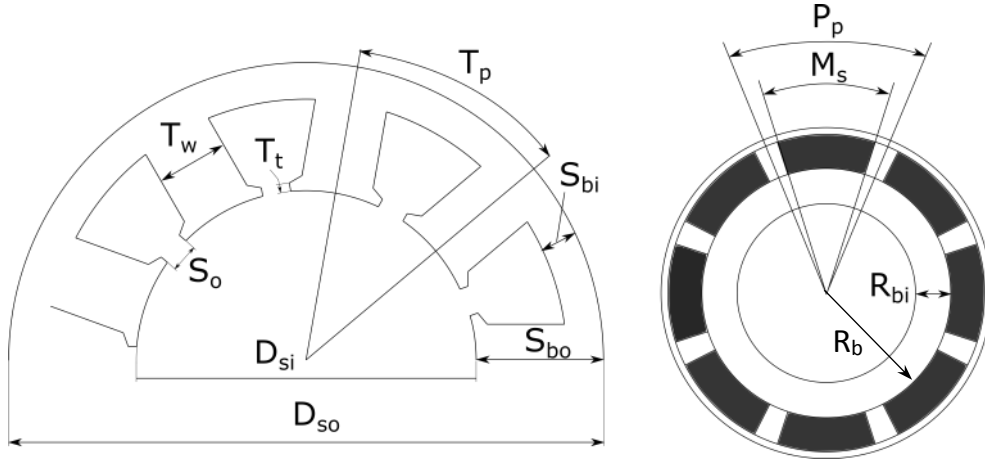


Figure 3-19: Cross-section of the 8p/9s PMSM

The main geometrical dimensions that are most influential on the performance of the 8p/9s machine, are the split ratio (D_{si}/D_{so}), relative tooth width (T_w/T_p), relative stator back iron (S_{bi}/S_{bo}), relative magnet span (M_s/P_p), relative slot opening (S_o/T_p), relative rotor back iron (R_{bi}/R_b).

In the work defined in the next sections, it is important to note that all sizing and sensitivity analyses procedures comprise a set condition of a fixed torque of 10.5 Nm.

3.9.1 Rotor dimensioning

The choice of the rotor diameter is critical for this application to ensure the low inertia and the fast-dynamic response. The machine is sized first based on initial design considerations discussed in Section 3.3. The rotor inertia was 34.2% higher than the requirements presented in Table 3-6. Although, the inertia is directly proportional to rotor length, a significant reduction can be achieved by minimizing the rotor radius as shown in (3-2).

$$J_r = \frac{\pi}{2} \rho_{fe} r^4 L_{ax} \quad 3-2$$

For this reason, a feasibility study has been completed to determine the limit on the rotor size. The rotor end-shaft and bearings are fixed mechanical components and for the case of this work, they don't have any design parameters. In this study, a factor of 10% may be added to the total inertia to include the non-active components such as bearings and end-shaft. The results are shown in Figure 3-20 for a rotor with a fixed stack length of 85 mm (chosen from Table 3-5). At 17.825 mm rotor radius, the inertia reaches its 0.9 of the limit. Indeed, this value has been identified as an optimal solution in this study.

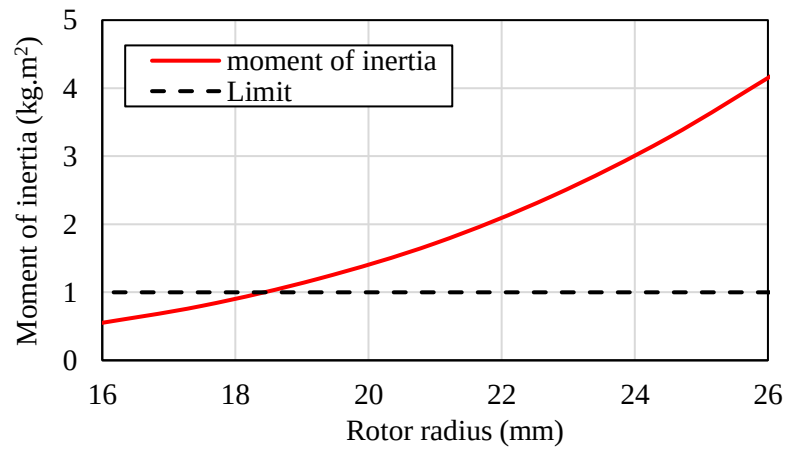


Figure 3-20: Rotor moment of inertia versus rotor diameter

3.9.2 Airgap and PM thickness

Generally, the thickness of the airgap is calculated based on manufacturing considerations such as manufacturing tolerances and rotor dynamics [121]. However, the airgap thickness L_g represents a critical parameter, since it affects the output torque and efficiency. For this study, a trade-off analysis has been performed to carefully select the airgap thickness value and the results are summarized in Table 3-9. To maintain the rated output power, the magnet thickness was increased with the increase of the airgap thickness. However, with an increased airgap (case 4 and 5), for the machine to have the same output torque, the electrical loading must increase accordingly, and hence this results in more copper losses as presented in Table 3-9.

The flux density distributions inside the airgap, with no load, armature reaction and full load conditions are shown in Figure 3-21 (a), (b) and (c). It can be noted that an increase

in airgap thickness reduces the flux density harmonics caused by the stator slotting effect and therefore less dips in the waveforms as shown in Figure 3-21 (a). Also, with the increase of airgap thickness, the flux density space harmonics due to armature reaction would significantly reduce due to the increase of magnetic reluctance, Figure 3-21 (b). Figure 3-21 (d) shows the harmonic components corresponding to Figure 3-21 (c), which confirm the above analyses.

Besides, the distribution of eddy current losses in the rotor components (i.e. magnet, sleeve and shaft) for different airgap thicknesses is shown in Figure 3-21 (e). Effectively increasing the airgap thickness results in lower losses in the rotor, due to the reduction of the harmonics as shown in Figure 3-21 (d). Figure 3-21 (f) shows the predicted windage losses for different speeds and airgaps. From the results, it can be seen that the windage losses increase with the operating speed increase whilst reduce with an increase in airgap thickness.

Therefore, increasing the airgap thickness is important to improve the efficiency. However, an air gap which is too big would require also more current in order to push the same amount of flux across the airgap and therefore more copper losses and lower efficiency. Finally, a 0.9 mm airgap was chosen for the design. For this airgap thickness, a PM with thickness of 4 mm was required to generate the desired torque of 10.5 Nm.

Table 3-9: Machine performance for different airgap thicknesses

Parameters	Case 1	Case 2	Case 3	Case 4	Case 5
Airgap thickness (mm)	0.5	0.7	0.9	1.1	1.3
Sleeve thickness (mm)	0.4	0.4	0.4	0.4	0.4
Magnet thickness (mm)	2.3	3	4	7	8
Stator inner diameter (mm)	37.45	37.45	37.45	37.45	37.45
Current (A)	85	85	85	87	92
Efficiency (%)	83.5	84.8	85.6	85.5	83.7
Total losses (kW)	1.88	1.71	1.6	1.61	1.82
Copper losses (W)	1298	1298	1297	1375	1587
Windage losses (W)	83	75.1	69	64.5	59.6
Stator core losses (W)	78.9	73.4	69.5	67.5	68.7
Rotor losses (W)	420.1	269.5	171.2	104.5	97.2
Moment of inertia ($10^{-4}(\text{kg} \cdot \text{m}^2)$)	1.14	1.067	1	0.924	0.88

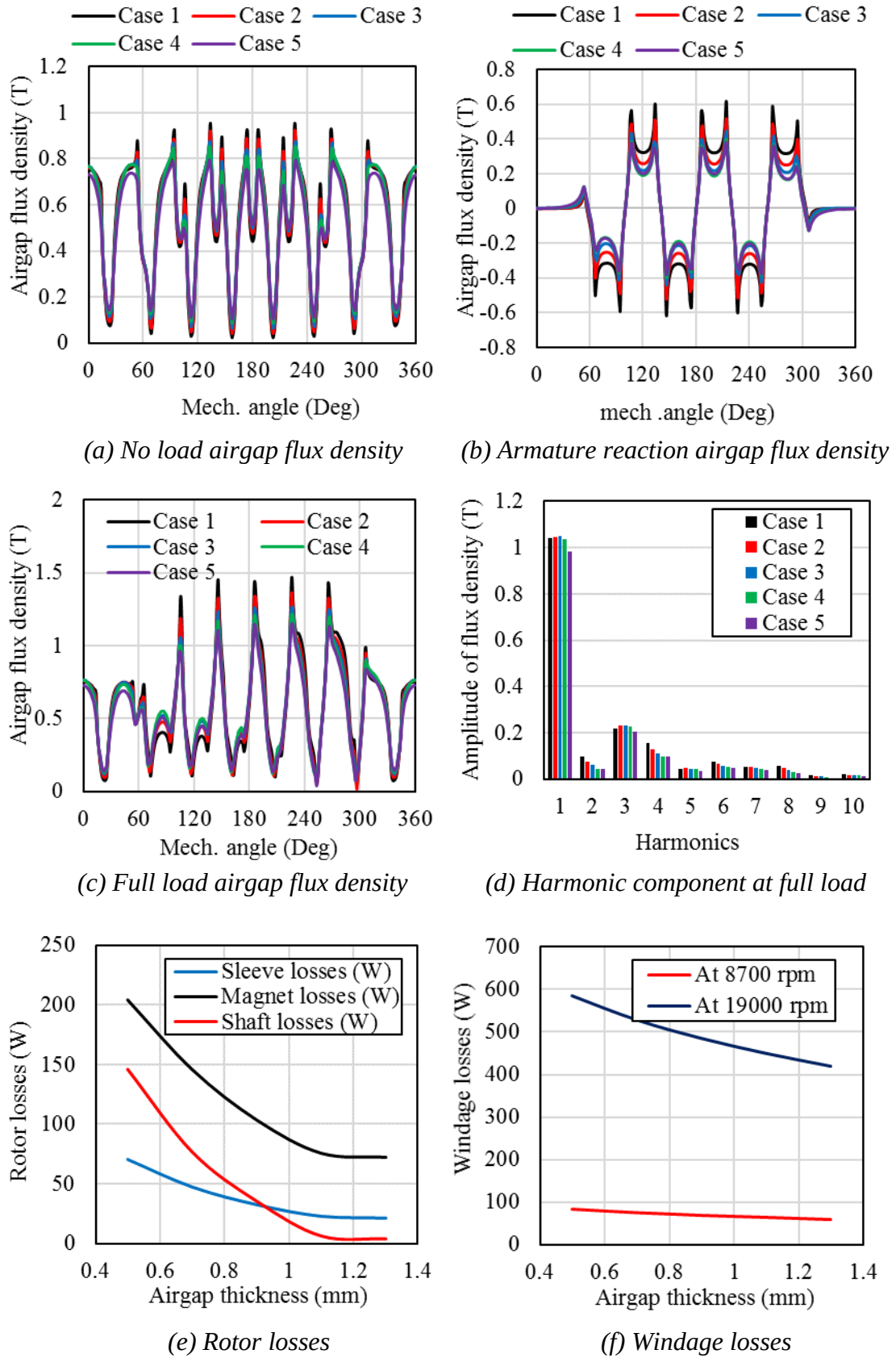
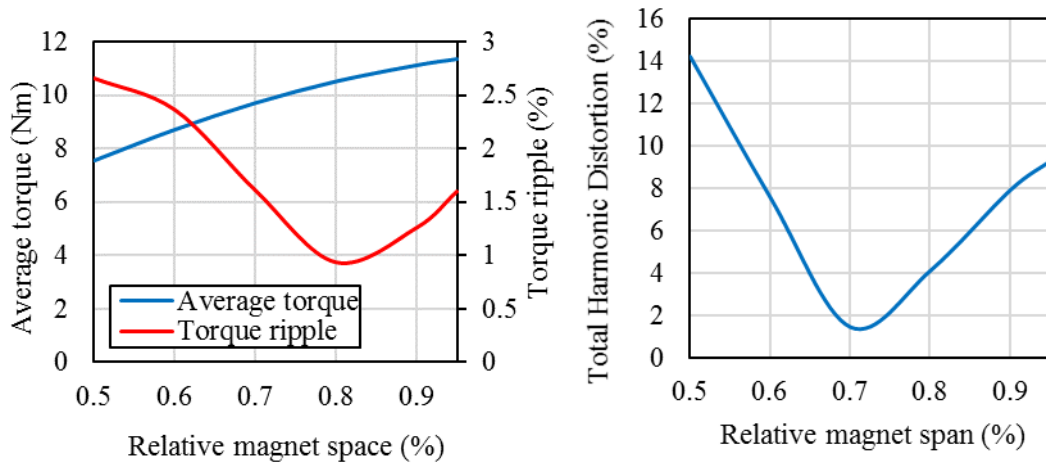


Figure 3-21: Influence of airgap and magnet thickness on

3.9.3 Relative magnet span

The next parameter to be optimized was the permanent magnet span, which mainly affects the average torque, torque ripple and the harmonic content of the back electromotive force. For the SMPM machine, the optimization results are summarized in Figure 3-22. The magnet span which provides a good compromise between torque ripple and total harmonic distortion is equal to 80% of the pole pitch. It is important to underline that choosing the magnet span value equal to 70% of the pole pitch will allow to reduce the total harmonic distortion of the back electromotive force from 4.09% ($M_s=80\%$) to 1.42% ($M_s=70\%$). However, the total harmonic distortion reduction will affect both developed torque and torque ripple. Indeed, adopting a 70% permanent magnet span will result in lower average torque and higher torque ripple.



(a) Average torque and torque ripple

(b) Total harmonic distortion

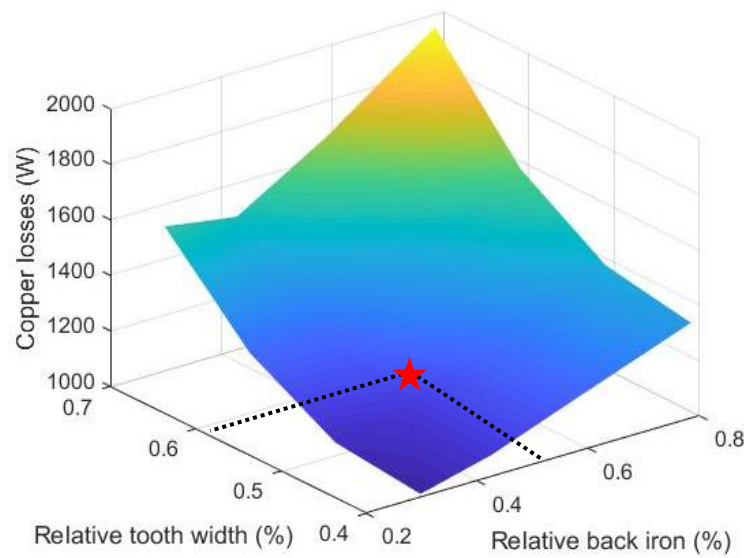
Figure 3-22: Influence of relative magnet span on

3.9.4 Relative tooth width and stator back iron

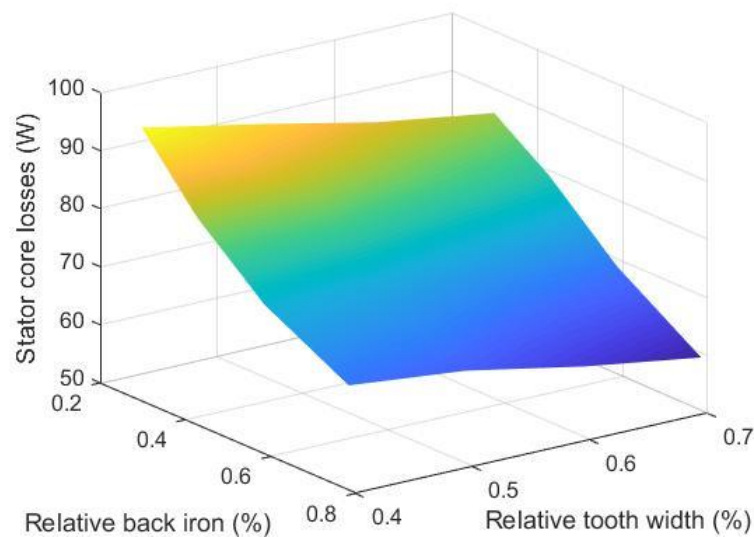
The stator tooth width and back iron have very strong effect on efficiency, as they represent the balance between the electrical loading and the magnet loading. The influence of the tooth width and back iron have been investigated for the fixed output torque of 10.5 Nm. Figure 3-23 shows the results for this analysis. For the rated speed of 8700 rpm, the copper losses and stator core losses are presented and discussed. From Figure 3-23 (a), it can be seen that, the copper losses increased with decreasing slot area, either via larger tooth width or back iron. As the slots get wider, the copper has more space and the resistance of the coils decreases. However, thinner teeth usually

imply more saturation in the magnetic circuit and higher iron losses as shown in Figure 3-23 (b). This increase in the stator core losses was due to the iron area saturating.

The lowest copper losses were achieved for a tooth width span of 40% and back iron span of 30%. However, higher current (95.5 A) is required to achieve the required torque. For this reason, a tooth width span of 55% and back iron span of 58% have been selected since this value provide the required torque (10.5 Nm) along with an acceptable value of copper losses.



(a) Copper losses variation with tooth width and back iron



(b) Stator core losses variation with tooth width and back iron

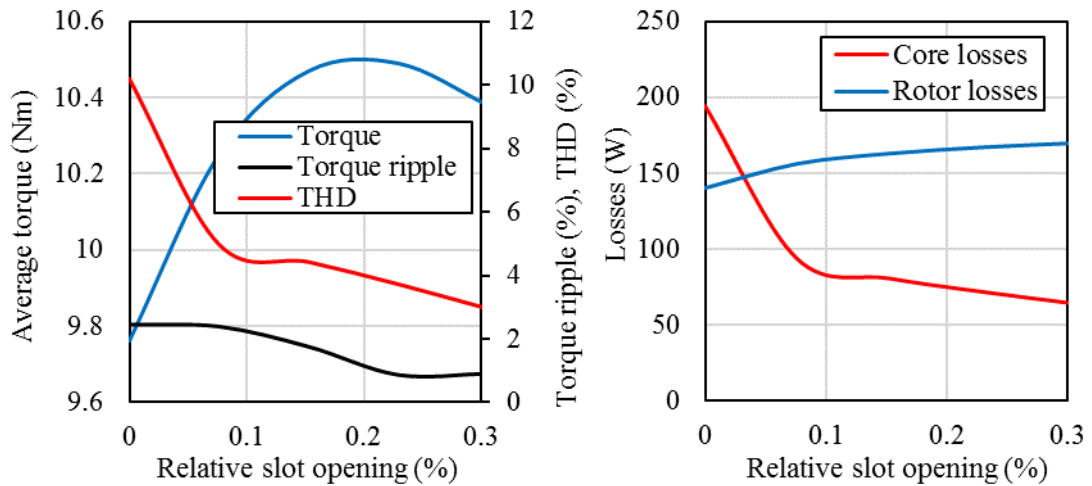
Figure 3-23: Variation of the losses with tooth width and back iron at rated speed

3.9.5 Relative slot opening

The stator slot opening of the PMSM has a direct effect on the leakage flux and therefore may have a significant impact on the flux paths and saturation within the tooth tip [74]. Therefore, the slot opening is optimized in order to achieve the desired torque as well as to minimize the rotor eddy current losses and stator core losses which are critical at high speed. As shown in Figure 3-24 (a), the slot opening significantly affects the output torque. For small slot openings, the leakage flux through the tooth tips increases and thus reduces the flux linkage and the output torque. On the other hand, when the slot opening increases, the slot leakage flux increases too and thus leads to a reduction in torque.

Figure 3-24 (b) shows the eddy current losses in the rotor component and the stator core losses as a function of slot opening at rated speed condition of 8700 rpm. Due to the increase of the slot opening, a significant increase in the rotor losses were noted due to the increase of the flux harmonics created by the stator slot opening. On the other hand, it can be observed that the stator core losses increase as the slot opening decreases. This is because of the increase in the iron losses in the tips of the stator tooth due to saturation.

In relative terms, the optimal slot opening value is equal to 22% of the slot pitch, which is equivalent to 3 mm, in absolute terms.



(a) Average torque, torque ripple and THD

(b) Rotor and stator core losses

Figure 3-24: Influence of relative slot opening span on

3.9.6 Relative rotor back iron (hollow shaft)

At this stage of the design, considering a solid shaft, the total mass of the electrical machine including the non-active parts is 2.86 kg. The total moment of inertia is equal to $1.01 \cdot 10^{-4} \text{ kg.m}^2$. Therefore, to meet the project requirement, the use of a hollow shaft was considered. The rotor back iron value has been optimized in order to avoid the magnetic saturation of the flux paths and to ensure the mechanical strength. As shown in Figure 3-25, the adoption of a hollow shaft has a direct effect on the rotor mass of the machine as well as the total moment of inertia.

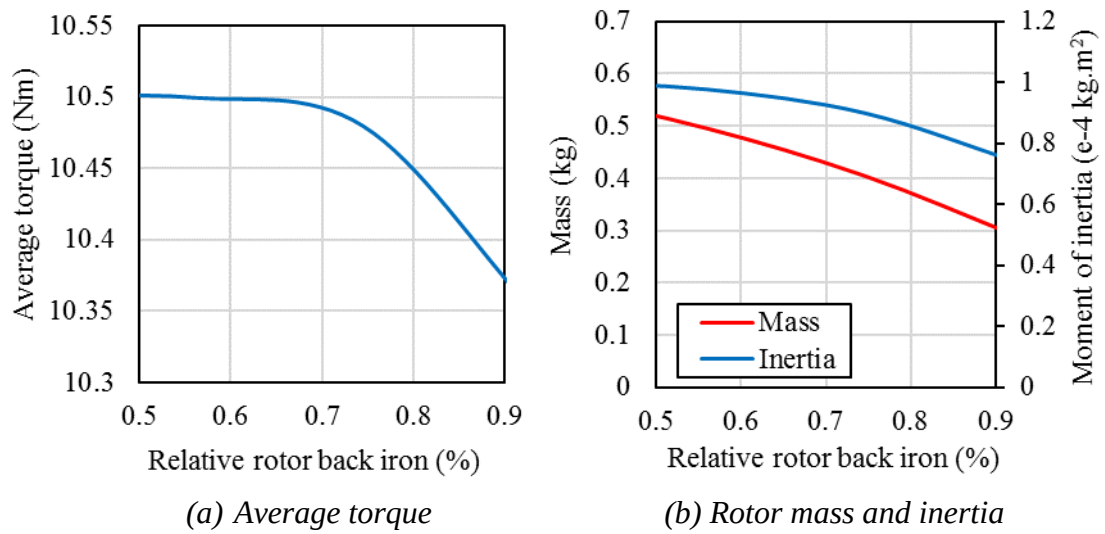


Figure 3-25: Influence of rotor back iron on

Therefore, the relative rotor back iron of 70% (4 mm) has been selected as an optimal value. With 4 mm, the total mass of the machine drops by 6.3 % for the same output torque. While, the rotor moment of inertia is within the specification.

3.10 Initial comparison to the machine targets

The machine model of the final post-optimisation design is shown in Figure 3-26 (a) and the flux density distribution at the rated torque condition of 10.5 Nm is reported in Figure 3-26 (b).

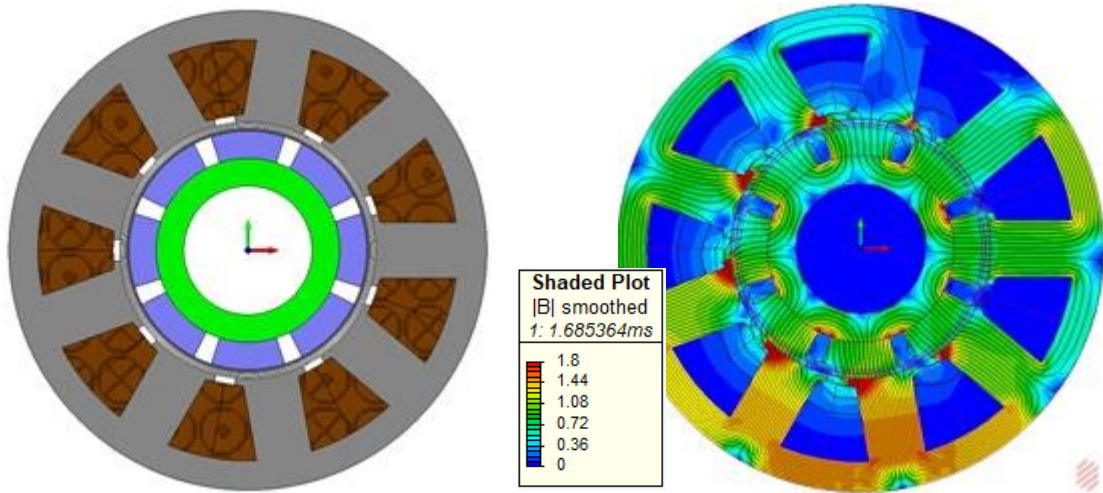


Figure 3-26: Cross-section of 8p/9s PMSM and the flux density map

Table 3-10 summarize the main design parameters, while the most important parameters of the designed machine for the two different operating conditions are listed in Table 3-11.

Table 3-10: Machine dimensions for preliminary design

Parameters	Values	Parameters	Values
Stack length	85 mm	Slot Depth	12.1 mm
Air-gap thickness	0.9 mm	Stator back iron thickness	4.275 mm
Sleeve thickness	0.4 mm	Slot opening	3 mm
Split ratio	53.5 %	Tooth Tip	1 mm
Stator inner diameter	37.45 mm	Magnet Thickness	4 mm
Rotor outer diameter	35.85 mm	Magnet Span	80 %
Rotor Back Iron	4 mm	Fill factor	45%
Tooth width	8.2 mm	Number of turn per phase	36

Table 3-11: Machine specifications for preliminary design

8p/9s PMSM	@ 8700 rpm	@ 19000 rpm
Rated torque	10.5 Nm	5.55 Nm
DC link voltage	270 V	
Current density, rms value	31.1 A/mm ²	
Peak phase current	85 Amp	
Peak phase Back-Emf (peak)	74.9 V	162.1 V
Torque ripple	0.76 %	0.85 %
Back-Emf THD	3.74 %	3.83 %
Total mass	2.68 kg	
Total moment of inertia	0.92·10 ⁻⁴ kg.m ²	
Efficiency	86.5 %	82.6 %
Total losses	1.48 kW	2.31 kW
Copper losses	1.18 kW	1.18 kW
Stator core losses	71.5 W	79.8 W
Rotor losses	166.51 W	570 W
Windage losses	69 W	486 W
Maximum winding temperature	202.78 °C	220.54 °C
Magnet temperature	110.4 °C	131.8 °C

As the goal of this section is to achieve the machine targets, Table 3-12 compare the design of 8p/9s PMSM against the targets. The tick (✓) shows that the target has been achieved and the cross (✗) represents a target that has not been met and needs modification. It can be safely said, that the machine meets the torque and current requirements. As can be observed, the main shortcomings of this design relative to the requirements are

- 1) Efficiency – 3.88% (at 8700 rpm) and 2.82% (at 19000 rpm) lower than the requirements, respectively.
- 2) Temperature, which was 5.01% higher than the set limits. In order to reduce inertia, more current was required to be pushed into the machine and therefore temperature rose.

- 3) Axial length – When considering the end windings, the total length of the machine was calculated to be 105 mm.

In the next chapter, methodologies to solve these last challenges are presented resulting in an improved design that is able to meet all the requirements.

Table 3-12: Comparison of PMSM design to machine targets

	Target	8p/9s PMSM	
Rated torque (Nm) @ 8700 rpm	10.5	10.5	✓
Torque (Nm) @ 19000 rpm	5	5.55	✓
Maximum current (A)	85	85	✓
DC link voltage (V)	210-270	210	✓
Efficiency (%) @ 8700 rpm	90	86.5	✗
Efficiency (%) @ 19000 rpm	85	82.6	✗
Winding temperature (°C)	210	220.54	✗
Maximum machine diameter (mm)	70	70	✓
Maximum machine length (mm)	100	105	✗
Mass (kg)	2.5	2.68	✗
Moment of inertia (kg.m ²)	$1 \cdot 10^{-4}$	$0.92 \cdot 10^{-4}$	✓

3.11 Highlights and contributions to research

The work related to this chapter has resulted in two conference papers [28, 48]. The main contributions to research from this chapter include

- 1) A defined study of the pros and cons of flooding of the rotors of high speed machined, that clearly inform future works about the adoption of such measures.
- 2) A ‘push the limit’ technique based on cooling channels in the slots of flooded machines. This was shown to be extremely beneficial for applications that ‘enjoy’ a cooling agent as part of the whole system.

3.12 Conclusion

In this chapter, the design of a high performance PMSM with high power density and a flooded airgap has been presented. The performances of different designs have been analysed and compared including different slot–pole combinations and winding configurations as shown in Section 3.5.1. From the initial trade-off studies, an 8-pole / 9-slot PMSM was identified as the optimal topology. To ensure a compact design of the 8p/9s PMSM, a high current density of $31.1 \text{ A}_{\text{rms}}/\text{mm}^2$ was used. To allow the use of such a high current density, an oil cooling system has been adopted, pushing the fluid directly through the windings in relatively innovative, customised slot ducts. While, the rotor is cooled by oil flow through gap between stator and rotor. A high windage loss was expected due to the machine rotating at relatively high speeds (19000 rpm). However, this technique was shown to be the most efficient way to remove the heat generated by the copper losses in the coil and the eddy current losses in the rotor as shown in section 3.7.

As the goal of this chapter was to meet the project requirements of Table 3-1. A sensitivity analysis was applied to investigate the effect of the main geometrical parameters on the machine performance as illustrated in Section 3.9. An excellent machine performance was achieved with high power density, low torque ripple and sinusoidal back-EMF. However, due to the high power to mass ratio and low inertia associated with this application, the machine struggled to meet the requirements of efficiency and thermal limits as shown in Table 3-12.

Therefore, as a final step towards the final design, several methodologies will be investigated in the following chapter in order to achieve the required performance level for the 8p/9s PMSM.

Chapter 4. FINAL DESIGN OF THE INDUSTRIAL MACHINE AND PERFORMANCE ANALYSIS

4.1 Introduction

In the previous chapters it has been shown that to improve the design further the losses and the mass have to be reduced and the thermal management must be improved. Therefore, the initial design is then put through more-detailed analysis in order to improve the performance so as to satisfy all the project requirements. The performance improvement techniques implemented here, include the use of high performance material such as Cobalt Iron (CoFe) laminations and the use of Halbach array configurations.

This chapter continues by introducing a complete losses analysis for the purpose of improving the efficiency and provide accurate estimation for the machine performance. Thermal and mechanical analyses are conducted to ensure the temperature and mechanical limits. Using all the information achieved from these analyses, the overall performance of machine was re-evaluated to ensure integrity of the design methodology. This chapter concludes by presenting the final design of the machine.

4.2 Improved machine design

To mitigate the design requirements that were not met in the previous section, different possible strategies were investigated to enhance the general performance of the machine and to satisfy all the requirements.

4.2.1 Changing the stator lamination material

In order to reduce the total copper losses (max. temperature) and improve the efficiency a material with better characteristics was investigated. The initial 8p/9s machine was optimized using SiFe material (M270-35A), with a lamination thickness of 0.35 mm and a saturation limit of 1.7 T. However, using Cobalt Iron (CoFe) material (VacoFlux 48) with a lamination thickness of 0.1 mm, the magnetic performance of the machine can be greatly improved as this material can achieve saturation points of about 2.3 T.

In Figure 4-1, the BH curve of both materials are reported and compared. By following the main parameters reported in Section 3.10, the 8p/9s machine is redesigned with these CoFe laminations. Figure 4-2 shows the field density map for the two optimized FE models. With the CoFe material the area of the slot is increased by 41.4% compared to SiFe material. The results obtained for the two different designs are presented in Table 4-1 for fixed envelope dimensions.

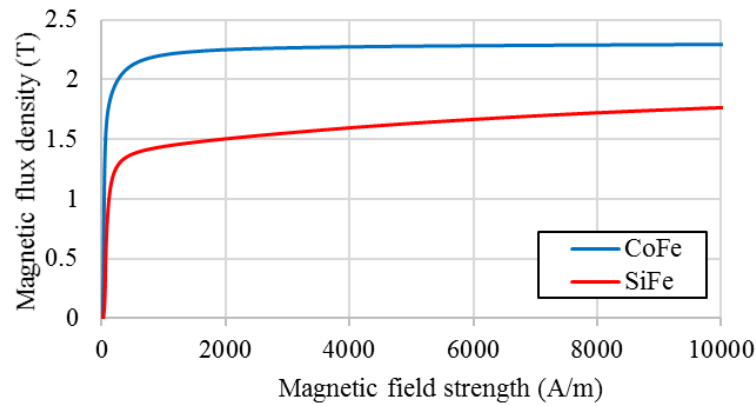
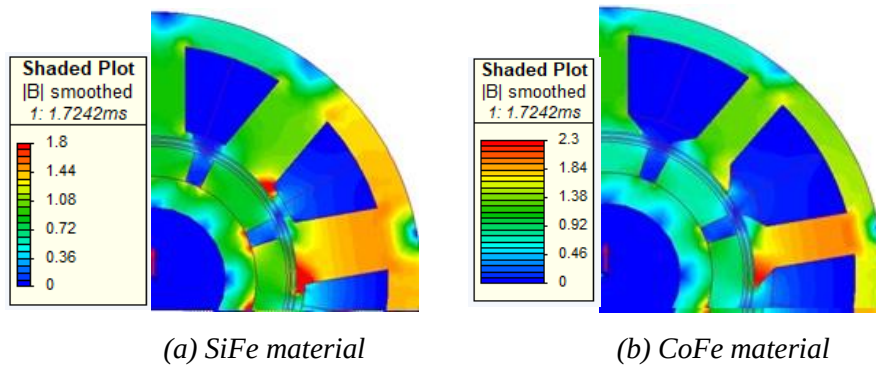


Figure 4-1: BH curve for the stator lamination material



(a) SiFe material

(b) CoFe material

Figure 4-2: Flux density map

Table 4-1: SiFe vs. CoFe performance comparisons @ 10.5 Nm

	SiFe	CoFe
Stator outer diameter (mm)	70	70
Split ratio (mm)	0.535	0.535
Stack length (mm)	85	85
Current density J (A/mm^2)	31.1	21.9
Total Copper losses (kW)	1.18	0.71
Efficiency (%)	86.5	91.2
Winding temperature ($^{\circ}C$)	202.7	150.7
Stator core mass (kg)	1.1	0.92

From Table 4-1, it is clear that, for the same fill factor the machine design using CoFe material led to lower current density and copper losses compared to the machine design based on SiFe material. Also having a CoFe material, the maximum temperature generated inside the slot was 25.6% lower compared to the design with SiFe material. Therefore, CoFe material is adopted for this application.

4.2.1.1 Stator fill factor

At this stage, a fill factor of 45% of the slot area has been used. While the cooling channel was assumed to be 10% of the slot area. However, in order to verify the feasibility of creating the proposed cooling channel for the new design, a dummy stator core was manufactured. To search for the best fill factor experimentally, the dummy stator core was rewound with different sizes of wires. From these tests, it is observed that the best results in terms of cooling channel are achieved by the use of 33.7% fill factor as shown Figure 4-3.

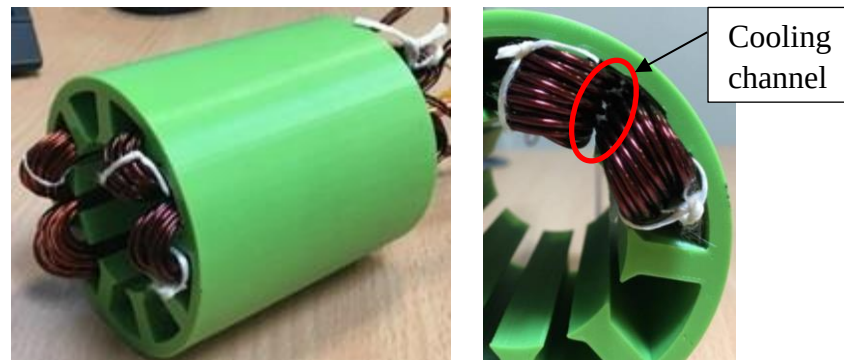


Figure 4-3: Cooling channel in a slot of dummy stator

With this new fill factor in place, the models were re-run, but now with them considering the effects of the cooling channel in the slots as shown in Figure 4-3. Table 4-2 tabulates the main design parameters, while the most important performance parameters of the designed machine for the two different operating conditions are listed in Table 4-3. From Table 4-3, it can be observed how the efficiency for the two different operating points is increased by 1.5% and 2.6%, respectively. However, with high copper losses, the maximum temperature generated inside the windings was still 2% higher than the requirements of 210 °C. While, the total length and mass of the machine were 5% and 7.2 % higher than the specifications.

Table 4-2: General Machine Dimension

Parameters	Values	Parameters	Values
Stack length	85 mm	Slot Depth	12.77 mm
Air-gap thickness	0.9 mm	Stator back iron thickness	3.5 mm
Sleeve thickness	0.4 mm	Slot opening	3.75 mm
Split ratio	53.5 %	Tooth Tip	0.5 mm
Stator inner diameter	37.45 mm	Tooth wedge (mm)	2 mm
Rotor outer diameter	35.85 mm	Magnet Thickness	4 mm
Rotor Back Iron	4 mm	Magnet Span	80 %
Tooth width	5.5 mm	Fill factor (FF)	33.7%

Table 4-3: General Machine Specifications

8p/9s PMSM	@ 8700 rpm	@ 19000 rpm
Rated torque	10.5 Nm	5.9 Nm
DC link voltage	270 V	
Current density, rms value	29.36 A/mm ²	
Peak phase current	85 Amp	
Peak phase Back-Emf (peak)	74.7 V	161.7 V
Torque ripple	0.43 %	0.55 %
Back-Emf THD	3.06 %	3.2 %
Total mass	2.68 kg	
Total moment of inertia	0.92·10 ⁻⁴ kg.m ²	
Efficiency	87.8 %	84.8 %
Total losses	1.32 kW	2.1 kW
Copper losses	1.06 kW	1.06 kW
Stator core losses	26.4 W	26.09 W
Rotor losses	171.3	526 W
Windage losses	69 W	486 W
Maximum winding temperature	209.3 °C	214.09 °C
Magnet temperature	104.9 °C	121.6 °C

4.2.2 Changing the magnetization structure of the PM

The performance of the electrical machine has been further improved with the adoption of Halbach arrays [101]. This arrangement led to a better air gap flux density with a more sinusoidal waveform which results in a more sinusoidal back-EMF, a lower torque ripple and lower iron losses. All these result in a higher torque [101]. Also, with Halbach arrays, other aspects of the machine, such as the rotor yoke can be significantly reduced [122]. The drawbacks related to this arrangement are the higher cost and complexity of manufacturing. Starting from the model shown in Figure 4-2, a quasi Halbach magnetized magnet is implemented instead of the conventional SMPM array as shown in Figure 4-4. It is important to note that the ratio between the transition pole (T) and magnet pole (M) which is defined as M:T in this work must be investigated in order to reduce the THD, torque ripple and satisfy the torque requirement.

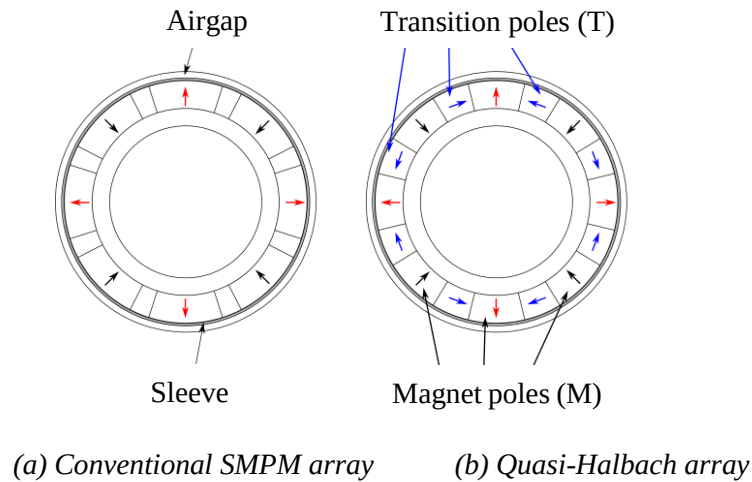


Figure 4-4: Cross-section of the PMSM with different rotor PM magnetization

It can be seen from Figure 4-5 that the utilization of Quasi-Halbach array has a significant impact on the output torque and THD, when compared with the general SMPM. The proper ratio of M:T is around 1.5. Table 4-4 tabulates the performance parameters of the two models for the same envelope size. It can be observed that, the total PM mass for the Quasi-Halbach model is increased by approximately 25.1%, which means further increase in the total moment of inertia. In fact, when considering the total machine mass, the Quasi-Halbach model results in a 2.05% increase. However, this is acceptable, when compared to the 12.7% increase in the output torque.

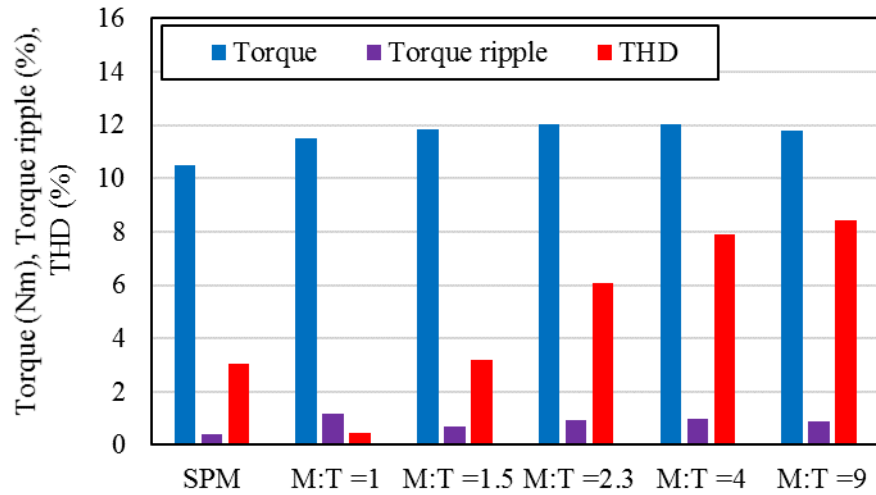


Figure 4-5: Performance parameters versus PM structures

Table 4-4: SMPM vs. Quasi-Halbach performance comparisons for a fixed envelop size

Performance Indicators	SMPM	Quasi-Halbach M:T=1.5
Torque (Nm) @ 8700 rpm	10.5	11.84
Peak phase current (A)	85	85
Torque density (Nm/kg)	3.97	4.46
Stator outer diameter (mm)	70	70
Stack length (mm)	85	85
PM mass (kg)	0.219	0.274
Total mass (kg)	2.68	2.735
Moment of inertia ($kg \cdot m^2$)	$0.92 \cdot 10^{-4}$	$0.98 \cdot 10^{-4}$

In general, the objective here is to satisfy the torque requirements of 10.5 Nm while respecting all the other constraints. Therefore, Table 4-5 presents the performance parameters of the two models for the same output torque. Compared with a conventional SMPM (parallel magnetization), a quasi-Halbach array configuration presents a reduction of the total machine mass of 6.7%, with a stack length decreased by 5.88%. Considering the stack length reduction, the windage losses at the maximum speed condition dropped by 5.9% to 457 W. Besides, the copper losses reduced by 18.86% due to the reduction of both stack length and phase current and consequently lower windings temperature.

From the above, the superiority of the Quasi-Halbach arrangement compared to the conventional arrangement was observed on the general performance of the electrical machine. Thus, considering all the above it was decided that the implementation of the Quasi-Halbach configuration is necessary for the high performance requirements of the application.

Table 4-5: SMPM vs. Quasi-Halbach performance comparisons for a fixed output torque

Performance Indicators	SMPM	Quasi-Halbach M:T=1.5
Torque (Nm) @ 8700 rpm	10.5	10.5
Stator outer diameter (mm)	70	70
Stack length (mm)	85	80
Peak phase current (A)	85	80
Current density J (A/mm ²)	29.36	27.2
Total Copper losses (kW)	1.06	0.83
Windage losses (W)	486	457
Winding temperature (°C)	209.3	182.07
PM mass (kg)	0.219	0.258
Total mass (kg)	2.68	2.5
Moment of inertia (kg. m ²)	$0.92 \cdot 10^{-4}$	$0.96 \cdot 10^{-4}$

4.3 Demagnetization analysis

One of the main challenges with PM machine is the requirement to ensure that no partial or full demagnetization occurs for all operating conditions. If the magnet thickness and coercivity are large enough, the risk of demagnetization is not present [122]. However, during the flux weakening operation when higher speed of the PM machine is required, the flux of the permanent magnet is reduced by applying negative d-axis current in order to drive the machine to higher speed and the possibility of demagnetization of the magnet will appear. Since, the cooling system for the rotor of this machine is selected to be a flooded. Therefore, it is important to check the demagnetization of the permanent magnet during the flux weakening operation with maximum working temperature (worst case scenario).

The maximum operating temperature of magnet is limited to 350°C. At this temperature, the knee point of the demagnetization characteristic of $\text{Sm}_2\text{Co}_{17}$ magnet appears at about 0.27 T. If the flux density is smaller than the knee point in any area of a magnet then, the magnet is partially or completely demagnetized.

A demagnetization prediction analysis is performed using finite element simulation. The risk of irreversible demagnetization occurs when the PM operating point goes down below the knee of the curves (shown in red in Figure 3-5). Even, in the worst case, with PM temperature set at 350°C and maximum speed operation of 19000 rpm, only a small amount of irreversible demagnetization occurs. This is shown in Figure 4-6, where the red regions represent the irreversible demagnetized regions.

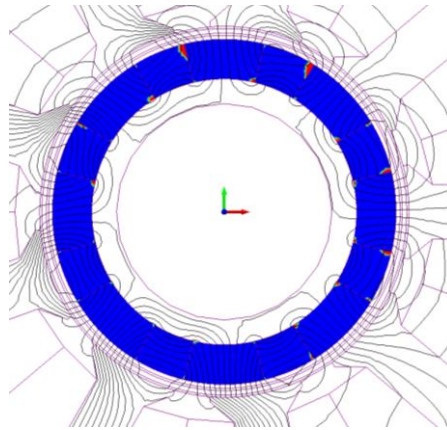


Figure 4-6: Demagnetization prediction at $T_{pm} = 350\text{ }^{\circ}\text{C}$ and 5 Nm @ 19000 rpm

However, for this specific application the maximum operating temperature generated inside the permanent magnet is about 130 °C. For this temperature, the flux density is greater than the knee point as shown in Figure 3-5. Therefore, the magnets are safe from demagnetization.

4.4 Updated comparison to the machine targets

From the aforementioned investigational studies, the final design of the 8p/9s PMSM comprises the following features:

- Hollow solid rotor shaft
- Completely flooded cooling system
- CoFe stator laminations
- Quasi-Halbach arrangement for the PM

The machine model of the final design is shown in Figure 4-7 (a) and the flux density distribution is reported in Figure 4-7 (b). The main results of this final model can be observed in Table 4-5 (for the Quasi-Halbach), so there is no need to repeat them in this section. However, for the sake of clarity, Table 4-6 compares the final machine design against the targets. It can be observed that the main requirements of an output torque of 10.5 Nm is achieved while respecting the key challenges of the targets such as thermal limit, envelop dimensions, and rotor moment of inertia.

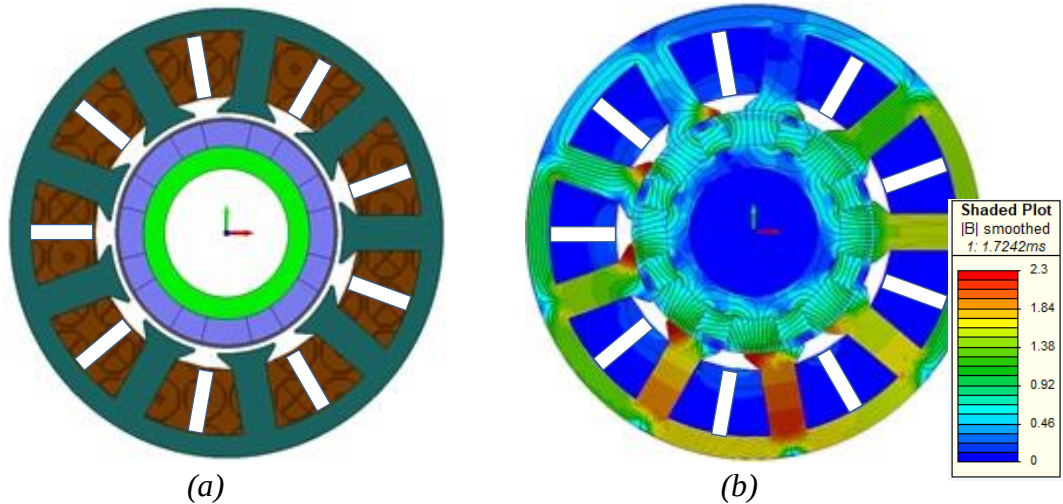


Figure 4-7: Cross-section of 8p/9s PMSM and the flux density map at rated torque

Table 4-6: Comparison of PMSM design to machine targets

	Target	8p/9s PMSM	
Rated torque (Nm) @ 8700 rpm	10.5	10.5	✓
Torque (Nm) @ 19000 rpm	5	6.27	✓
Maximum current (A)	85	80	✓
DC link voltage (V)	270	210	✓
Efficiency (%) @ 8700 rpm	90	90.14	✓
Efficiency (%) @ 19000 rpm	85	88.27	✓
Winding temperature (°C)	210	189.16	✓
Maximum machine diameter (mm)	70	70	✓
Maximum machine length includes end-winding (mm)	100	100	✓
Mass (kg)	2.5	2.5	✓
Moment of inertia (kg.m ²)	$1 \cdot 10^{-4}$	$0.96 \cdot 10^{-4}$	✓

4.5 Losses Modelling

Considering the main aim of this PhD thesis, then an important point of interest is to achieve an understanding of the loss distribution in the designed machine. This characterisation exercise is critical in order to set the context for the developments presented in Chapter 6. This Section is therefore focused on characterising and mapping all the losses of the machine designed above and this is done using 2D and 3D FE models.

4.5.1 Copper losses

In order to improve the thermal estimation, it is important to include the AC losses effect in the copper losses prediction, to ensure that they do not change the temperatures significantly. The high fundamental frequency of the machine (1266.67 Hz) required the use of stranded wire to reduce skin and proximity losses. Figure 4-8 shows the cross-section of stator core with slot winding for the 8p/9s PM machine with different number of parallel strands but for the same number of turns per coil (12 turns) and same phase current to investigate the effect of using parallel strands on reducing the ac losses. In addition, a refined mesh for the copper wire strands has been set for a precise distribution of the currents at high frequency. The proximity losses have been investigated including the effect of both armature reaction and rotor flux.

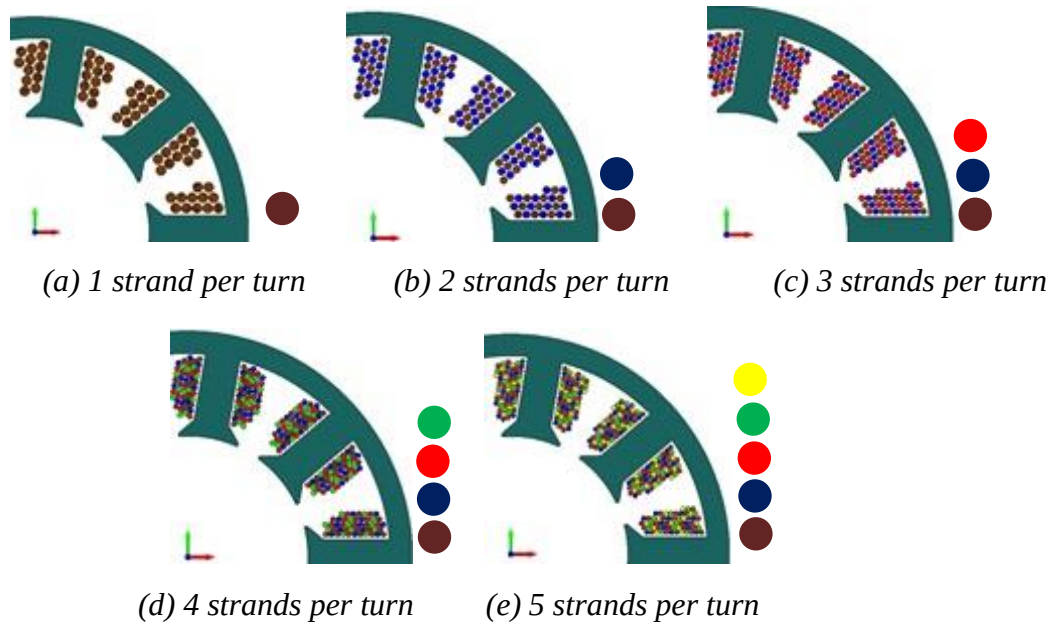


Figure 4-8: Cross-section of stator core with different number of strands per turn

Figure 4-9 shows the ratio of (AC+DC)/DC copper losses versus the operating frequency of the machine. It can be seen that using only 1 strand per turn the total copper losses at the maximum speed of operation was about 26% higher than the dc losses. Without considering proximity and skin losses, the copper losses will be largely underestimated. On the other hand, the copper losses decrease with the increase of strand number and the change gradually slows down. The changes in copper losses are very small and can safely be ignored when the number of strands is 5. Therefore, 4 strands per turn have been selected. In addition, increasing the number of strands even more is not attractive option due to the fact that the circulating current will increase the ac losses even more, the difficulty for winding the machine and the small difference on reducing the total ac losses. Figure 4-10 depicts the 3D-FEA, where the phase winding consists in 3 series connected coils, which are made of 12 turns each. The turns are arranged on 4 strands per turn and a 0.873 mm diameter wire (including the insulation thickness) is adopted for the strand.

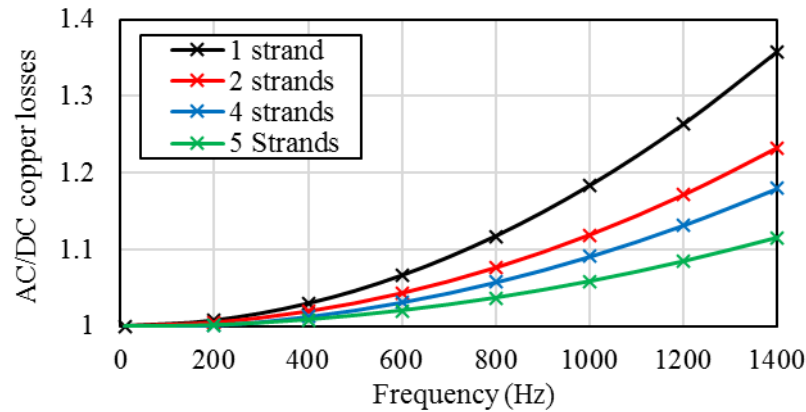


Figure 4-9: Ratio of ac to dc copper losses with different number of strands

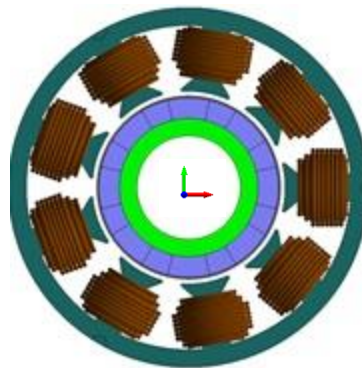


Figure 4-10: Cross-section of 8p/9s PMSM with 12 turns per coil and 4 strands

The trend of the AC winding losses ratio ($R_{ac}/R_{dc}|_T$) at 80 A and as a function of the

frequency is presented in Figure 4-11. For the end-winding, the R_{ac}/R_{dc} ratio is very small compared to the one of the active-length winding. This is due to fact that the leakage flux along the active-length winding is higher than in the end-winding region, where the windings are surrounded by air. Therefore, lower proximity losses occur in the end-winding area.

Figure 4-12 shows the distribution of the magnetic flux density in the stator conductors. The stator slot leakage flux is higher in the slot opening region compared to the slot top region and the largest eddy currents are induced close to the slot opening. Therefore, it's preferred to place the conductors towards the top of the slot and away from the slot opening region in order to reduce the ac losses due to the leakage flux which results in non-uniform current density distribution inside the conductors in one slot.

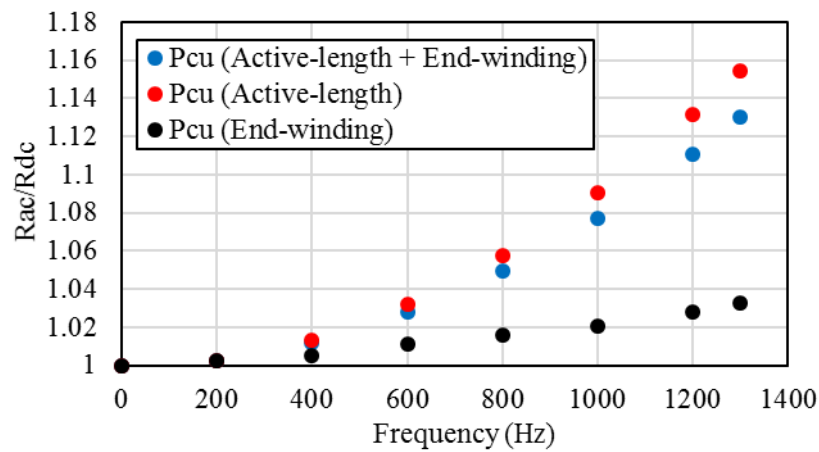


Figure 4-11: Ratio of ac to dc copper losses

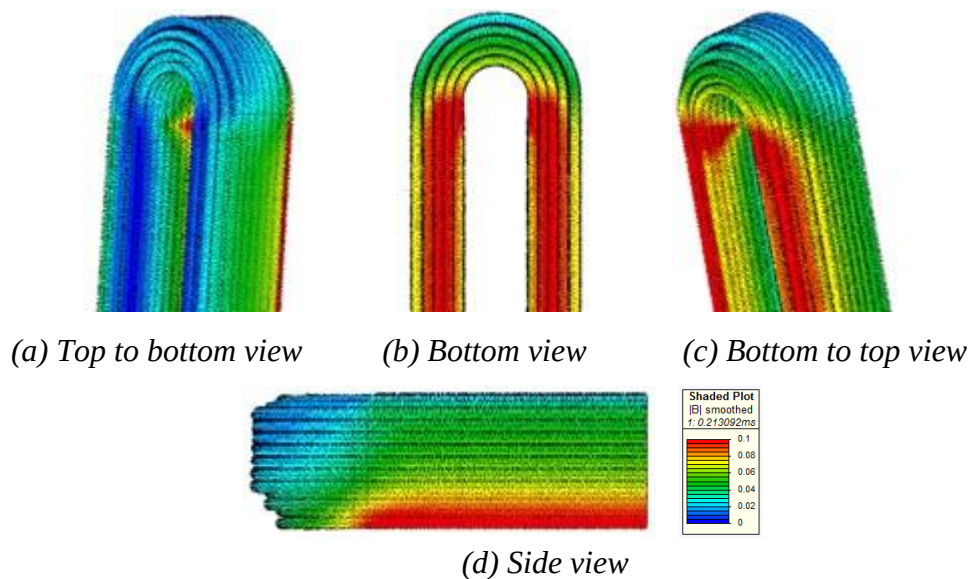


Figure 4-12: Flux density distribution in the stator winding using 3D FE model

4.5.2 Stator core losses

Figure 4-13 (a) shows the stator core losses of the 8p/9s PMSM for different operating speeds. It is clear that, the core losses are increasing with speed at no load. For this condition, the magnetic flux in both teeth and yoke is related to the permanent magnet and is almost constant while the frequency is increasing which lead to increasing the iron losses. However, at full load condition and from (0 to 9000) rpm a constant torque operation is applied and the iron losses increasing related to both frequency and flux density (permanent magnet and armature current). While, from (9000 to 19000) rpm a flux weakening operation is applied to drive the machine to higher speed operation. Even, the speed is increasing the flux density in the stator core is reduced and distorted due to the flux weakening. Therefore, the total iron loss is reducing during the flux weakening operation to certain frequencies and then start to increase. Due to the end effect, the core losses resulting from the 3D-FEA are higher than the ones provided by the 2D-FEA. Indeed, the end-winding leakage flux causes a local flux density increment in the edge of tooth-shoe region, as highlighted in Figure 4-14, which leads to extra core losses.

Figure 4-13 (b) shows the core losses of the SPM machine at rated speed of 8700 rpm for several q-axis currents. At low current levels, i.e. below 60A, the core losses predicted by 3D-FEA are lower than that of the 2D-FEA due to the leakage effect, which results in lower magnetic flux density in the machine core. Conversely, at higher current levels, i.e. above 60 A, the 3D-FEA predicted core losses exceed that of the 2D-FEA. This is due to the flux produced by the current flowing inside the end-windings.

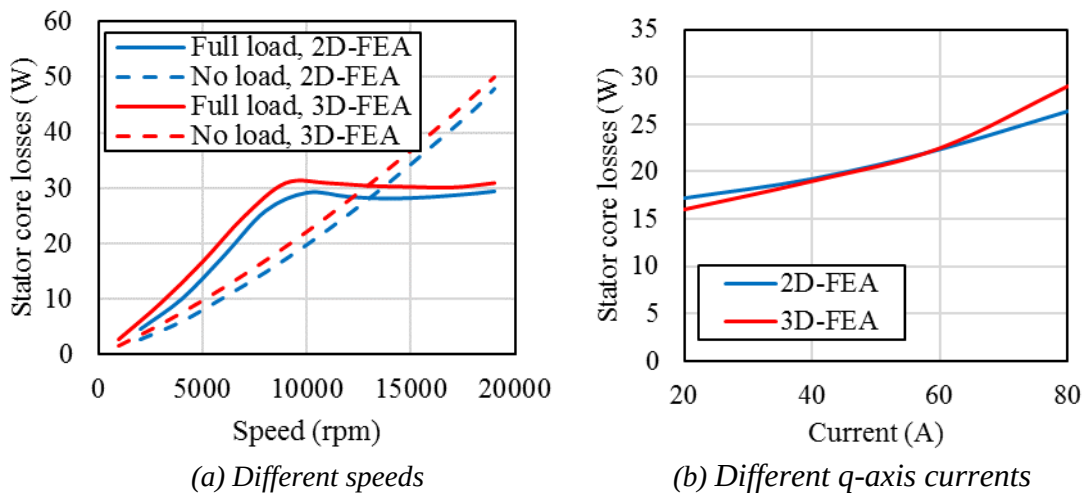


Figure 4-13: Stator core losses of 8p/9s PMSM

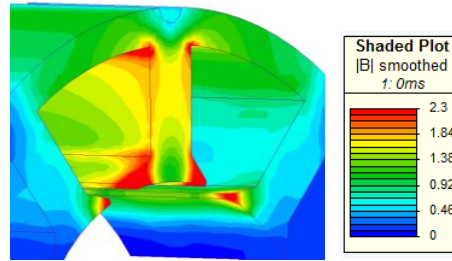


Figure 4-14: Flux density distribution in the stator core

4.5.3 Rotor eddy-current losses

Eddy current losses might also be generated in the PMs, sleeve and the stainless-steel shaft, due to high conductivity of these materials. The main causes of rotor eddy current loss are the slotting effect and winding space harmonic. In case of fractional slot PMSM with concentrated winding, the losses are more critical due to the high harmonic content of the armature field. Figure 4-15 shows the distribution of the current density in the rotor components at maximum speed of 19000 rpm.

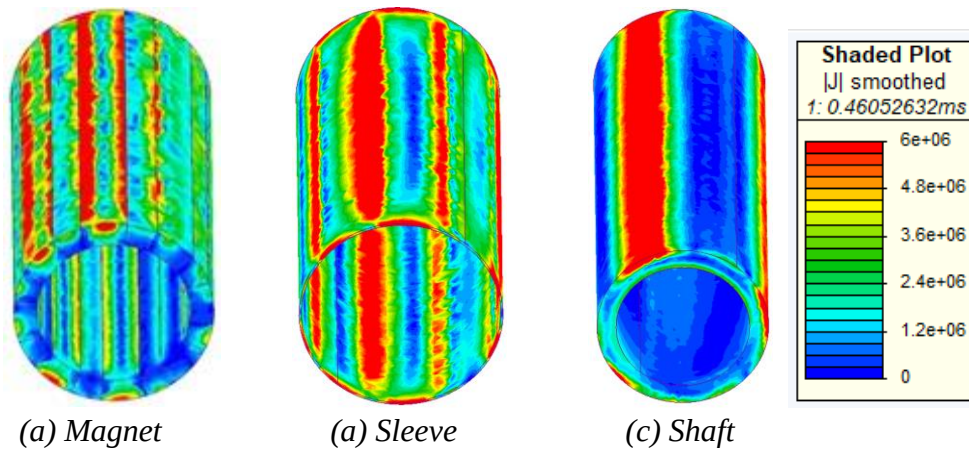


Figure 4-15: Current density distribution in solid rotor shaft at 19000 rpm

The eddy current losses in magnet, sleeve and solid shaft are calculated accurately by using FEA and the results are shown in Figure 4-16. Since the machine is designed with a semi-open slot the influence of the no load eddy current losses caused by the stator slotting is not considerable for all the conductive components. However, at full load condition the magnet losses vary significantly with the speed. This clearly indicates the impact of the armature reaction field harmonics. Considering the same load condition, lower eddy current losses are observed in 3D-FEA compared to the 2D-FEA. This is the result of lower leakage flux in 2D-FEA, which does not model end-paths and

restrains the flux lines on a plan. Since a higher amount of ‘active’ flux is considered by the 2D-FEA, the resulting eddy current losses are higher.

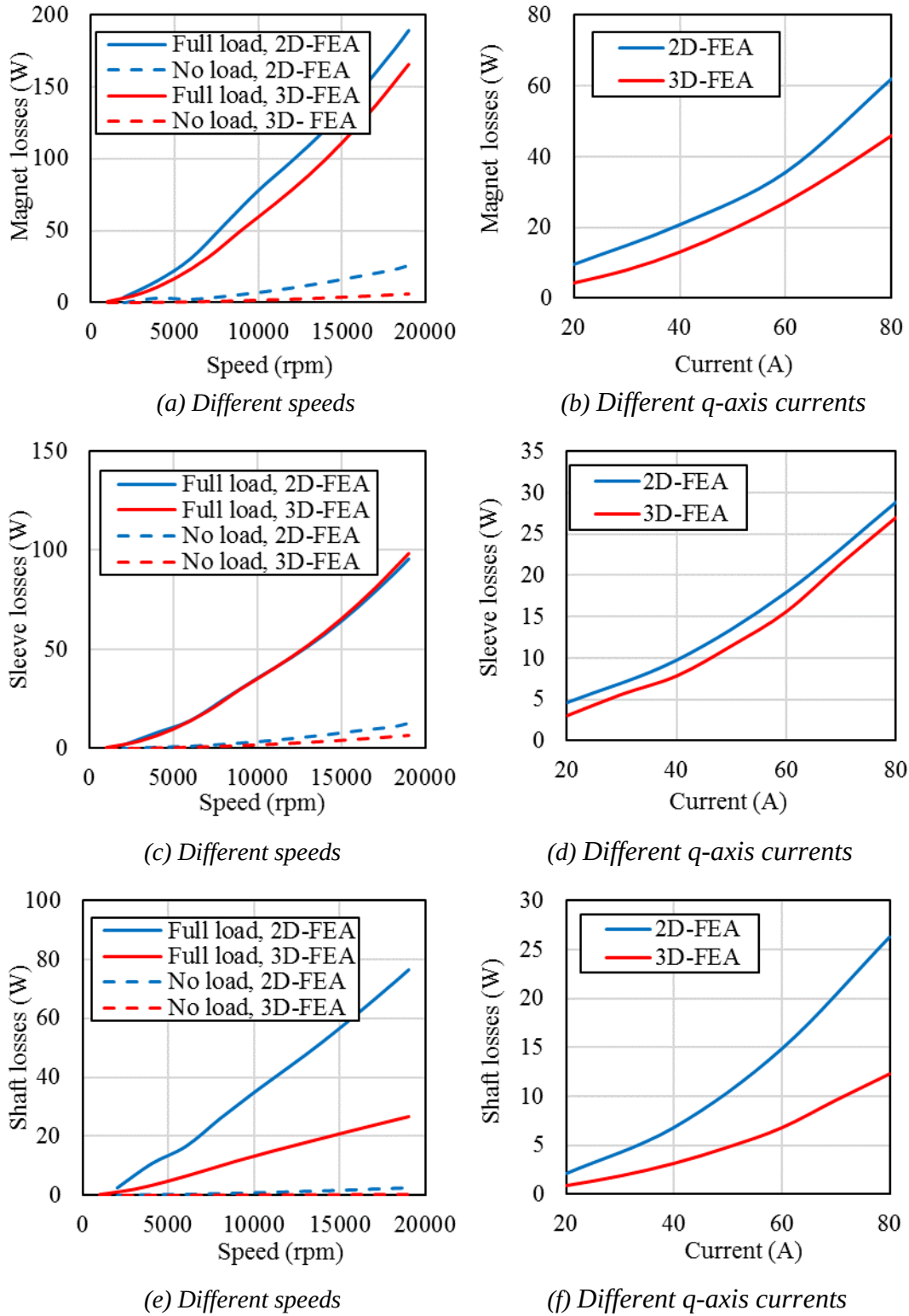


Figure 4-16: Magnet, sleeve and rotor shaft losses of 8p/9s PMSM

However, in order to improve the PMSM efficiency, the eddy current losses in the permanent magnets are reduced by adopting the segmentation of the PMs. The performance of the PMSM using magnets axially segmented is investigated by means of 3D FE simulations. Figure 4-17 shows the effect of the axial segmentation of the magnet on the eddy current losses at 580 (8700 rpm, base speed) and 1266 Hz (19000 rpm, maximum speed) for the 8p/9s PMSM. The magnet segmentation in the axial direction with high number of divisions can be expensive for example, (double the number of segments will increase the cost per rotor by 20%) and of course more complex to assembly. Also, with the increased number of magnets segments a gradual decrease of the magnet eddy current loss can be achieved as shown in Figure 4-17 (a).

Therefore, to provide a compromise between the manufacturing cost and the magnet losses, 10 segments of permanent magnet per pole has been selected where the magnet eddy current losses are drop by 43% at the maximum speed.

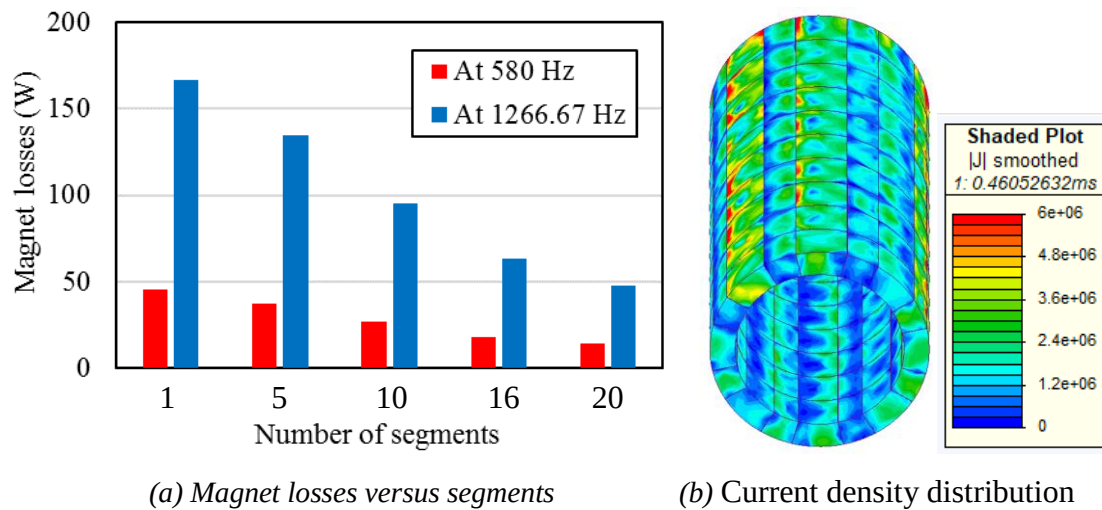


Figure 4-17: Effect of segmentation of eddy current losses

Figure 4-17 (b) shows the current density distribution in a permanent magnet with 10 segments at 19000 rpm. By segmenting the PMs, the eddy currents paths are reduced, since the eddy currents are constrained inside each PM single segment. Hence, a higher number of segments results in shorter paths and lower eddy currents losses compared to non-segmented magnet shown in Figure 4-15 (a).

4.5.4 Mechanical losses

4.5.4.1 Windage losses

The machine housing is designed to be oil flooded. The maximum oil flow rate is 4.2 L/min and the maximum oil inlet temperature is 90 °C. As the rotating velocity of the machine is high, the windage loss at the rotor surface becomes critical in terms of machine efficiency. While, at low temperature the viscosity of the fluid increases significantly.

Therefore, it is quite necessary to evaluate the windage losses considering different rotational speeds and temperatures to have a better understanding for the windage losses mechanism.

A. The Effect of Fluid Temperature

The cooling oil properties are strongly dependent on the temperature. In particular, its kinematic viscosity significantly varies with the temperature (nonlinear dependency), according to the trend shown in Figure 4-18. In particular, the kinematic viscosity decreases (i.e. lower windage losses and drag torque) as the oil temperature increases. Therefore, the influence of the temperature on the kinematic viscosity of the cooling oil is taken into account, when both windage losses and drag torque are determined.

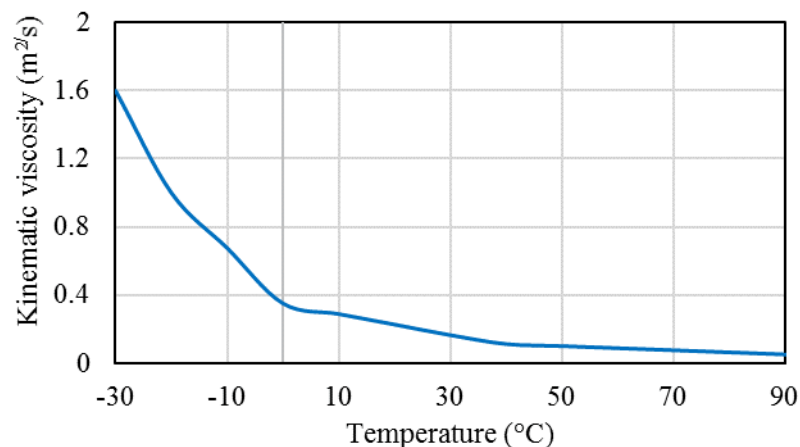


Figure 4-18: Kinematic viscosity vs. temperature for the adopted cooling oil

Using equation (2-7), windage losses and drag torque were analytically determined in the range from -40°C to +100°C, for several speeds up to 20 krpm. The obtained results

in terms of windage losses are shown in Figure 4-19 (a), while the drag torque results are reported in Figure 4-19 (b). At the lowest operating temperature of -40°C the motor will start to run, and a torque will require to overcome the oil drag of the flooded airgap. Thus, the output torque required to accelerate the rotor inertia will reduce. Also, it can be seen that at the maximum speed of 19000 rpm and for the maximum temperature the output torque of the electrical machine should be about 0.3 Nm higher than the torque required to drive the pump or 3.5 Nm at the lowest temperature. As a result, the output torque of the electrical machine should be increased by this value in order to provide the required power to the pump. Therefore, it is necessary to calculate accurately the drag in the output torque over the entire operating conditions for designed machine.

Despite the wide temperature range considered in the study, the oil temperature adopted for the machine design is equal to 90°C (as required by the project specifications) Indeed, the cooling oil used in the actual application does not exclusively cool the designed machine, but being already available, it also performs other cooling functions.

In particular, the windage losses represent 4% and 28% of the total losses, respectively at 8700 rpm and 19 krpm, when the temperature is equal to 90°C . Additionally, the drag torque varies from zero to 0.3 Nm, within the all operating speed rage (i.e. up to 19 krpm), at the same temperature (i.e. 90°C).

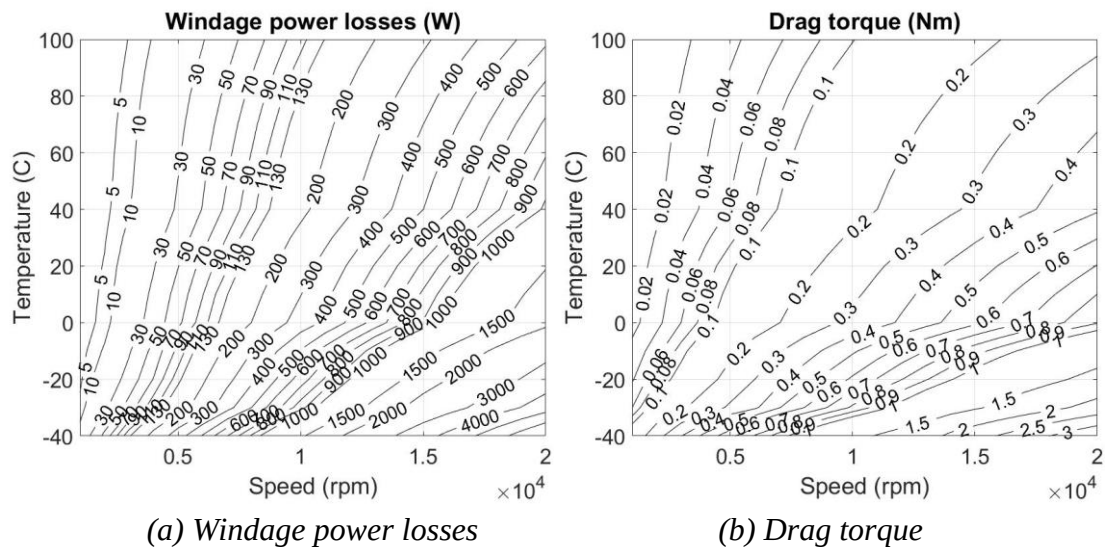


Figure 4-19: The effect of temperature and speed on

4.5.4.2 Bearing losses

The bearings used for the PMSM are two units of a single row, deep groove, ball bearings (61903). The following graph shown the bearing losses at different operating speeds using SKF model [93].

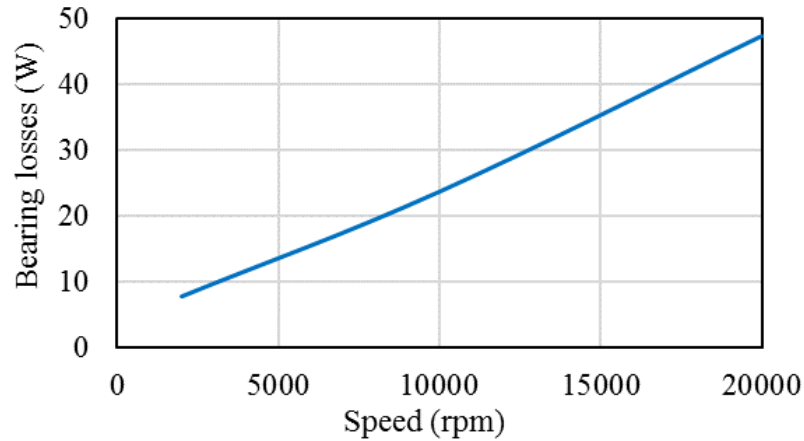


Figure 4-20: Bearing losses at different speeds using SKF model

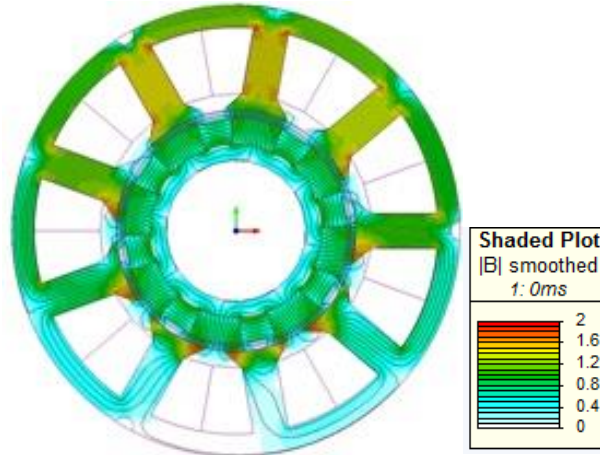
4.6 Performance evaluation

In this section, the performance of the designed PMSM is highlighted, under both no-load and full-load conditions. In order to evaluate the effect of the PM segmentation and the Joule losses in the end-windings, 3D FE simulations are performed.

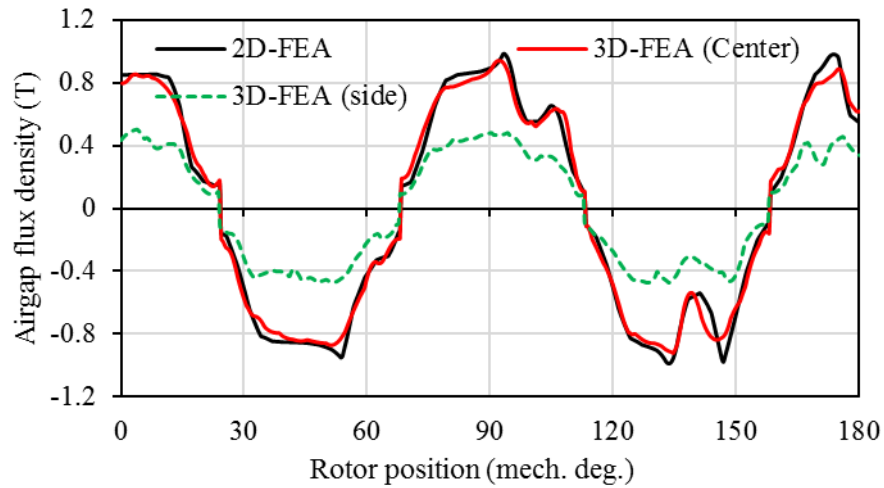
A. No load condition

The 2D no load flux density map of the machine is presented in Figure 4-21 (a), while the airgap flux density distribution is shown in Figure 4-21 (b). In the 3D model, the airgap flux density has been detected in two regions: a) at the middle of axial length and b) at one rotor end. Lower amplitude and higher distortion of the airgap flux density is observed at the end of the machine, due to the influence of the end leakage.

Considering the airgap flux density at the middle of the axial length, similar distribution is obtained for both 2D and 3D-FEAs, although 3D-FEA distribution reveals lower peaks, this is due to axial leakage in the 3D-FEA.



(a) Flux density map



(b) Flux density distribution

Figure 4-21: No load flux density

The machine back electromotive force (Back-EMF) waveform and its harmonics at 8700 rpm are shown in Figure 4-22 (a) and Figure 4-22 (b). The Fourier series transformation has been applied to the estimated Back-EMF waveform. The total harmonic distortion (THD) at the base speed of 8700 rpm are 3.13% and 4.8% for the 2D and 3D FE results, respectively.

The Back-EMF waveform at the maximum operating speed of 19000 rpm is obtained and the results reported in Figure 4-22 (c), while the harmonic content is shown in Figure 4-22 (d). The THD is 3.53% using 2D-FEA and 4.76% using 3D-FEA. Lower back-EMF amplitudes are observed when the end-windings are included in the 3D model. This is due to the end magnet leakage which reduces the amplitude fundamental harmonic of the back-EMF. While the higher THD is mainly given by the increase of the 3rd harmonic.

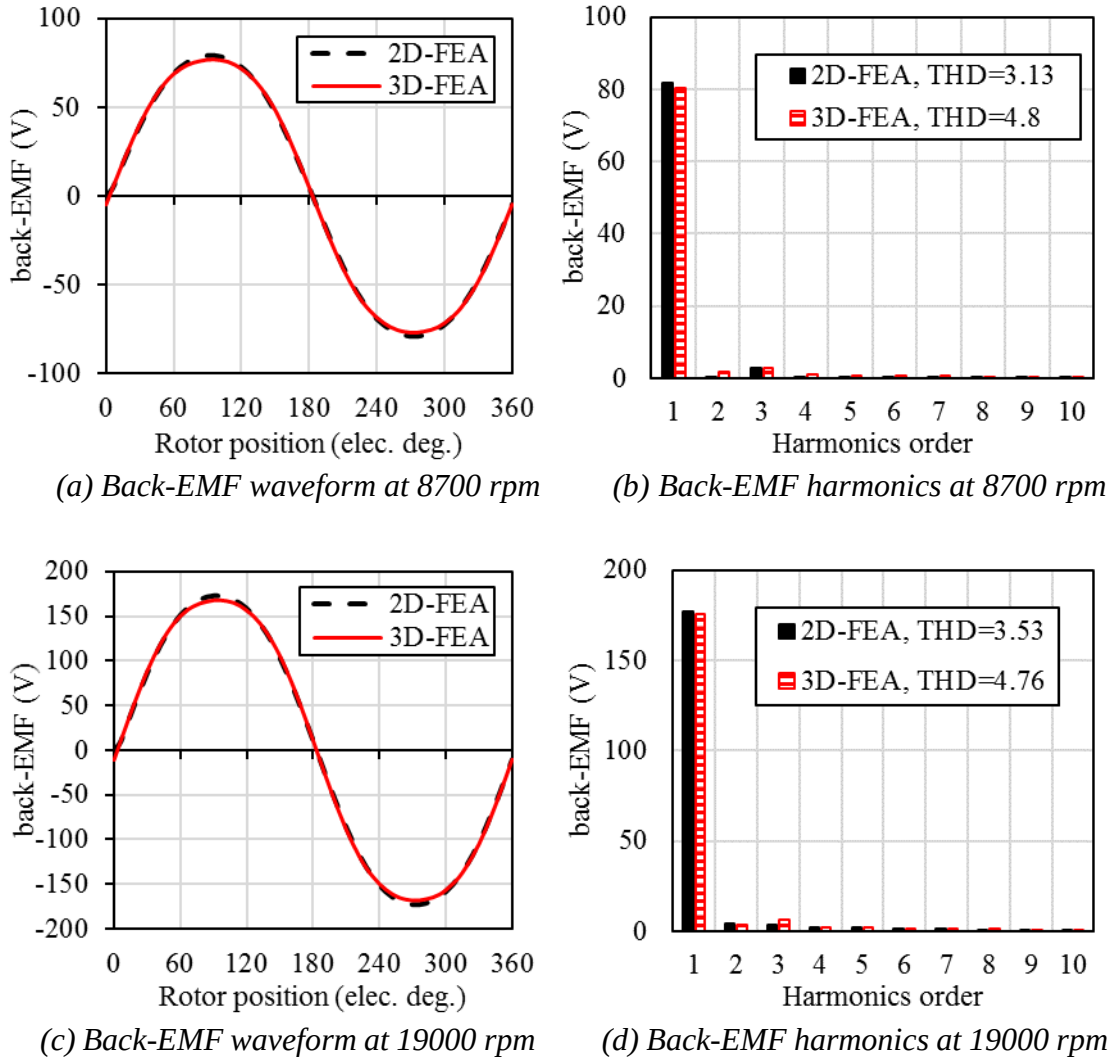


Figure 4-22: Back-EMF of the 8p/9s PMSM

B. Full load condition

The machine full load performance has been investigated to evaluate the electromagnetic torque and the torque ripple. 2D and 3D (transient with motion) simulations are considered for this analysis. The machine torque waveform at full load

of ($I_q=80\text{A}$, $I_d=0$) with rated speed of 8700 rpm is presented in Figure 4-23, as calculated by FEA analysis. The electromagnetic torque of the 2D and 3D-FEA has an average of 10.52 Nm and 10.3 Nm, respectively. Also, the torque waveform has significantly low torque ripples (0.9% for the 3D-FEA and 0.5% for the 2D-FEA) due to low cogging torque and Back-EMF total harmonic distortion. Therefore, the torque ripple difference between the 2D-FEA and 3D-FEA can be neglected. Also, the torque constant K_T has been calculated where only the q-axis current is applied to the machine and the results obtained from the 2D model was 0.1315 Nm/A while 0.1283 Nm/A for 3D model. The flux density map for the two operating points are presented in Figure 4-24. All the above indicates that the machine design satisfies the requirements of the application at hand. From the numerical analyses presented above, it is perceived that the 3D-FE model considers more aspects than its 2D counterpart and therefore gives a more accurate representation of the electrical machine.

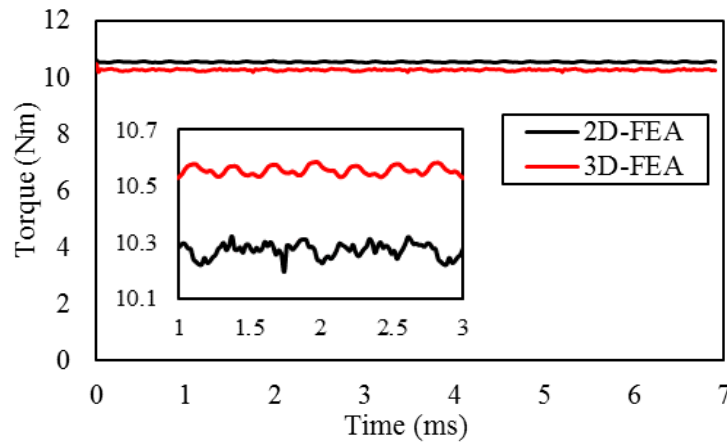
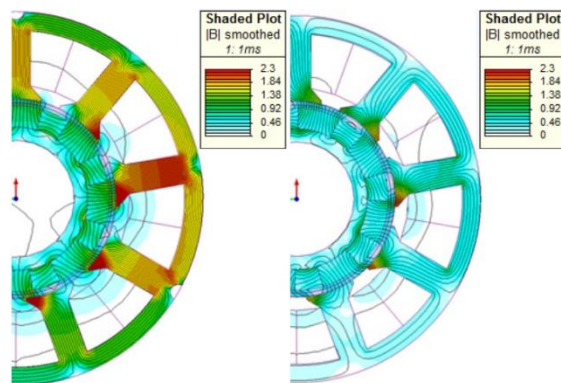


Figure 4-23: Torque at full load condition @ 8700 rpm and 80 A



(a) 10.5 Nm @ 8700 rpm

(b) 5 Nm @ 19000 rpm

Figure 4-24: Full load flux density map and flux line distribution

C. Torque-speed characteristics

In order to determine the torque-speed characteristic of PMSM, the method proposed in [123] is used to account for both magnetic saturation and cross-coupling effects. Figure 4-25 and Figure 4-26 show the PM flux linkage at different q-axis currents and the d- and q-axis inductances against d- and q-axis currents using 2D and 3D FEA. Due to the PM end-leakage flux, i.e. the end-effect, the PM flux-linkage predicted by 2D-FEA is higher than the 3D-FEA. In addition, higher armature reaction on the PM flux-linkage is observed in the 3D-FEA results compared to its counterpart of 2D-FEA, as the high current level leads to higher magnetic saturation in the stator and therefore higher stator end leakage. Overall, only a small reduction is observed, i.e. approximately 2.04% and 3% with an open circuit and a current of 80 A, respectively. In general, the inclusion of the end-winding in the calculation of the inductances results in higher values than the expected outcomes of 2D-FEA. This is mainly due to the end leakage inductance. It can also be seen that magnetic saturation at higher currents is larger and therefore, due to the higher leakage flux at the stator core of the machine, low inductance values are observed.

Figure 4-27 shows the torque and power curves calculated by the 2D and 3D FEA. When the end-effect is not taken into account, the 2D-FEA results provide higher torque and a higher base speed values in the constant torque region.

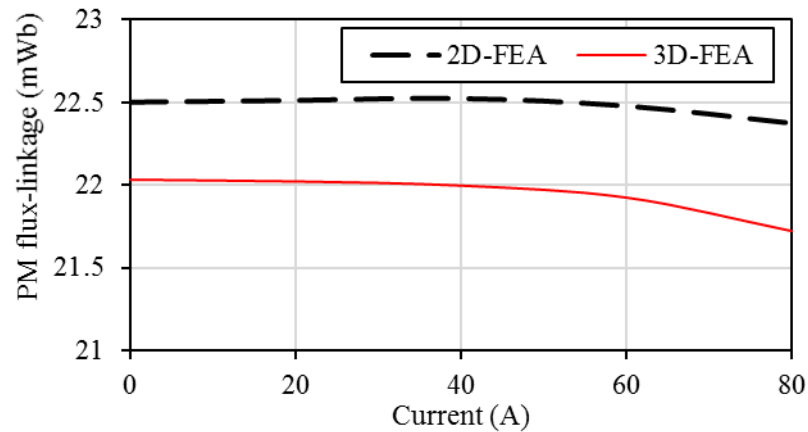


Figure 4-25: PM flux linkage at different q-axis currents

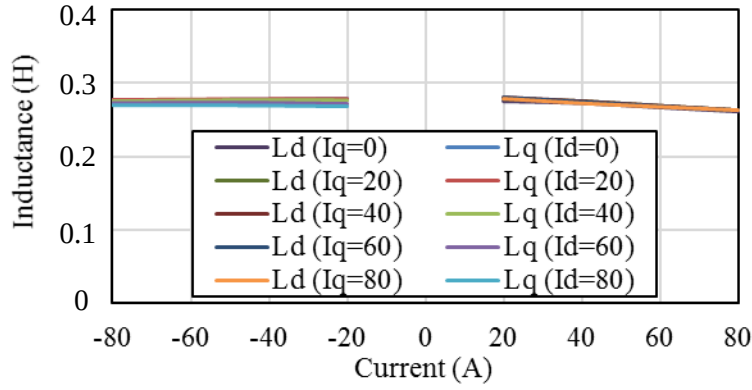
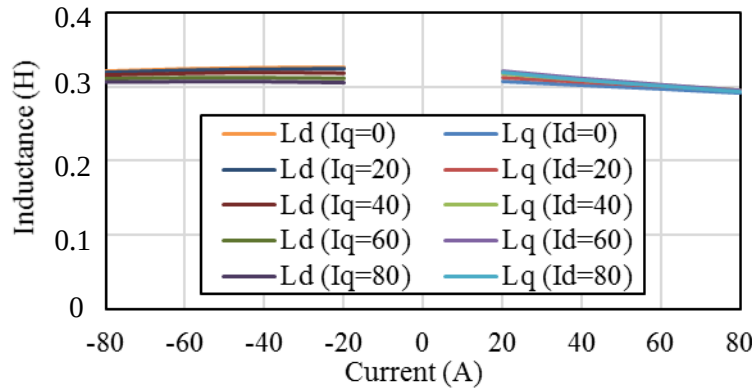
(a) *D and q-axis inductances using 2D-FEA*(b) *D and q-axis inductances using 3D-FEA*

Figure 4-26: Inductances calculations at different currents

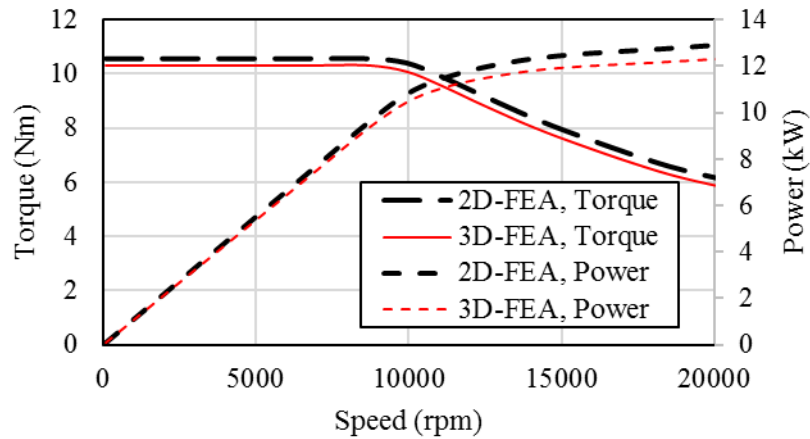


Figure 4-27: Torque-speed, Power-speed characteristics

This outcome was expected since the PM flux-linkage predicted by 3D-FEA is lower. In addition, the 3D-FEA results show a further reduction in the flux weakening capability (power and torque) due to increased d-axis inductance and reduced in PM flux linkage.

D. Efficiency

The losses terms previously predicted by FEA have been employed to determine the machine efficiency at several operating speeds, using (4-1), where P_{cu} is the total copper losses ($P_{DC}+P_{AC}$), P_{core} is the stator core losses, P_{eddy} is the rotor eddy current losses and P_{mech} is the mechanical power.

$$\eta = \frac{P_{mech}}{P_{mech} + P_{cu} + P_{core} + P_{eddy}} \quad 4-1$$

The PMSM efficiency results are presented in Figure 4-28. Low efficiency was expected at the low speed region due to the large amount of copper losses. However, the efficiency is then increased with speed increases up to 11000 rpm and then decreased with speed increase as the windage and rotor eddy current losses are higher in the high-speed region. Despite the reduction in the output power obtained by means of 3D-FEA simulation, a higher efficiency was observed in the high-speed regions. The reason of this, was due to the reduction of losses (i.e. rotor eddy current losses) as shown in Section 4.5.3. For the final design and by using 3D-FEA, the efficiency was 90.06% and 87.93% at the speeds of 8700 rpm and 19000 rpm, respectively.

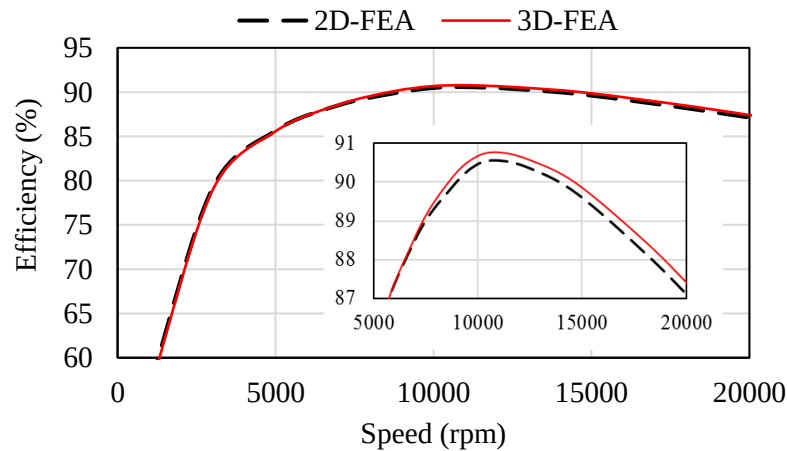


Figure 4-28: Efficiency characteristics

E. Thermal analysis and validation

For the worst operating full load case of 19000rpm, the AC losses were estimated to increase the DC losses by 12%, giving a machine total copper loss of 929.6 W. This could then be used to re-estimate the machine temperatures. The rectangular cooling channel was adjusted to be 2 mm wide as shown in Figure 4-29, to replicate the actual cooling channel. Using MotorCAD, the temperatures distribution inside the machine

are shown in Figure 4-30. As can be seen from Figure 4-30, for a limited oil flow rate of 4 L/min at temperature of 90°C, the coil maximum temperature will be 189°C despite the increased copper losses. This is due to the improved volumetric cooling channel that increases the cooling area and thus improved the conduction path for the windings at the middle of the slot. Given that increase copper losses, showed no significant problems with the coil temperature. While, the temperature at magnet is under 120°C which is well below the limit temperature of magnet. Therefore, the machine can be well cooled by the oil flows available through the machine.

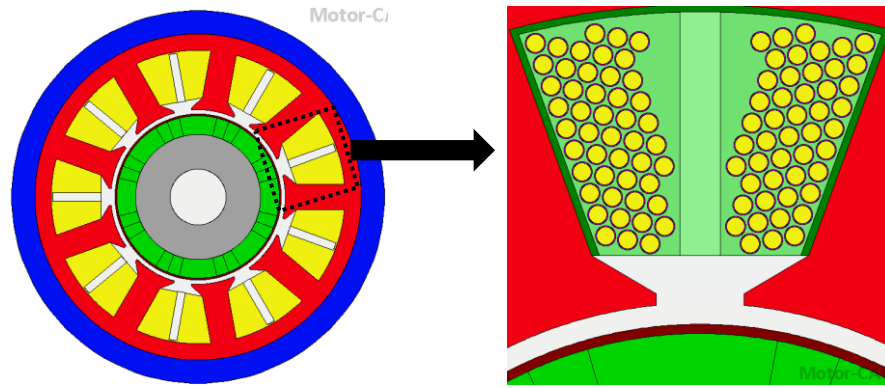


Figure 4-29: Cross-section area of the 8p/9s PMSM

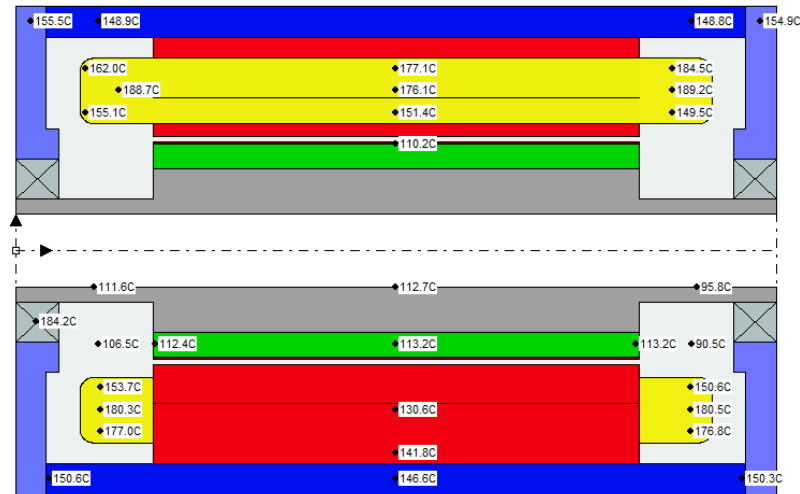


Figure 4-30: Temperature distribution in the machine for a condition 19000 rpm

F. Mechanical analysis and validation

The mechanical stress and rotor dynamics analysis have been performed to determine the rotor stability using a commercial software ANSYS. The mechanical analysis was carried on the final construction of the rotor to achieve high accuracy.

1) Mechanical stress

To operate the rotor safely at its maximum speed, the mechanical stresses that the structure can withstand need to be estimated. Figure 4-31 presents an axisymmetric structural FEA analysis model of the rotor used for the mechanical stress analysis. An interference fit of 40 microns (μm) between the sleeve and the magnets and shaft endcaps are used. A rotational speed of 20krpm has been assigned to add 5% of safety factor to the maximum speed of the machine. Figure 4-32 presents the equivalent (von-Mises) stress distribution in the rotor. It can be seen that the equivalent stress in different rotor components are lower than the yield strength of the materials used and summarized in Table 4-7.

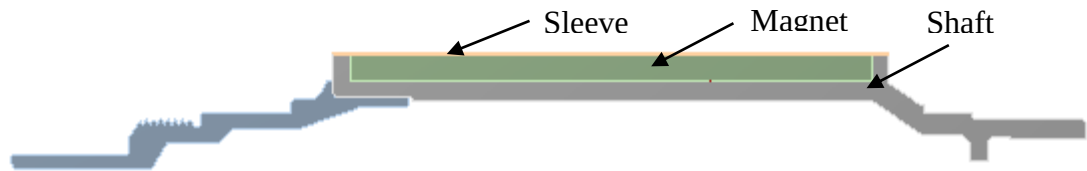


Figure 4-31: An axisymmetric structural FEA analysis model of the rotor

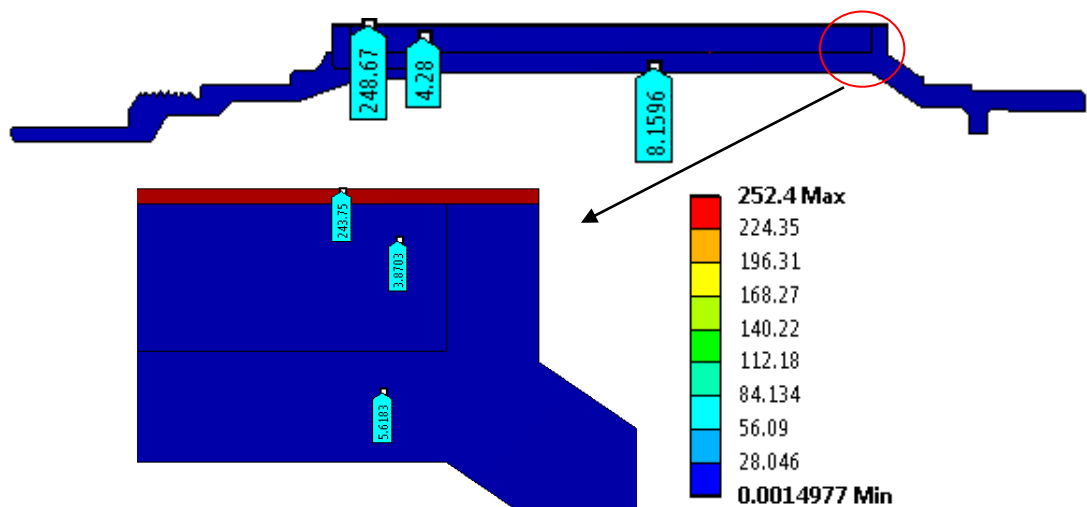


Figure 4-32: An equivalent (von-Mises) stress in the rotor at 20 krpm

Table 4-7: Maximum allowable stress limit for rotor components

Component name	Stress Limit (MPa)
Shaft	1379
Permanent magnet	35
Sleeve	825

2) Rotor dynamics

Another important mechanical aspect that is relevant in high speed machines is the rotor dynamic performance. The rotor speed should be kept away from the critical speed, otherwise serious vibrations would occur, and the machine might be destroyed [105]. In particular, the effects of unbalanced force due to inherited from machine topology are simulated and presented. A total 600 N maximum unbalanced force has been estimated from electromagnetic analysis with 10% safety margin and therefore shaft deformation and rotor dynamic performances are predicted under the unbalanced force conditions. Figure 4-33 shows the results of shaft deformation due to the 600 N force applied at the active part of the shaft. The maximum deformation of shaft is 0.003 mm at middle of the shaft. Comparing with 0.9 mm air gap between stator and rotor, the deformation due to unbalanced force is less than 0.3% and the effect is limited.

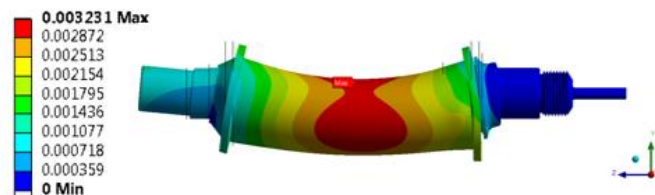


Figure 4-33: Shaft deformation due to unbalanced force

The shaft deformation due to unbalanced force could also affects the dynamic performances of rotor assembly. The bearing-rotor assembly dynamical performances are predicted for condition of with and without unbalanced force. It was found that for bearing stiffness of $5e7$ N/m at both ends of shaft, the frequencies of the first mode decreased from 2087.8 Hz to 1924.5 Hz. This is due to extra deformation of shaft caused by the unbalanced force. However, it's still 36% far from the machine operating speed of 19000 rpm. Therefore, the effects of the unbalanced force on rotor assembly dynamic performance can also be ignored. Figure 4-34 shows the first mode of rotor assembly under unbalanced force.



Figure 4-34: The first bending mode of rotor assembly

4.7 Conclusion

This chapter presents the optimization of a high performance PMSM design. A detailed procedure on the design and modelling of the final machine is given, illustrating the significant effort made to improve the machine performance. This is achieved by implementing several strategies such as the use of high performance material (CoFe) for the stator laminations and the adoption of Halbach array configuration for the PM. It was shown that by these techniques a power density of 4.87 kW/kg (including active and non-active weights) at the maximum speed can be achieved while still respecting the thermal limit of the machine. A main point to highlight is the power losses analysis illustrated in Section 4.5, which resulted in achieving an accurate estimation of the machine performance.

The final results presented and discussed in section 4.6, show an excellent performance achieved by the motor. From Figure 4-22 to Figure 4-34, it can be observed that the application requirements of Table 3-1 have been already reached.

In general, it can be concluded that the design procedures combined with the accurate modelling presented in this work have resulted in a machine which satisfies all the project targets.

The work related to this chapter has resulted in a journal publication [124] and a conference publication [125]. The main contribution to research from this chapter include

- 1) A compiling in one holistic process of all the different analysis required for electrical machines, including electro-magnetic, thermal, mechanical and general losses mapping. This optimisation and development process as described in this chapter can be used by other researchers to achieve a complete design of a high performance electrical machines.

Chapter 5. MANUFACTURING AND EXPERIMENTAL VALIDATION

5.1 Introduction

In the previous chapter, the procedure for the design of a flooded airgap, surface mounted PM machine for a direct coupled pump application was presented. This Chapter describes in detail, the manufacturing and assembly procedure related to the motor designed in Chapter 4 and its testing. The specific construction and assembly procedures are presented and investigated including a brief overview on different parts of the designed motor and the test-rig. Finally, the modelling results of Chapter 4 are compared to the experimental results obtained from the test-rig to validate the design procedures of the PMSM. After a detailed discussion of the obtained results, the conclusions for the chapter are stated.

5.2 Manufacturing and construction

In order to build up the PMSM of Chapter 4, the construction and assembly procedures are presented in this section. The geometrical dimensions of the PMSM are given in Table 4-2.

5.2.1 Stator construction

The stator core of the PMSM consist of a stacked Cobalt Iron (VacoFlux 48) laminations with thickness of 0.1 mm as shown in Figure 5-1. The stator core has a length of 80 mm with 9 slots.



Figure 5-1: Stator core of PMSM

5.2.2 Coils, cooling channels and oil compatibility

A. Coils

A double layer arrangement of concentrated coils has been used for the designed PMSM. For this configuration, 4 strands per turn has been selected, where the wire gauge is AWG 20 with diameter of 0.8735 mm and bare wire (copper diameter) of 0.8075 mm. There are 4 parallel strands per turn and 12 turn per coil connected in series, leading to 96 total number of conductors in one slot. A number of thermocouples (type k), placed within both the end-winding and in the middle section of the winding for each phase, were used to monitor the temperatures.

B. Cooling channels

As stated in Chapter 4, the PMSM is designed to operate with a high current density, therefore, a cooling channel has been created inside the slot. The cooling channels inside each slot are made of air, therefore, no additional or specific material for cooling purposes is placed within the slots. In other words, the cooling channels have been obtained during the winding procedure of the stator core, by leaving enough space between the two coil sides belonging to the same slot, as shown in the Figure 5-2 (c) (i.e. red circles). This was physically done by having plastic spacers in place during the winding process and then removing them gently once the winding has been completed as shown in Figure 5-2 (a and b). Thanks to this “free space” between the two coil sides

of the same slot (i.e. cooling channel), a passage for the cooling oil is created. This cooling strategy allows a direct contact between cooling oil and winding surface, which enhances the cooling capability. Indeed, the heat produced within the slot (due to the Joule losses) is directly extracted from the windings.



Figure 5-2: Stator core with winding and details of the cooling channel

C. Oil compatibility

The chemical compatibility between the cooling oil and the slot insulation materials (i.e. coating wire insulation and slot liner) has been carefully considered at the design stage. Since, the chemical deterioration and/or corrosion will affect the insulation lifetime leading to the winding failures. For this reason, compatible cooling oil and

insulation materials have been selected, in order to avoid and/or minimize potential impacts on the winding lifetime.

In particular, the selected copper wire uses a Polyester-based insulation coating, which is designed for oil flooded applications. Its chemical compatibility with the selected cooling oil has been verified together with both copper wire and cooling oil manufacturers. The adopted slot insulation (i.e. liner) is represented by a “sandwich structure” made of Mylar tape, which is covered on both sides with Nomex. This choice ensures chemical compatibility with the cooling oil and enhances the electrical insulation between copper wire and stator core.

The corrosion risk of the wire insulation has been also assessed and the AEROSHELL FLUID 31 has been selected as a cooling oil [126]. This cooling oil is formulated with high technology additives, which prevent both oxidation and corrosion.

5.2.3 Rotor construction

Figure 5-3 (a and b) shows the rotor construction with PMs, quasi-Halbach array on the shaft without and with retaining sleeve. The rotor magnet is Samarium Cobalt permanent magnets (Recoma 33E (Sm₂Co₁₇)). The selected material for the rotor sleeve is Titanium Ti6A14V designed with a thickness of 0.4 mm, to reduce the moment of inertia and increase the airgap width. The permanent magnets are segmented axially into 10 segments to reduce the eddy current losses.

A hollow shaft made of 17-4 PH Stainless material is designed to reduce mass and meet the moment of inertia requirement. The rotor is supported by bearings at each end. In particular, a deep groove ball bearing is used at the drive end, while at the non-drive end, a pair of back to back bearings is adopted for increasing the rotor stability. While, an internal spline coupling is designed at the drive end where the hydraulic pump is connected. For the preliminary tests in the of University of Nottingham, a shaft spline coupling is designed to connect the motor to the load motor via torque meter as shown in Figure 5-3 (c).



(a) Without sleeve



(b) With sleeve



(c) Shaft spline

Figure 5-3: Rotor shaft construction of PMSM

5.2.4 Housing construction

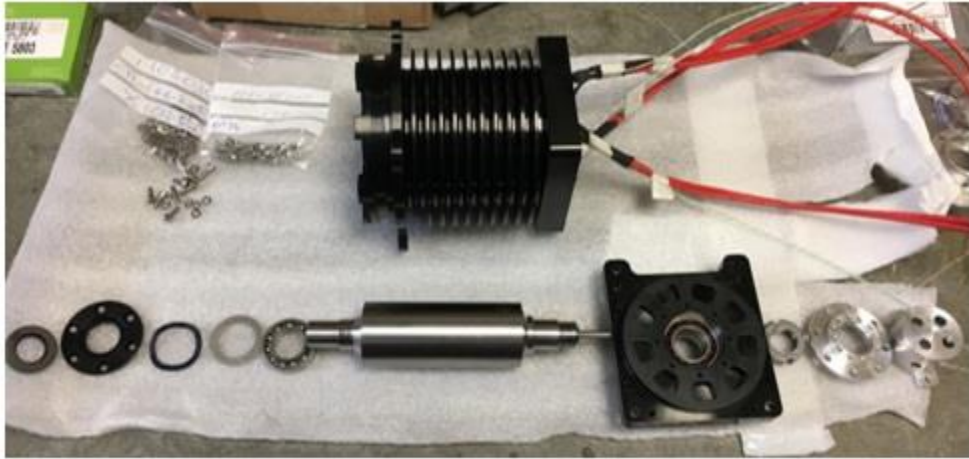
Figure 5-4 shows the housing construction of the PMSM. The machine housing is designed to be oil flooded. Hydraulic oil from the pump flows into motor at the front of the machine housing passes through the gap and cooling channels inside the machine and flows out at the rear end. A fin array is attached to the machine housing, so as to enhance the heat dissipation to air. The machine housing then serves as an additional heatsink for the whole pumping system.



(b)



(b)

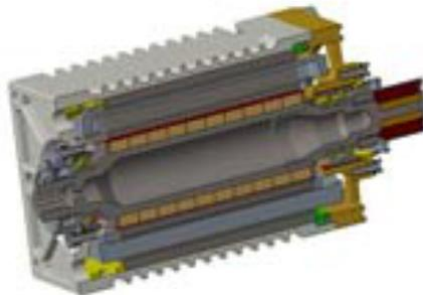
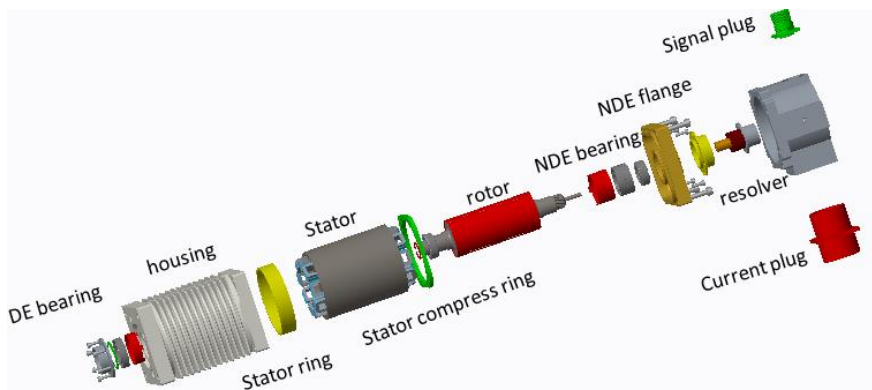


(c)

Figure 5-4: Housing construction of PMSM

5.3 Overall structure and assembly

A cross-sectional view of the flooded PMSM assembly is shown in Figure 5-5 (a), while an exploded view is given in Figure 5-5 (b).

*(a) Cross-sectional view**(b) Exploded cross-sectional view**Figure 5-5: PMSM final motor assembly*

The complete motor assembly, which is shown in Figure 5-6, is then assembled on a high speed test rig. Figure 5-7 shows the test rig equipment used in this work. The prime mover or (load motor) is connected to the rig using flexible high speed coupling on one side, on the other side the PM motor is connected to torque meter through a shaft splined. The load motor is a standard industrial (DPM17100CM brushless servo motor), 25 kW, capable of 20 Nm continues load and maximum speed of 30000 rpm. The torque is measured using a high precision torque meter (Magtrol torque-meter TMHS 309-111) with a maximum rating of 50 Nm with an accuracy of 0.1%. Figure 5-8 shows the equipment used in this work.

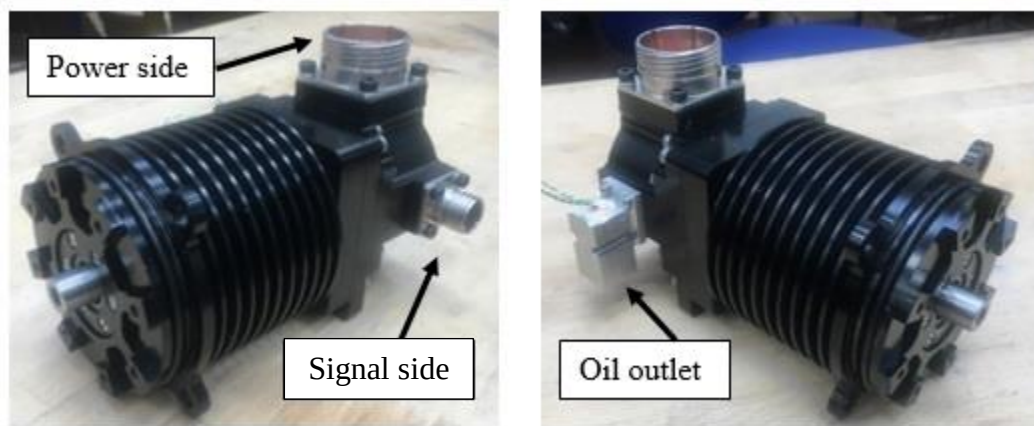


Figure 5-6: PMSM: final prototype

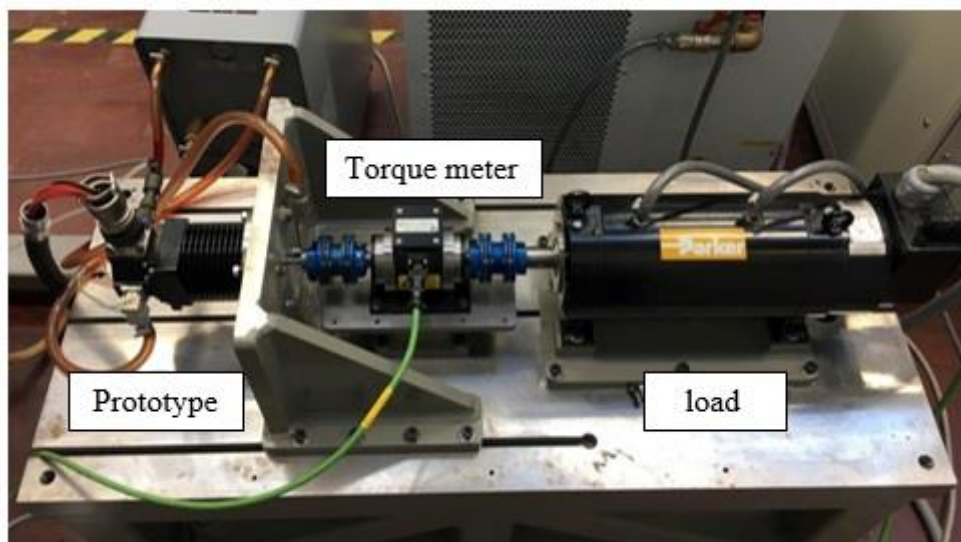


Figure 5-7: Experimental test bench



(a) DC power supply



(b) Power converter



(c) Water cooled chiller



(d) Oil cooling unit



(e) Speed control panel



(f) Measuring equipment

Figure 5-8: Test rig set-up high

5.4 Experimental measurements and validation

In order to validate the performance of the prototype machine, a series of tests were carried out. These are presented and described below.

5.4.1 Resistance measurements

The stator resistance is an important factor because it determines the copper losses and thus the total PMSM efficiency. Accurate measurements of a stator resistance need to consider temperature and frequency dependence and the manufacturing unknown variables.

A. DC resistance measurements

The DC winding resistance of each phase has been measured by using a Milli-ohmmeter Valhalla Scientific 4300B and the obtained values are listed in Table 5-1.

Table 5-1: DC resistance measurments using the Valhalla Scientific 4300B and FEA

Resistance	Measured	FEA	
	Valhalla Scientific 4300B	2D-FEA	3D-FEA
Phase A (mΩ)	67.31	50.9	63.49
Phase B (mΩ)	67.38	50.9	63.49
Phase C (mΩ)	67.44	50.9	63.49
R_{avg} (mΩ)	67.37	50.9	63.49

These parameters are measured at ambient temperatures. As can be observed, good similarity exists between the measured and FE results with error of 6.11%. This difference in results is due to fact that the FE results does not take into account the length and the resistance of the connection leads.

B. AC resistance measurements

Under AC conditions, the winding resistance value is affected by skin and proximity effects, which increase the resistance value. In order to take into account, the influence due to the skin and proximity effects, the AC resistance is measured using an LCR Bridge meter (HM8118). This method uses the four-wire method with two wires injecting a known current and the other two connected to a voltmeter. The obtained results are reported and compared to the predicted FE results in Table 5-2, while Figure 5-9 shows the AC resistance trend as function of the frequency.

Table 5-2: AC resistance measurements using the LCR Bridge HM8118 and FEA

F (Hz)	Measured				FEA	
	Phase A (mΩ)	Phase B (mΩ)	Phase C (mΩ)	R_{avg} (mΩ)	2D-FEA	3D-FEA
200	67.68	68.83	67.93	68.16	51.03	63.64
400	68.99	69.09	69.05	69.04	51.6	64.24
600	71.59	70.56	71.14	71.13	52.5	65.26
800	73.602	73.88	74	73.83	53.85	66.63
1000	76.96	76.84	77.61	77.14	55.54	68.3
1200	81.08	79.804	81.91	80.94	57.6	70.53

The last column of the measured values of Table 5-2 reports the line-to-neutral AC resistance (phase resistance) average values, which are determined by using the equation, shown in (5-1)

$$R_{phase_avg} = \frac{R_A + R_B + R_C}{3} \quad 5-1$$

Summarizing, the obtained results are:

- The measurements of AC resistance have been conducted at ambient temperature. Compared to the FE results almost no variation at different frequencies (in the range 0 Hz – 400 Hz).
- At 580 Hz (base speed), the AC phase resistance is 8.85% higher than the 3D FE result.

- At 1200 Hz the AC phase resistance is 14.76% higher than the 3D FE result. The frequency of 1200 Hz is considered, since it is the closest value to 1266 Hz, which is the frequency at maximum speed (19000 rpm).

The higher values of the measured phase resistance at maximum frequency can be attributed to the accuracy of LCR bridge meter.

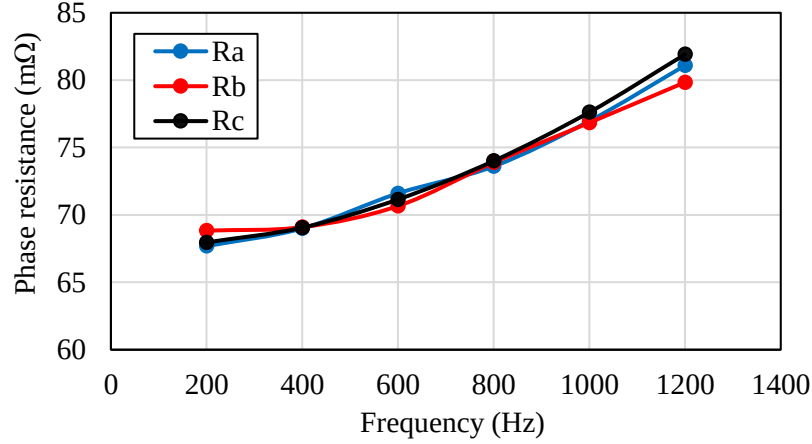


Figure 5-9: Phase resistance as function of frequency (LCR bridge measurements)

C. Temperature dependency

The total winding resistance is a sum of the active-length and end-winding components and can be expressed as given in (5-2), where, $R_{act}(T, f)$ and $R_{end}(T, f)$ are the active length and end-winding resistances under the operating temperature and frequency, respectively.

$$R_{tot}(T, f) = R_{act}(T, f) + R_{end}(T, f) \quad 5-2$$

Under DC operation, the resistance can be corrected easily to consider the influence of the temperature. For this analysis, the electrical resistivity of the copper can be corrected considering equation (5-3) for both the active length and end-winding to account the temperature variation in FEA, where ρ_o is the electrical resistivity of copper at $T_o = 20^\circ\text{C}$ and is $1.68 \cdot 10^{-8} \Omega m$, α is the temperature coefficient of resistivity and is 0.00393 and T is the temperature at operating point.

$$\rho = \rho_o * (1 + \alpha(T - T_o)) \quad 5-3$$

However, in case of AC operation the expression given in (5-3) cannot be used directly to correct the operating temperature, since, the DC and AC resistance components are depending on the temperature in different manner [60]. Therefore, to investigate the variation of DC and AC resistance with temperature, a variable-frequency power supply with sinusoidal current (Chroma 61511) was used to energise the winding under test. The power, voltage and current were measured using a high bandwidth power analyzer (KinetiQ PPA2530), which allows to take into account the resistance variation with frequency. A number of thermocouples (type k), placed within both the end-winding and in the middle section of the winding, were used to monitor the temperatures. The experimental tests have been carried out on a complete wound stator as shown in Figure 5-10. While, Figure 5-11 shows the setup used to measure the winding losses.



Figure 5-10: Analysed stator / winding assembly

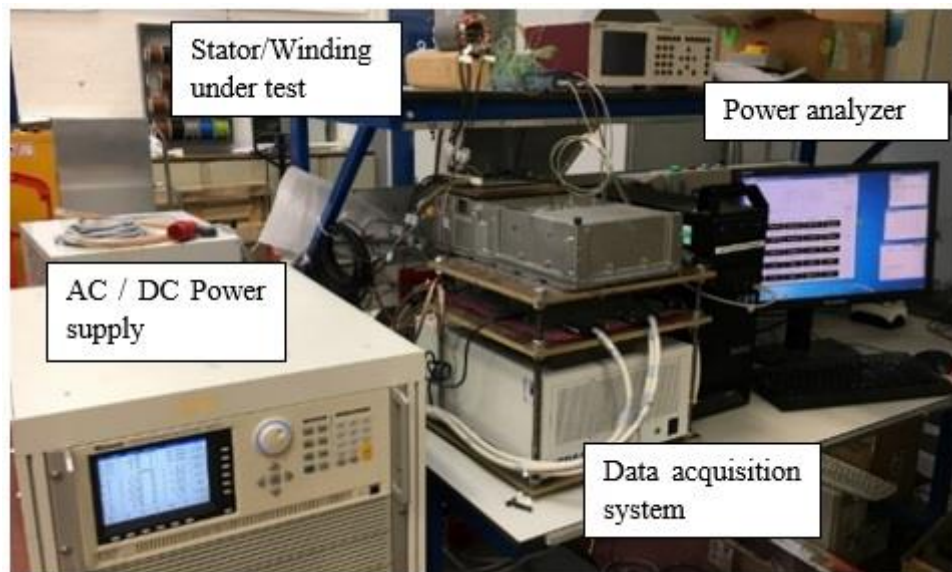


Figure 5-11: Experimental setup

In order to obtain the DC and AC phase resistance R_{dc} and R_{ac} , a high-bandwidth power analyser was used to measure the input power per phase and the rms current. However, it is not possible to experimentally separate laminated stator core losses from measured power (i.e. copper losses + iron losses) using a power analyser. Therefore, accurate estimation of iron losses was required. For this purpose, 3D-FE simulations of the stator core were performed using the same current and frequency values used during the experiments.

In order accurately estimate the phase resistance, the effect of core losses P_{fe} (calculated by 3D-FEA) is subtracted from the total measured power P_{in} , as shown in (5-4).

$$R_{ac} = (P_{in} - P_{fe})/I_{rms}^2 \quad 5-4$$

Further, the winding temperatures are recorded at several excitation frequencies. The thermal test for a single excitation point is performed until the thermal steady state is reached [61]. Hence, the thermal test is stopped once the winding temperature varies with a ratio lower than 1°C over 10 minutes. The variation of the winding temperature (hot-spot) due to the frequency under different currents is shown Figure 5-12.

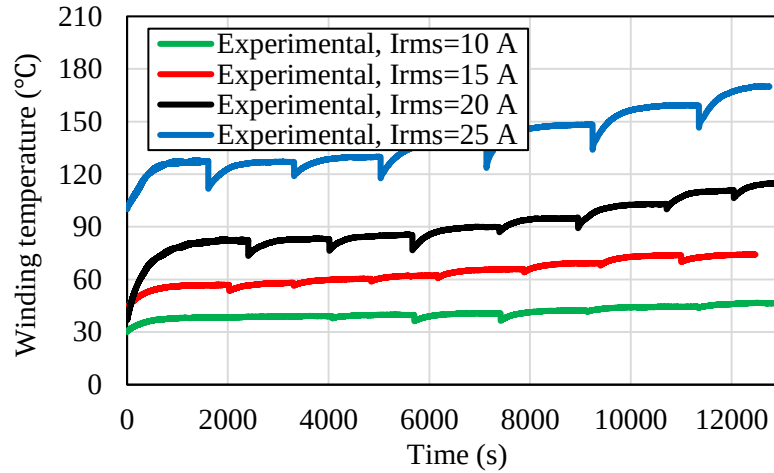


Figure 5-12: Winding temperature under $I_{rms}=25$ A with different frequencies

In Figure 5-13, the link between the DC phase resistance and the winding temperature is highlighted, whereas the dependency of the AC phase resistance with the frequency and temperature is outlined in Figure 5-14. It is obvious that the DC resistance change

linearly with temperature as shown in Figure 5-13. This is due to fact that the copper resistivity changes linearly with temperature as expressed in equation (5-3). Looking at Figure 5-14, the AC phase resistance value increases with the frequency, as expected.

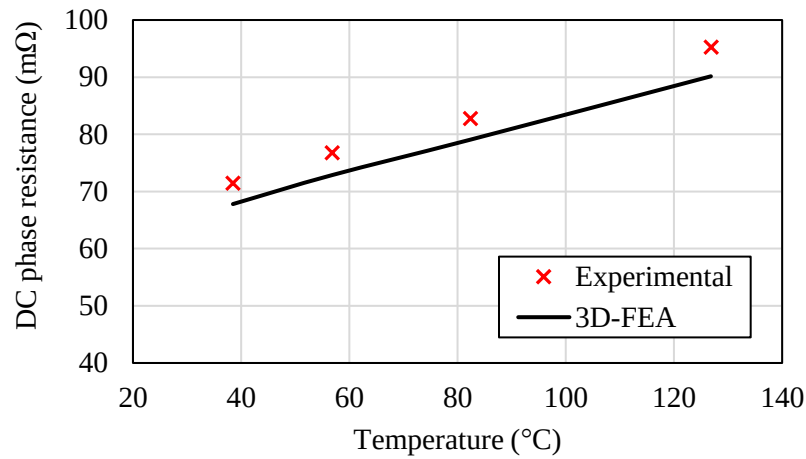


Figure 5-13: DC phase resistance versus temperature

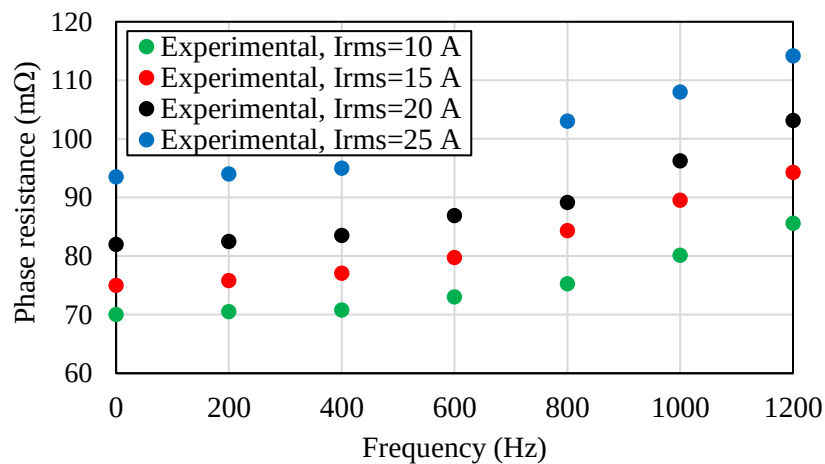


Figure 5-14: Phase resistance versus frequency

5.4.2 Inductance measurements

Another factor which plays a critical role in the control strategies of the PMSM is the machine inductance. The AC standstill method is commonly used for the inductance measurements [127]. This method is based on the indirect measurements of the instantaneous flux linkage. Figure 5-15 Show the schematic of the inductance measurements.

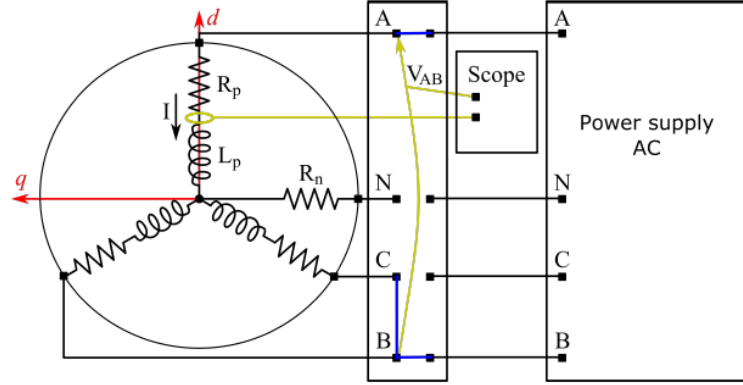


Figure 5-15: Oscilloscope and AC power supply connection for L_d inductance measurement.

In this test, the L_d and L_q inductances are calculated from the measured terminal voltage V_{ab} and the measured current I with the rotor locked in two different positions. The alignment of the d-axis with phase A is obtained by supplying a DC current between terminal A and terminals B and C shorted as shown in Figure 5-16. While, the alignment of the q-axis with phase A is obtained by supplying a DC current between terminal B and C with terminal A floating as presented in Figure 5-17. The rotor should be locked in position before measuring the inductance so that additional induced voltages due to changes in the magnetic fields associated with each phase are prevented [128].

The flux linkages in d and q axis, ψ_d and ψ_q can then be calculated by integrating the measured voltages as given in (5-5) and (5-6), where R is the stator winding phase resistance and i_d and i_q are the d and q axis currents.

$$\psi_d = \int (v_d - Ri_d)dt \quad 5-5$$

$$\psi_q = \int (v_q - Ri_q)dt \quad 5-6$$

The v_d and v_q are directly proportional to the measured voltage V_{ab} because both of v_b and v_c are equal as shown in Figure 5-15 and can be calculated as shown in (5-7) and (5-8), where θ is the position of the rotor and it's constant since the rotor is locked.

$$v_d = \frac{2}{3}(V_{ab}) \cos \theta \quad 5-7$$

$$v_q = \frac{2}{3}(V_{ab}) \sin \theta \quad 5-8$$

From (5-5), when the rotor is in d-axis position, the d-axis inductance can be calculated as:

$$L_d = \frac{\psi_d - \psi_{PM}}{i_d} \quad 5-9$$

Similarly, the q-axis inductance can be calculated from (5-6), when the rotor is in the q-axis position as:

$$L_q = \frac{\psi_q}{i_q} \quad 5-10$$

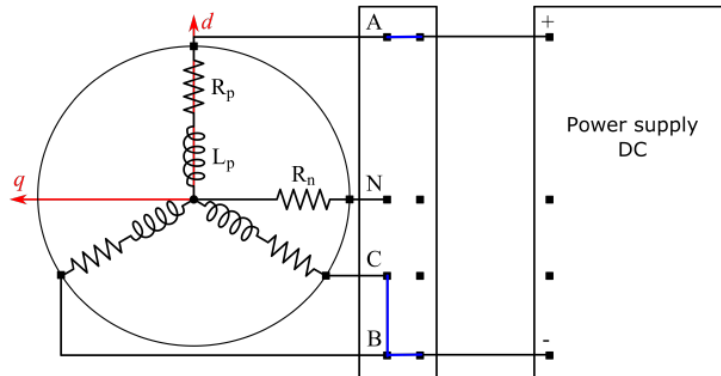


Figure 5-16: DC power supply connection for d-axis alignment

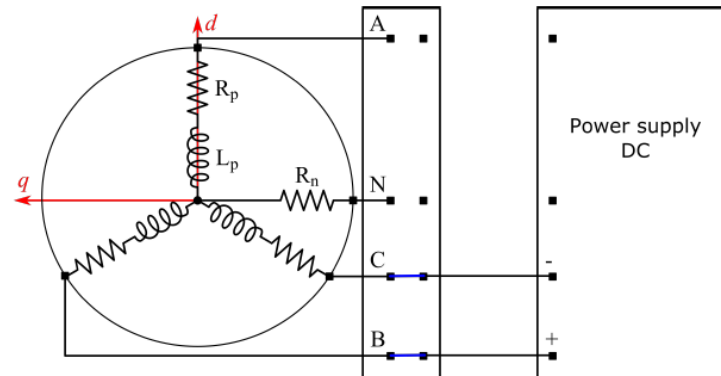


Figure 5-17: DC power supply connection for q-axis alignment

The experimental setup used for measuring the inductance of the prototyped PMSM using AC standstill test is presented in Figure 5-18. The test rig has been modified by including a clamp for the alignment of the rotor shaft as shown in Figure 5-18. The measurements were done at 580 Hz (based speed) with an RMS currents of (5, 15, 25) A. Figure 5-19 shows the measured voltage and current obtained when the magnet axis is aligned along phase a-axis at 580 Hz and 25 A. While, the calculated flux linkage from the measured voltage and current is presented in Figure 5-20.

Figure 5-21 and Figure 5-22 show the measured inductances and the 3D and 2D-FEA results, from which the influence of magnetic saturation can be clearly observed. Due to the end-windings, higher inductances are observed in the 3D-FEA and the measurements compared to its 2D-FEA counterpart. With an RMS current of 25 A, there is a difference of 16.82% between measurements and 2D-FE results for the d-axis inductance at positive d-axis currents and zero q-axis current. This difference is drop to 3.14% when the measured results are compared to 3D-FEA results. Also, with positive d-axis current (pro-magnetizing), the stator mmf causes an increase in the resulting airgap flux density resulting higher saturation at high currents. While, for the q-axis inductance with positive q-axis current and zero d-axis current, the measured inductance is 0.33 mH. The obtained L_q inductance from measurements are 20.14% and 6.58% higher than the one calculated using 2D-FEA and 3D-FEA, respectively.

In general, the difference is mainly due to the leakage flux at the ends of the machine. As can be seen, there is a good matching between the 3D-FE results and the measurements.

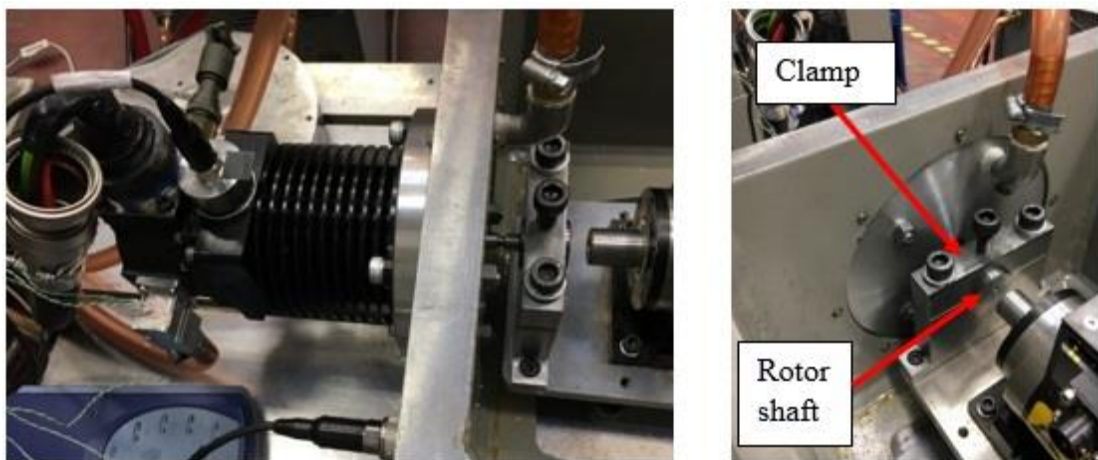


Figure 5-18: Experimental setup for the d and q-axis alignment

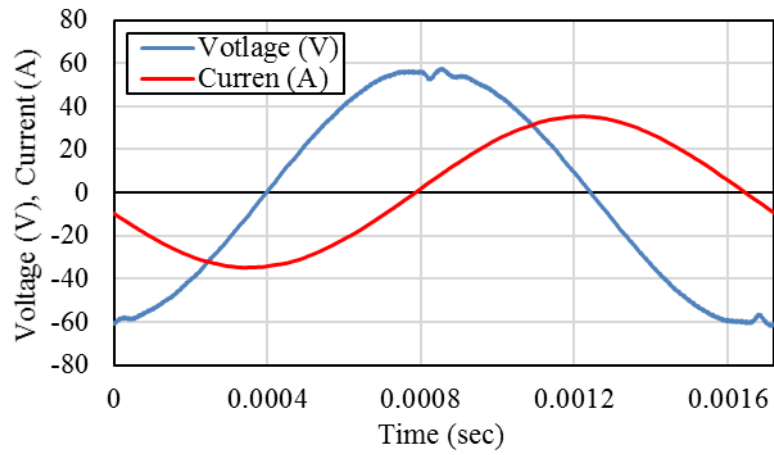


Figure 5-19: Measured voltage and current waveforms

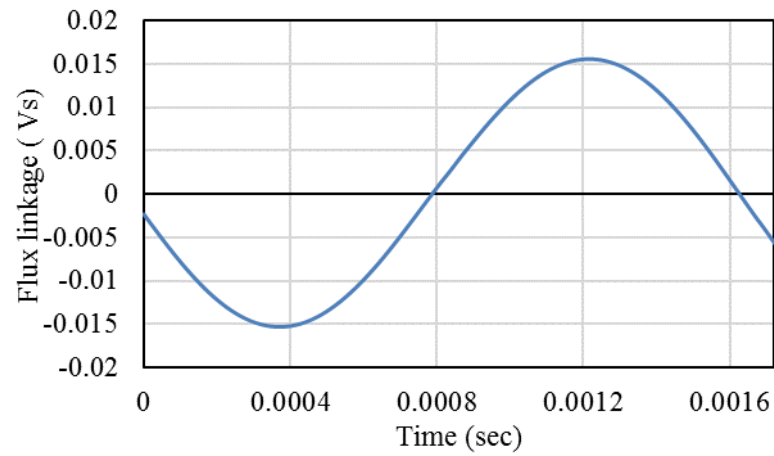


Figure 5-20: Measured flux linkage waveform

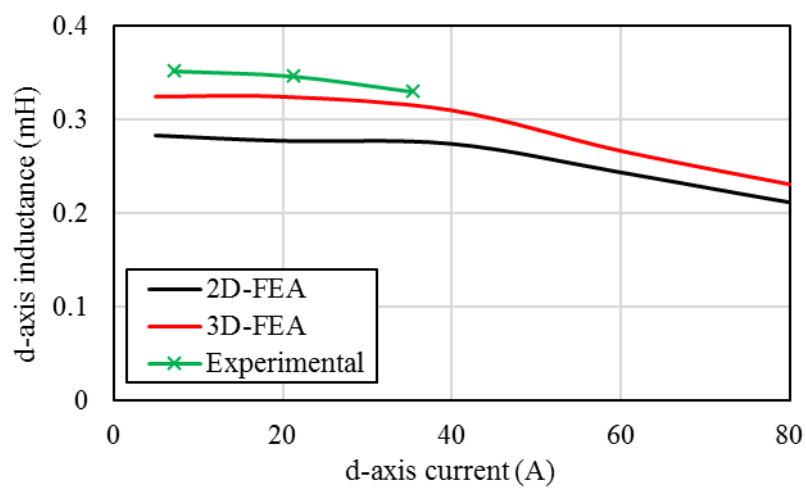


Figure 5-21: Measured and predicted d-axis inductance

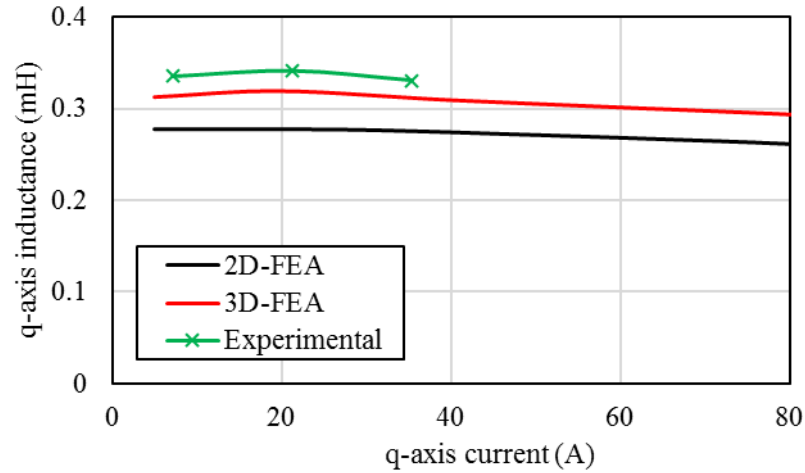


Figure 5-22: Measured and predicted q -axis inductance

5.4.3 Back-EMF measurements

One of the most important PMSM parameters is the back-EMF, because it affects the electromagnetic torque of the machine. For the no-load test, the motor has been mechanically coupled to the load motor (Parker). During the test, the Parker motor is speed controlled with constant speed reference, while three differential voltage probes have been used to measure the line-to-line back EMFs. The back-EMF is measured and recorded via a high bandwidth (MSO-X 2024A Oscilloscope).

A. No load results at 8700 rpm

The measured line-to-line back EMFs at 8700 rpm (base speed) are shown in Figure 5-23, while the comparison between the measured results and the FEA results are presented in Figure 5-24. Figure 5-25 reports the back EMF spectrum, while the detailed results and the comparison are shown in Table 5-3.

Based on the above results, the waveforms of the line-to-line back EMF are quite similar and the difference between the measured value and the FE results is small as shown in Figure 5-25. This is also confirmed by the harmonic content of the measured waveform showing a THD of 0.577% at 8700 rpm compared to 0.145% and 0.211% for the 2D and 3D FEA, respectively.

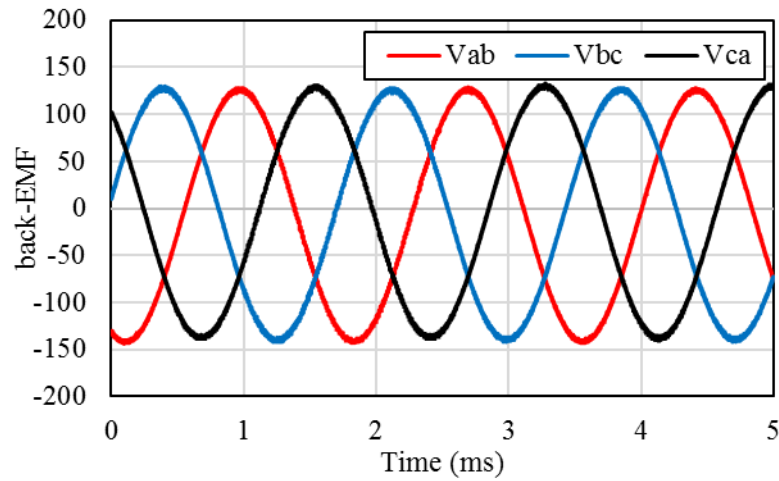


Figure 5-23: No load line to line Back-EMF waveform at 8700 rpm

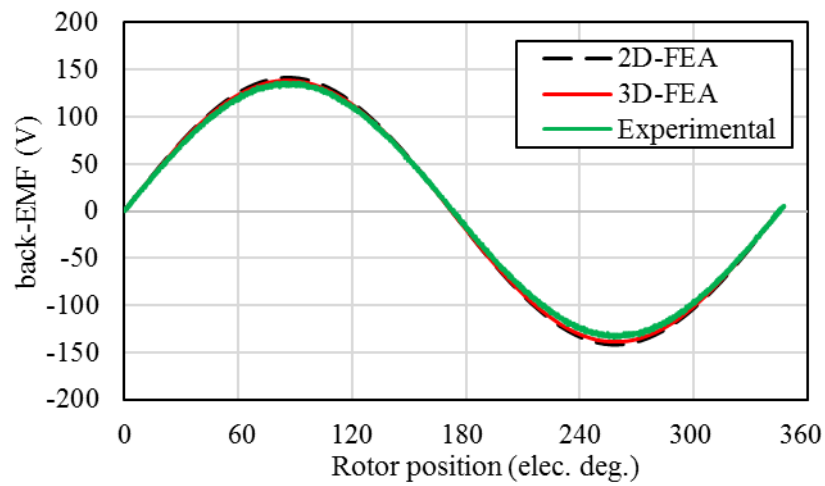


Figure 5-24: Comparison of the no load line to line back-EMF waveform

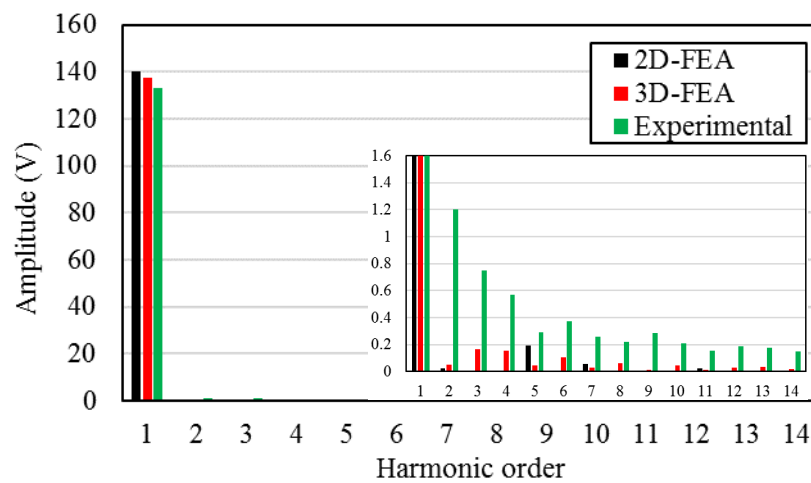


Figure 5-25: FFT of the line-to-line back EMF (between phase A and B)

Table 5-3: No-load results at 8700 rpm

Speed 8700 rpm					FEA	
	Phase AB	Phase BC	Phase CA	Phase Avg	2D FEA	3D FEA
V_{pk} (V)	133.11	133.59	132.41	133.03	139.2	136.4
K_{v_mech} (V/rpm)	0.0153	0.01535	0.0152	0.0152	0.016	0.0156
K_{v_el} (V·s/rad)	0.0365	0.0366	0.0363	0.0364	0.0381	0.0374
ψ_{PM} (Wb)	0.021	0.0211	0.0210	0.0210	0.0219	0.0216
THD (%)	1.3	0.433	0.936	0.577	0.145	0.211

B. No load results at 19000 rpm

The measured line to line back-EMF waveform at the maximum operating speed of 19000 rpm is shown in Figure 5-26. The comparison between the measured results and the one obtained from FE, are reported in Figure 5-27 while, the harmonic content is shown in Figure 5-28. The detailed results and the comparison are shown in Table 5-4.

A good match can be observed between the measured and the calculated back-EMF waveforms which are sinusoidal as shown in Figure 5-27. This is also confirmed by the harmonic content of the measured waveform showing a THD of 0.6% at 19000 rpm compared to 0.149% and 0.211% for the 2D and 3D FEA, respectively.

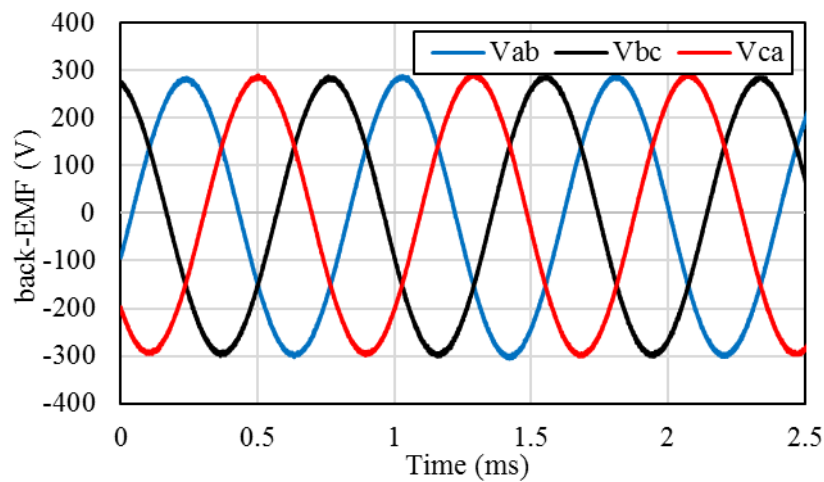


Figure 5-26: No load line to line Back-EMF waveform at 19000 rpm

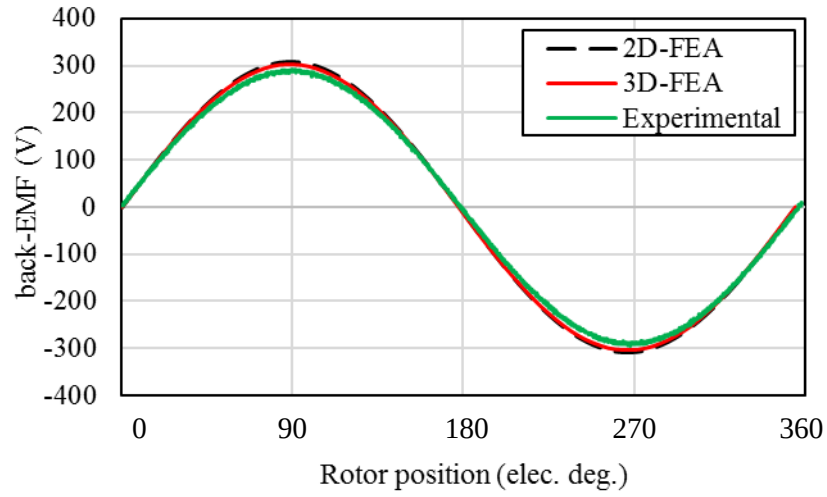


Figure 5-27: Comparison of the no load line to line Back-EMF waveform

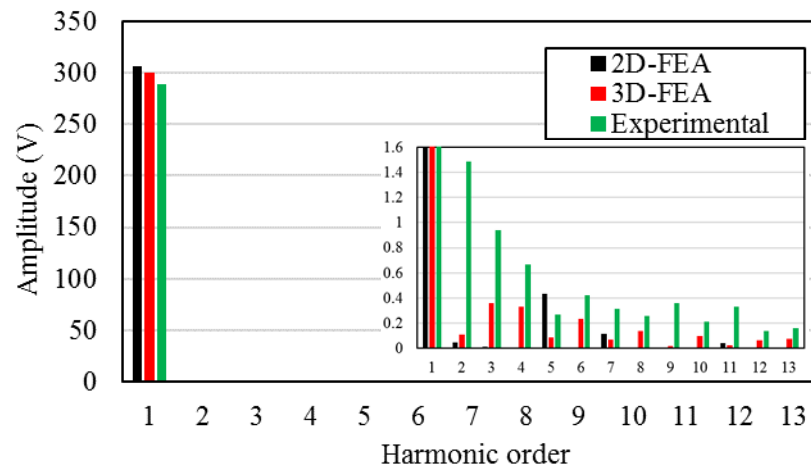


Figure 5-28: FFT of the line-to-line back EMF (between phase A and B)

Table 5-4: No-load results at 19000 rpm

Speed 19000 rpm					FEA	
	Phase AB	Phase BC	Phase CA	Phase Avg	2D FEA	3D FEA
V_{pk} (V)	288.75	289.9	289.6	289.41	310.4	299.84
K_{v_mech} (V/rpm)	0.0151	0.0152	0.0152	0.0151	0.0160	0.157
K_{v_el} (V·s/rad)	0.0362	0.0364	0.0363	0.0363	0.0384	0.0376
ψ_{PM} (Wb)	0.0209	0.02103	0.021	0.0209	0.221	0.0217
THD (%)	0.727	0.579	0.495	0.6	0.149	0.2115

C. No load overall results

The No-load tests have been performed at several operating speeds from 1000 rpm to 19000 rpm using 1000 rpm step and the obtained results are plotted in Figure 5-29. The tested results are in good agreement with the 2D and 3D FE simulations, although the measured values of the Back-EMF are slightly low at the maximum speeds with K_v 0.0363 V·s/rad. The obtained K_v from measurements are only 5.46% and 3.45% lower than the one calculated using 2D-FEA and 3D-FEA, respectively.

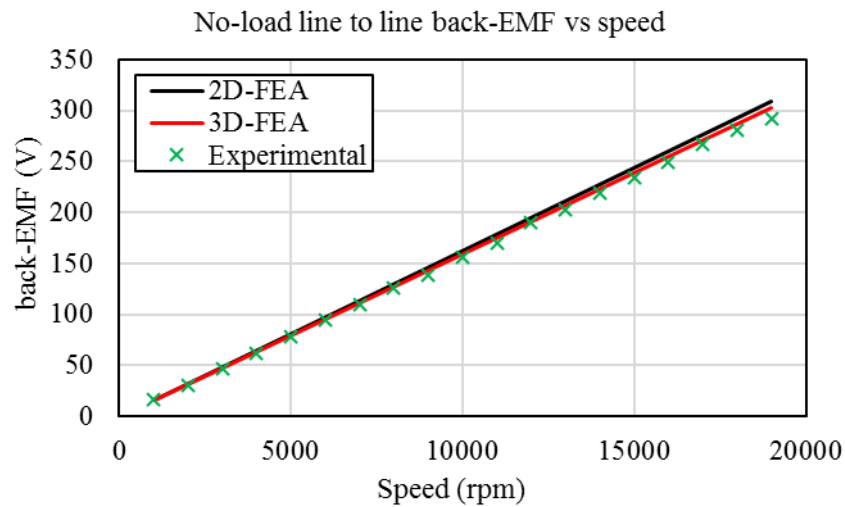


Figure 5-29: Amplitude of the line-to-line back EMF as function of the speed

5.4.4 Windage losses measurements

The windage losses has been investigated experimentally in this section to provide an accurate estimation for the machine performance. The standard open circuit test method cannot be used to segregate the windage losses with permanent magnet machines due to the existence of a constant magnetic field. Therefore, the direct test method is a preferred solution for the windage losses calculation due to simplicity and high accuracy. Normally, for the direct test method, either a dummy rotor without permanent magnets or stator with a non-ferromagnetic material or a completely dummy model (stator and rotor) can be used. An external driver was used to drive the motor and the mechanical losses will be measured by measuring the torque and speed of the rotor. In this case, the bearing losses will need to be subtracted to obtain only the windage losses.

In this test both the stator and the rotor should simulate accurately the geometry of the electrical machine.

Figure 5-30 shows the components of the dummy motor which will be used for the windage losses investigation which replicate the actual dimensions of the 8p/9s PMSM. The assembled dummy motor has been instrumented with a number of type K thermocouples (mineral insulated probe). These thermocouples were placed at three different locations inside the airgap (middle, inlet and outlet) by drilling some holes into the airgap and one near to the airgap surface to measure the temperature at the stator surface facing the airgap. In addition, thermocouples were located also in the chamber at the front and rear end plates. These thermocouples were fixed by thermocouple compression fitting to provide a flexibility in placing the thermocouples as shown in Figure 5-30. While, Figure 5-31 shows the experimental test rig for windage losses measurements including the dummy motor, torque transducer and the drive machine.

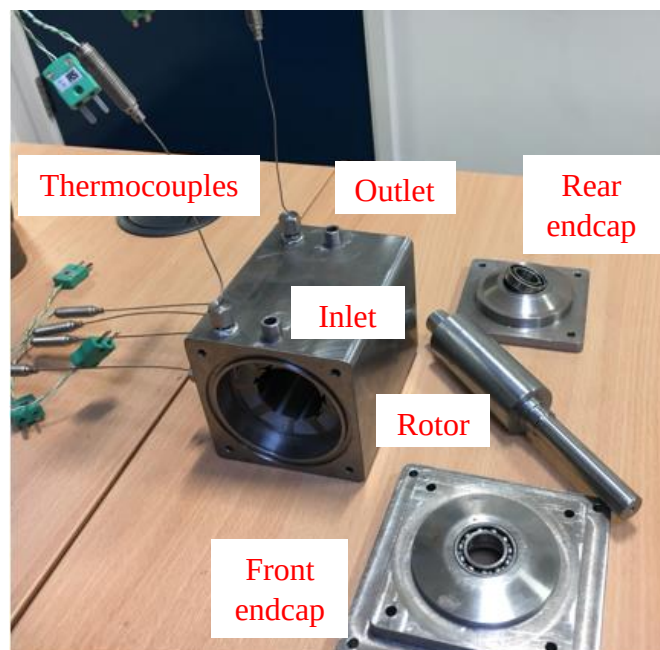


Figure 5-30: Dummy motor components for windage loss measurement.

For the experimental setup, it is important to make sure that the accuracy of the measuring system is sufficient for measurements of the windage losses. Therefore, high accuracy torque transducer with measurement error $\pm 0.1\%$ of the rated torque were used

to measure directly the torque and speed transmitted to the dummy motor under windage losses test. For the rated torque, the measurement error was translated into ± 0.02 Nm. The uncertainty of the temperature measurement using thermocouples were considered. The error in the temperature measurement using type k thermocouples is expected to be $\pm 1.5^\circ\text{C}$ for the measured temperature range. A cooling system with controllable flow rate and temperature has been used in the investigation. The temperature condition of the coolant flow inside the airgap may different with different speeds and conditions tests and this will affect the measured value of the windage losses. Therefore, to make the measured results comparable, the temperature of the coolant flow inside the airgap has been fixed to a reference temperature of 90°C . The dimensions of the rotor and stator enclosure used in this experiment are shown in Table 5-5.

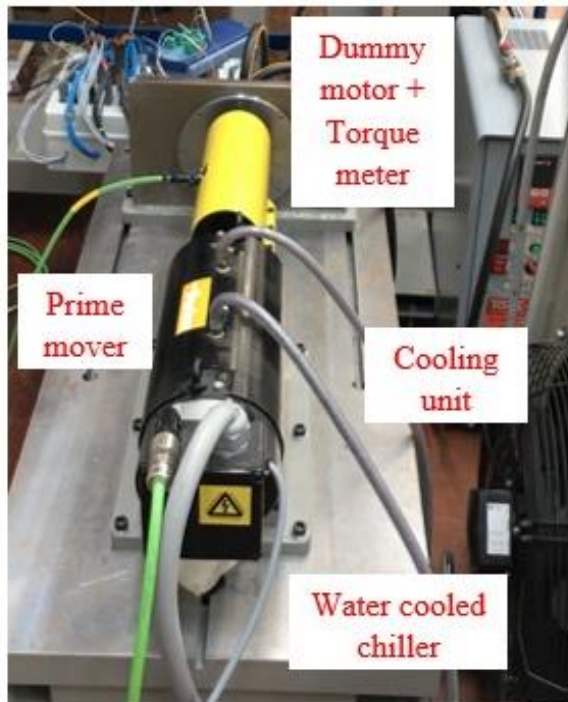


Figure 5-31: Windage losses experimental setup

Table 5-5: Rotor shaft dimensions

	Values
Rotor radius (r) mm	17.825
Stator inner diameter (D_{si}) mm	37.45
Airgap width (L_g) mm	0.9
Rotor length (L_{ax}) mm	80

Initial tests have been applied on the dummy stator with slot opening (axially slotted) and dummy rotor to evaluate the windage losses. For the initial test, the temperature was controlled by the cooling unit and the flow rate was 4.2 L/min. Figure 5-32 shows the measured windage losses obtained from tests with the dummy motor. While, the properties of the coolant oil were presented in Table 5-6. It can be seen that, there is a difference in results between the measured data and the calculated data at low speeds and this difference becomes higher at the maximum speed of 17000 rpm. For the top speed of 17000 rpm, it is clear that the error is approximately 3.66% lower than that obtained by the analytical method. The reason is that the measured windage losses of the dummy motor is produced not only by rotor but also by the bearings which included in the calculation based on the datasheet provided by the manufacturer which is of course not the actual case. Also, the analytical calculations are based on empirical coefficients which have been developed according to the accumulated experience from the experimental studies.

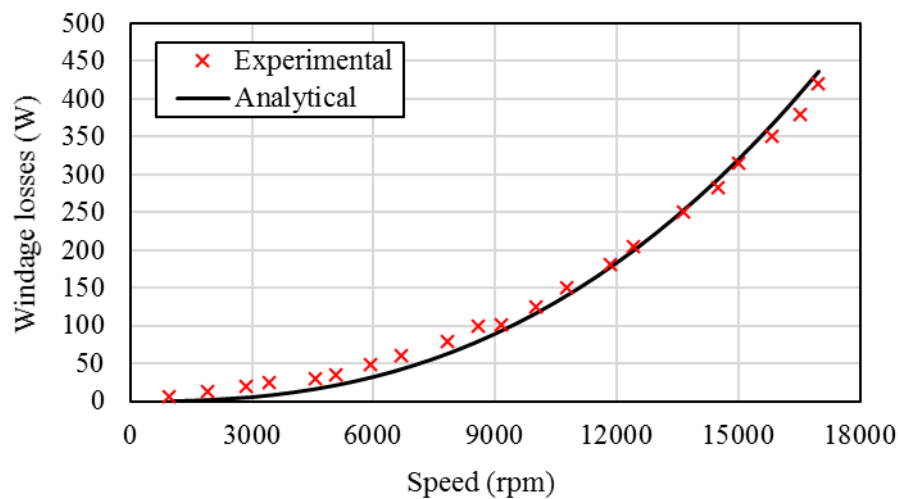


Figure 5-32: Measured windage losses versus speed

Table 5-6: Physical properties of the coolant oil

Properties	Values
Mass density (kg/m ³)	951.6
Dynamic viscosity (kg/m/s)	0.0047
Specific heat capacity (J/kg/°C)	1968
Thermal conductivity (W/m/°C)	0.14

5.4.5 Characteristics torque vs current measurements

The aim of this test was to characterise the prototype and its torque linearity. The torque constant K_t has been determined through experimental tests. During this test, the load motor (IPM motor) was under speed control, with a set point equal to 100 rad/s, 200 rad/s, 300 rad/s and 910 rad/s (base speed). The designed motor was under current control and coupled to the Parker motor. The current reference has been increased from 0 A to 80 A. For each value of axis current, the developed torque has been measured using a Magtrol torque-meter TMHS 309-111 at thermal steady state. The obtained results are shown in Figure 5-33. Looking at Figure 5-33, it is clear that the electromagnetic characteristic of the machine is linear for the whole range being investigated and thus validates the machine design.

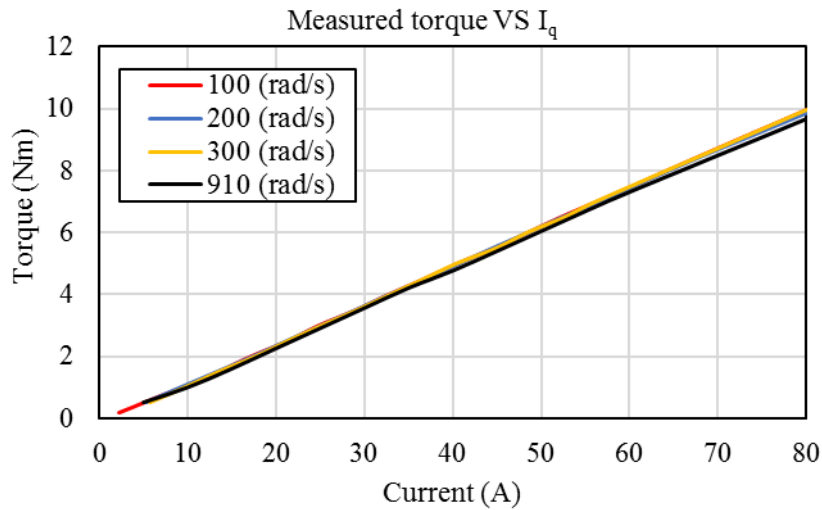


Figure 5-33: Measured torque versus current at different speeds

Figure 5-34 shows the measured and the calculated torque as a function of the I_q current at 910 rad/s (base speed). Based on the experimental results, the calculated torque constant with only q-axis current applied to the machine ($I_q=80$ A, $I_d=0$ A) is 0.123 Nm/A which is 6.9% and 4.37 % lower than the values obtained using 2D-FEA and 3D-FEA simulations, respectively.

The measured torque is in good agreement with the calculated torque using FEA. A discrepancy is however noted as the current increases, leading to a difference between the measured and calculated torque. This is mainly due to the saturation of the stator

laminations which is giving higher flux leakage. Therefore, more accurate results could be achieved using 3D FEA which accounts for the end effects.

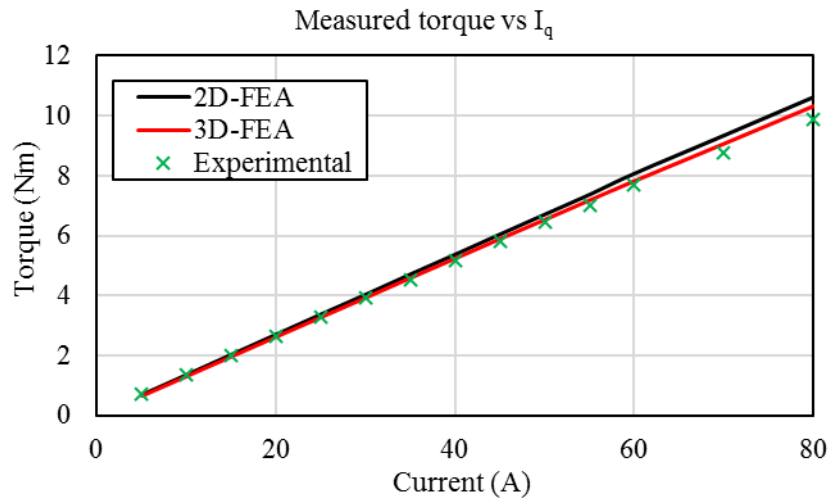


Figure 5-34: The measured and calculated torque versus current

5.4.6 Motor efficiency measurements

The efficiency of the machine has been computed as the ratio between the mechanical output power (P_{out}) measured using a Magtrol torque-meter at thermal steady state, and the electrical input power (P_{in}), measured via the electrical power analyzer as shown in the schematic of the efficiency measurements set-up in Figure 5-35. Tests were done by spinning the motor from 100 rad/s to 910 rad/s (base speed) and changing the load from 1 to 10 Nm.

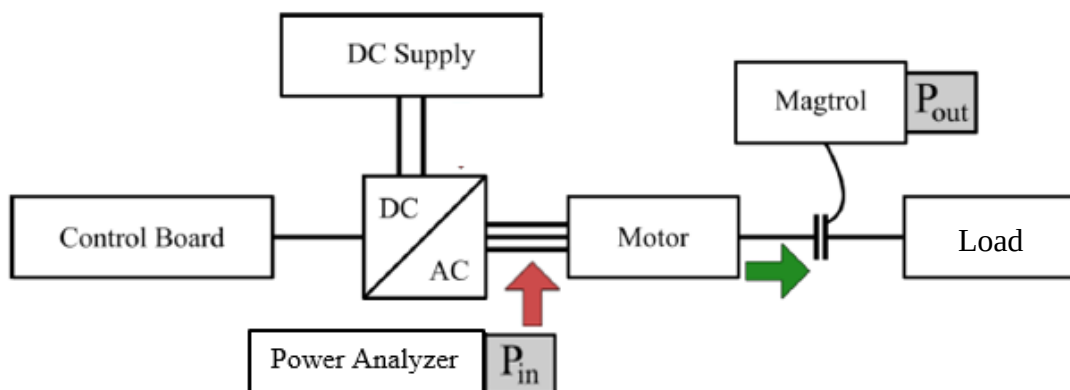


Figure 5-35: Schematic of the efficiency measurements set-up

The measured losses of the prototype machine are shown in Figure 5-36. The measured losses tend to be much higher than the simulated values. This difference in results is due to the additional losses caused by the stator current harmonics which are produced by the Pulse Width Modulation PWM inverter. Also, additional losses consist of the increased mechanical losses (windage and bearings) and the AC copper losses as explained in section 3.4.1 and 3.4.5. In Figure 5-37, the overall efficiency map for different speeds and load conditions is shown. Figure 5-38 shows the comparison of the measured and the calculated efficiency for a rated load of 80 A and up to the based speed of 910 (rad/s). For the low speeds, the losses represent 45% of the input power, and this results in a low efficiency. While at high speeds, the efficiency is increase due to the increase of the output power and then decrease. This reduction in efficiency is mainly due to the increase in the eddy current losses components and due to the increase in the windage losses. Based on the experimental results, the measured efficiency at the based speed and rated load condition was 3.25% lower than the value obtained using 3D-FEA, Figure 5-38. This difference may lie in the evaluation of the losses accurately. Overall, the measured efficiency is in good agreement with the calculated one using FEA, indicating the accuracy of the FE models.

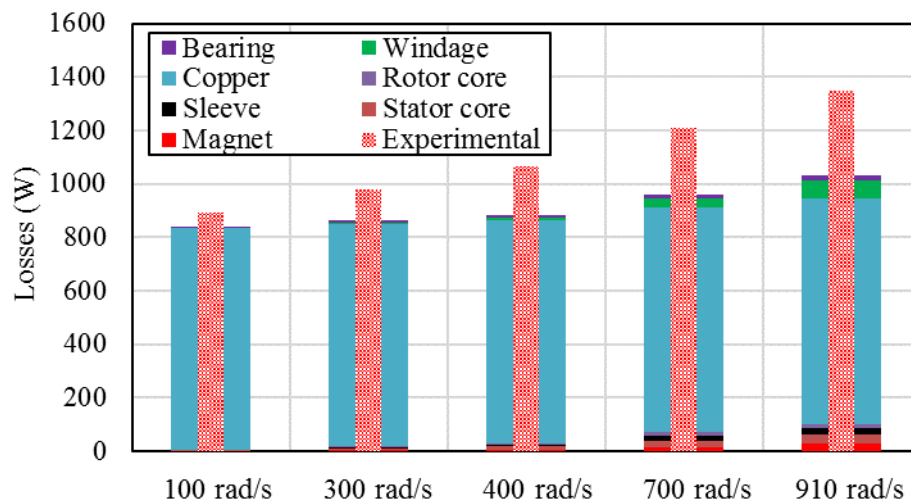


Figure 5-36: Comparison between the measured and calculated losses

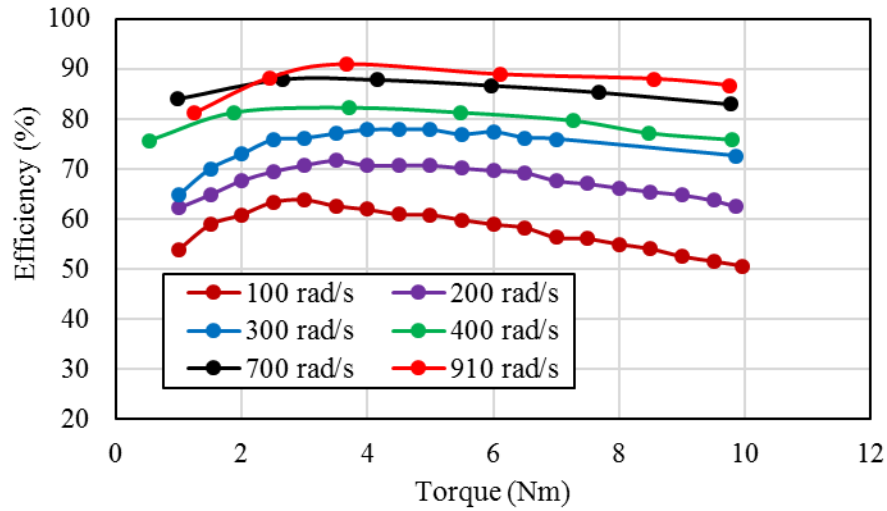


Figure 5-37: Efficiency map of the machine at different speeds and loads

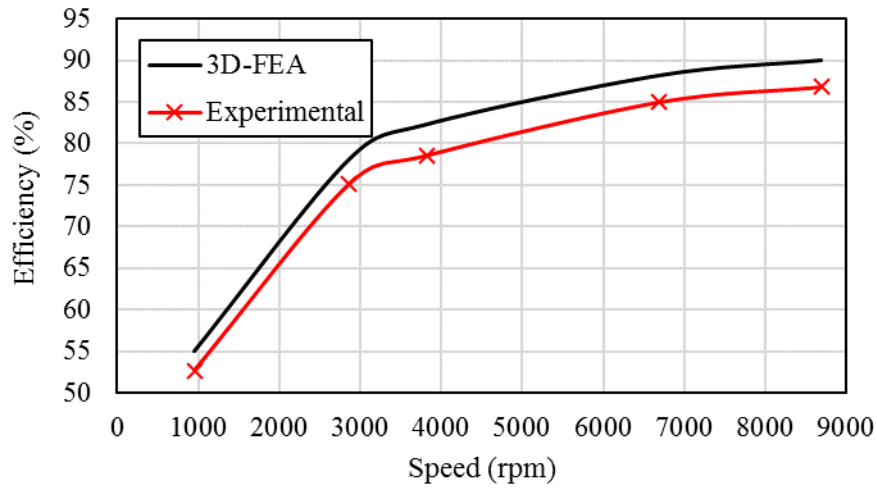


Figure 5-38: Efficiency map of the machine at base speed and maximum load

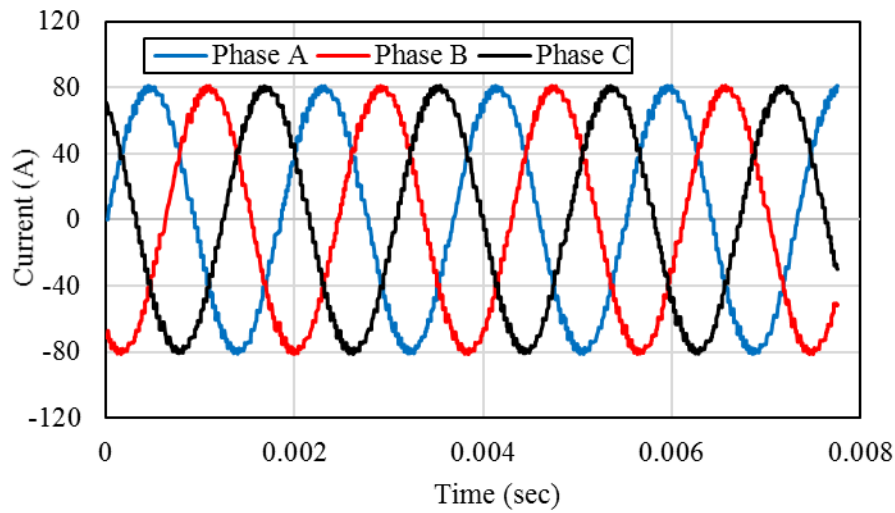
5.4.7 Harmonic losses

As shown in the previous section, another reason for the additional losses in the machine are the stator currents harmonics, which are produced by the PWM inverter. The PWM technique results in higher harmonic components in the output phase current of the inverter, which generates high frequency magnetic fields [129]. These fields induce an eddy current loss in the stator, windings as well as the conductive components of the rotor. This section focuses mainly on calculate the eddy current losses in the PMSM supplied by the PWM. In fact, the losses are estimated by post processing based on the obtained measured data.

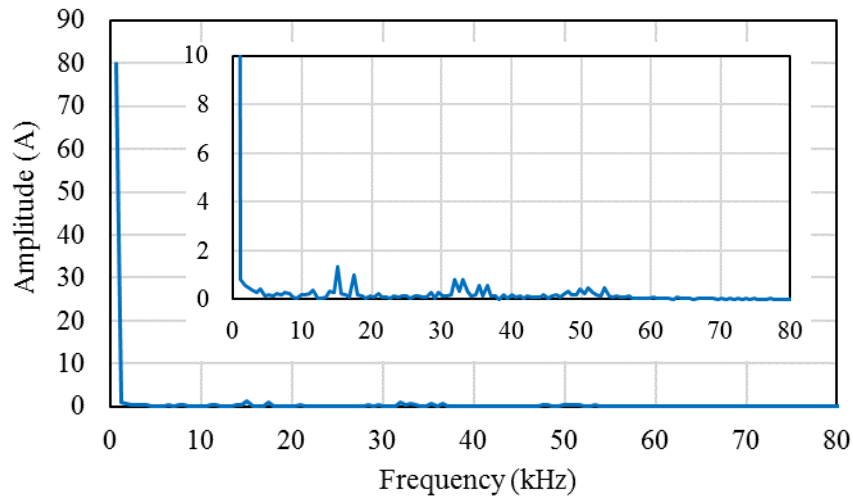
In Figure 5-39, the measured phase current waveforms and its harmonic components are shown for rated load condition at 8700 rpm. Figure 5-40 shows the difference in the eddy currents distributions in the rotor with PWM and sinusoidal supplies. This difference can be considered as the high frequency magnetic field caused by the inverter. Also, the highest time harmonic losses on the rotor side occur very close to the magnet and sleeve surfaces. While, less effect can be seen on the rotor shaft. This is because, skin depth of the magnet regarding to the major field harmonics are less than the magnet thickness, hence the harmonics does not penetrate deep into the rotor shaft.

Figure 5-41 shows a comparison between the measured losses and the losses calculated in the FE model with sinusoidal current and current with PWM harmonics. It can be noted that the additional losses in the rotor shaft remains nearly constant. However, the PM and sleeve loss seems to increase by almost 74.9% and 81.48%, respectively as the PWM. Since the eddy current losses in the stator core for the PMSM is very small (0.1 mm stator lamination). The additional stator core losses caused by the inverter are also small when compared to the rotor additional losses and can be neglected. Among the additional losses shown in Figure 5-41, the ac copper losses rise to 7.5% when compared to sinusoidal supply.

Since the largest percentage of losses arise from the DC copper losses, which makes up 72% of the total losses at the based speed as shown in Figure 5-41. The increase of losses due to the PWM time harmonics in the current was about 1.12 times higher than the sinusoidal fed machine, and this would result in almost a 1.78% difference in efficiency. In spite of including the effect of PWM, the measured losses were 13.8% higher. This is mainly due to the increase in AC copper losses according to the windings arrangement inside the slot (section 5.4.1). While, the increase in iron losses for the stator core caused by the manufacturing process and temperature rise was neglected.

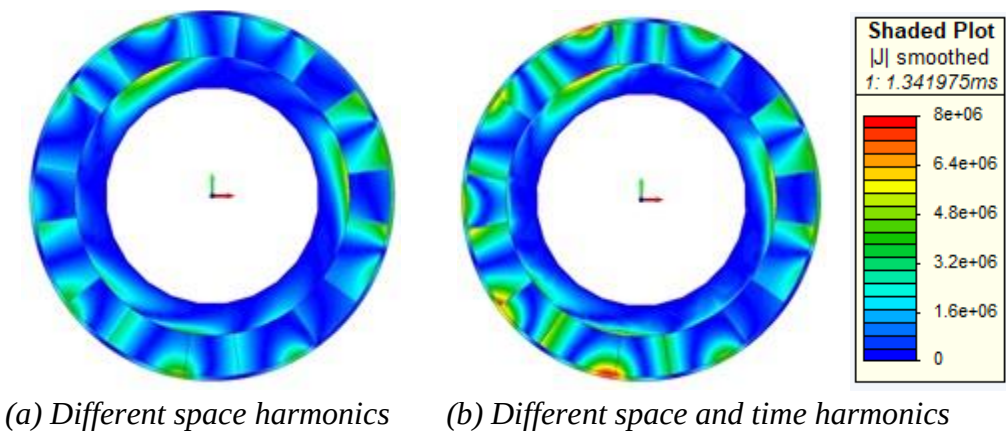


(a) Current waveforms



(b) Harmonic components

Figure 5-39: Phase current waveforms and harmonic content at $F_o=580$ Hz and $F_c=16$ kHz



(a) Different space harmonics (b) Different space and time harmonics

Figure 5-40: Eddy current density distribution in the rotor at 8700 rpm

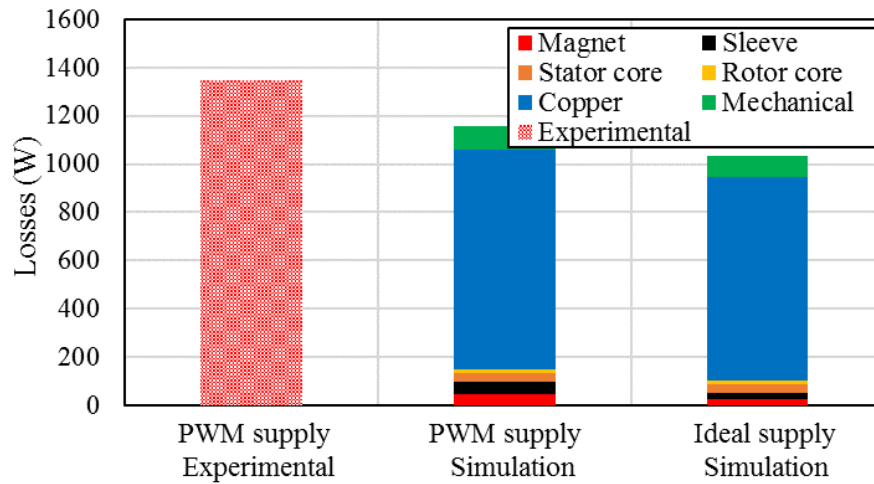


Figure 5-41: Comparison of the losses at $F_o=580$ Hz and $F_c=16$ KHz

5.4.8 Thermal measurements

As shown in Section 5.4.1.c and Figure 5-10, the PMSM is equipped with several thermocouples to monitor the operating temperatures in different parts of the windings (hot spot). Figure 5-42 illustrates the measured temperature rise in the middle of stator winding slot and in the end winding under rated load condition at 8700 rpm. During the experimental tests, the cooling oil is pre-heated at temperature of 90°C before running the machine. With a limited oil flow rate of 4 L/min, the coils between the active winding and the stator core experiences the highest temperature (191°C) due to the absence of direct oil cooling. This test shows that the motor with flooded cooling system could operate under steady state condition without exceeding the temperature limit of 210°C.

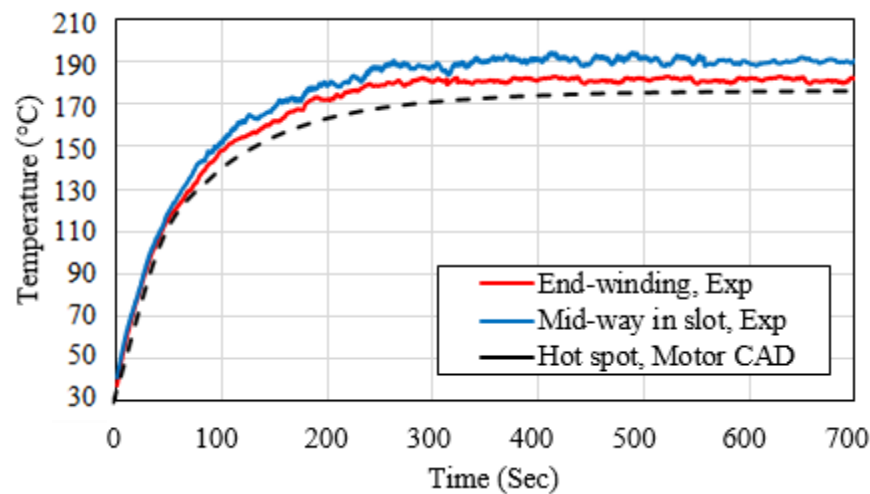


Figure 5-42: Measured temperatures under rated load condition at 8700 rpm

5.5 Conclusion

This chapter has confirmed how accurate the 2D and 3D models that were developed for the case study machine are. The experimental tests presented in this chapter show a clear similarity between the modelling and the prototype results. An important point to mention is how the modelling processes developed here remain faithful to the experimental tests, even in the higher operating points in terms of speed and power.

The experimental results for this chapter were used in two journal publications [124, 130]. The main contributions to research from the work related to this chapter are

- 1) The validation of the relatively new aspects proposed in this thesis, including the cooling channels in the slots
- 2) A robust process how to predict and measure the windage losses for a rotating body.
- 3) Detailed experimental tests in a wide range of speed and load conditions on the prototype with a fully flooded cooling system.

Chapter 6. INFLUENCE OF STATOR CORE MANUFACTURING ON MACHINE PERFORMANCE

It is important to note that, the manufacturing processes on the machine cores have a significant and permanent effect on the magnetic properties of the material, especially on the BH curve and the specific core loss properties of the material [13, 15, 16, 131-137]. This implies that the materials used for the machine no longer reflect the material properties given by their original data sheets. If these variations are not taken into account by the machine designer, then these can have a significant effect in terms of the general performance characteristics of any prototyped machine. In the design and development Chapters above, these effects have been ignored. Therefore, the main aim of this chapter is to investigate these manufacturing effects on the machine developed in Chapter 5.

6.1 Introduction

In the world of electrical machines, a very important design aspect is related to the properties and the characteristics of the machine materials, including those of the copper and the hard and soft magnetic materials. With these properties in hand, electrical machine designers have for decades developed their machines in accordance with their own design and analysis procedures. With recent advances in computational resources, finite element (FE) software packages and the advent of whole suites of new materials, these material properties are today becoming ever more important in the quest towards higher performance, higher speeds and higher efficiency machines.

Traditionally, machine designers perform their development by considering the data provided by the material suppliers. For example, the BH characteristics of the laminated material used for the stator and rotor cores are usually available by either the material supplier or the material distributor datasheets. These characteristics are imported into

the designer's modeling tools with the assumption that the data provided by the manufacturer are accurate and representing exactly the material characteristic for the specific application.

However, it is important to note that such data are usually based on conventional testing and characterization methods. For example, in the case of soft magnetic materials, the data will be achieved from standard tests such as the Epstein frame test and the single sheet test. These conventional characterization methods achieve the data based on tests performed on representative test samples, such as a simple toroid or square-shaped cubes.

However, in reality the machine designer is using those data (obtained from simplistic test samples) to develop a much more complex geometry and shape, such as for example a stator core geometry. For a preliminary sizing exercise this could be sufficient, but when a motor needs to be designed for manufacturing, then the question relative to the validity of the original data arises. It is well known that, the manufacturing of a stator core is a process comprising several steps, e.g. cutting, welding, stamping. Thus, these processes can have a significant effect on the magnetic properties of the material, especially on the BH curve and the specific core loss properties [15, 16]. Which means that the materials being used for the actual machine no longer reflect the material properties given by their original data sheets, and this could lead to a significant effect in terms of the general performance characteristics of any prototyped machine

Usually discrepancies between the expected and the measured results, resulting from these effects are simply waived off and lumped with the "manufacturing error". However, it is also true that these discrepancies and variations do have an effect on the performance characteristics of the machine [138]. Today there is an ever-increasing need for higher performances and cost reduction (e.g. highly demanding application areas such as for aerospace and automotive applications). In such cases the electrical machine is typically already pushed to the technological boundaries represented by the thermal and magnetic limits [116, 139, 140]. Thus, to achieve the required specifications, even the smallest discrepancy and/or variation needs to be taken into account. For this reason, it is necessary to determine the effects on electrical machines'

performance, due to changes and variations in the magnetic properties of the material resulting from manufacturing processes.

6.2 Material and manufacturing effects on machine performance

In literature, the variations in the material properties and core loss due to the manufacturing process on the stator core are investigated in [13, 15, 16, 63, 132-137]. However, there is a very small amount of literature that considers the effect of machining processes on the actual machine performance. In [13], the influence of material degradation on the performance of PM machine, due to cutting, has been studied. In this work, a single sheet lamination, with different number of cuts, has been used for evaluating the magnetic properties which then is used to determine the operating behaviour of the PM machine. In [135], the authors investigate the effect of 6.5% Silicon Steel (SiFe) and CoFe alloy materials on machine performance where the application is quite high speed and therefore material properties play even a bigger role. The magnetic material properties are experimentally characterized by using a conventional single sheet tester in order to obtain the anhysteretic BH curves and the specific core loss. However, the influence of the manufacturing process on the machine performance is not considered. In other works, the performance of a small power induction motors have been investigated using a suite of different magnetic materials, such as M600-50A, NO20, NO12 and amorphous steel alloy 2605sA1, for the stator core [136, 137]. The anhysteretic BH curves and the specific core loss have been obtained using a toroidal sample and the induction motor performance have been evaluated in the speed range from 0 to 3000 rpm at full load condition. The effect of welding processes is investigated in [16], with focus on the influence of welding on the material losses. Experimental results using a ring core at different frequencies and flux densities are presented. The results show an increase of the core loss by 10% due to the welding process, while the effect on the machine efficiency is about 0.11%. The welding effect on small slot-less PMSM performance has been also studied in [15]. Stator cores made of SiFe and nickel iron steel sheets are experimentally characterised to investigate the changes in magnetic material properties. A main outcome of the study

is that negligible influence on both the induced back-EMF and output torque was found, while an increase of core loss from 32% to 44% was observed. This is due to the deterioration of the material properties in the area of the welding seams. Since, the welding seams to partly destroy the sheet insulation layer.

All the above indicates that while a number of previous works do acknowledge and study the effects that manufacturing processes have on the material itself, only a couple of cases are actually translating the analysis into the domain of the machine performance.

Thus, the major objective of this chapter is to accurately map the machine's efficiency ratings at different operating points, while of course taking into account the influence of manufacturing process on the magnetic properties of the hard magnetic material.

In order to do this, three main data sets (DS) of material properties are used, namely

DS1 - The data provided by the material supplier for the laminated sheet steel (as provided in the data sheet).

DS2 - The data of the same material characterized in house, using standard, conventional test samples.

DS3 - The data of the same material characterized in house, using test samples, composed of completely manufactured stator core geometries (i.e. that include all the manufacturing processes).

For DS2 and DS3 above, the data is captured on a purposely-built, magnetic material characterization set-up as introduced later in Section 6.3.2.

6.3 Characterisation of magnetic material properties

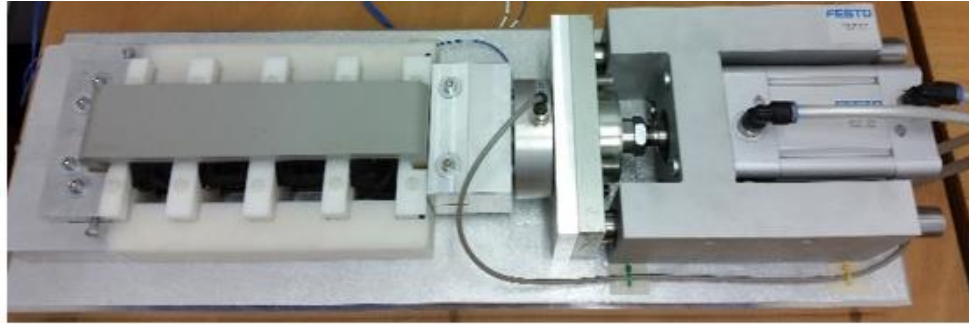
6.3.1 Description of the test lamination sheet and the manufactured stator core

Different methods can be employed to investigate the magnetic properties and the core loss for the laminated material [141, 142]. As mentioned above, the most commonly used methods for measuring the magnetic properties of a laminated sheet are the Epstein frame test and the single sheet test. These testing methods are usually preferred due to their accuracy and simplicity [141, 142]. However, these methodologies are usually performed on representative test samples, such as a simple toroid or square-shaped cubes, where any effects of machining are only limited to the outer perimeter of the sample.

When the actual stator core shape and geometry is considered, the degradation effect of machining is higher as now the length of the “edges” of the sample that have seen machining is much longer. This can be appreciated by a visual comparison of Figure 6-1 and Figure 6-2, where it can be observed that the square sample of Figure 6-1 is much less impacted by machining than the actual core geometry of Figure 6-2. All this indicates that for a standard test sample, the measured BH curve and the specific core loss do not accurately represent the properties of the actual manufactured stator core of the machine. This indicates that much more relevant and accurate material characteristics could be achieved from the actual machine stator core itself [141]. By using the actual stator core for material characterization, the effects of manufacturing processes, such as punching or cutting, stacking, welding and annealing, which might have a significant influence on the magnetic properties of the material [143], can be taken into account and fed back into the machine design process.



(a) Sheet sample



(b) Single sheet tester

Figure 6-1: Prepared material sample for testing.

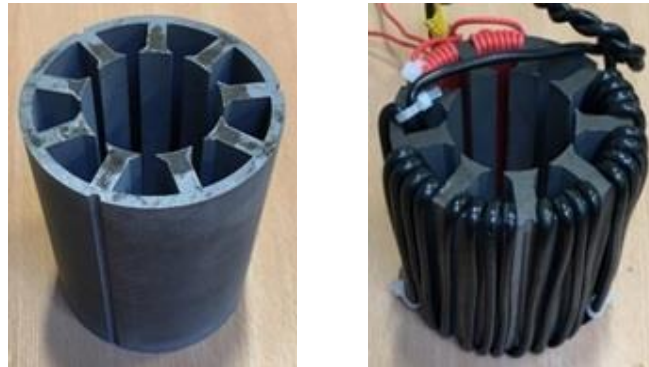


Figure 6-2: CoFe material stator core (left) and wound core under test (right)

As explained in Section 6.2, three different sets of material data are considered in this work. The data produced in-house (DS2 and DS3 in Section I) is achieved by using two different methods to evaluate the magnetic material properties. For DS2, i.e. data achieved on simplistic test samples, a single sheet of the lamination material is used. Figure 6-1 (a) shows the sample of lamination sheet used for the test, while testing setup for this is shown in Figure 6-1 (b). For DS3, i.e. with data achieved from a completely manufactured stator core, the test sample is shown in Figure 6-2.

Both the stator core and lamination sheet used for the experiment are made from a CoFe alloy with 0.1mm lamination thickness. The CoFe sheets are made of approximately 49% cobalt, 49% iron and 2% Vanadium [100, 144]. Table 6-1 gives an overview of the CoFe material properties that were obtained from the manufacturer's datasheet.

Table 6-1: Material properties of the investigated CoFe material

Lamination Thickness	Density	Electrical Resistivity	Package Density	Saturation Flux Density
0.1 mm	8120 kg/m^3	$0.4 \mu\text{ohm}$	98%	2.3 T

6.3.2 Description of the measuring setup

In order to accurately characterise these samples, a state of the art testing facility that is available in-house was used to characterize these samples. Figure 6-3 shows the main interface of the magnetic measurement system (MPG 200 Brockhaus) used to perform the experimental tests on the stator core and lamination sheet. The measuring system is equipped with a high-power amplifier, a digital frequency generator and a set of safety protections in order to ensure high accuracy and high fidelity of the tests along with high level of safety while testing. A high speed real-time software interface allows easy programming and monitoring of the experiments. Indeed, a wide range of excitation waveforms (e.g. DC bias offset, Sin Wave, Pulse Width Modulation and Higher Harmonics), can be applied to the samples under test. Table 6-2 presents the specifications of the measuring system which is presented more in detail in [47, 135, 145]. The core loss tests and BH magnetization curves at high frequency and high flux density for the slotted stator core and for the lamination sheet are performed under pure sinusoidal wave excitation.

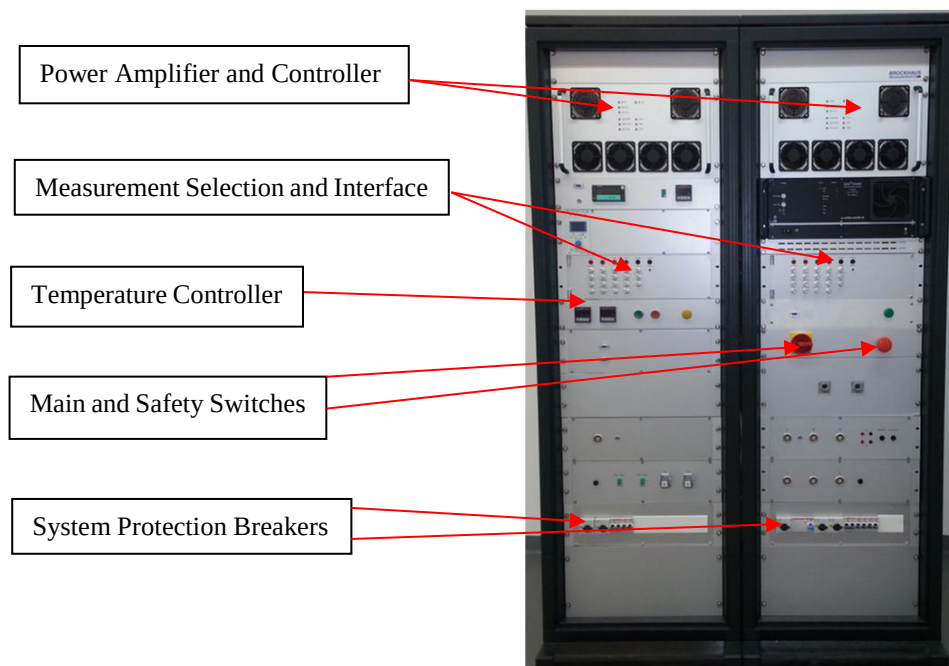


Figure 6-3: Magnetic measurements system, MPG 200 Brockhaus setup bench

Table 6-2: Specification and capabilities for the measurement system

Parameter		Quantity
Fundamental Frequency		2 – 100 kHz
Low Frequency Amplifier Capacity		100V 52 A (Peak)
High Frequency Amplifier Capacity		50V 10A (Peak)
Operating Temperatures	Soft Magnetic Materials	-40°C to 180°C
	Hard Magnetic Materials	-40°C to 300°C
Excitation Waveforms		Sine Wave
		DC / Normalized
		PWM
		Harmonic

6.3.3 Investigated samples

For the single sheet test, the investigated sample is a standard geometry with dimensions of 400 mm $\times$ 30 mm $\times$ 0.1 mm and magnetically characterized according to the specifications of the IEC 60404 standard. The lamination sheet is placed within two coils (primary and secondary) as shown in Figure 6-1 (b) and the magnetic circuit is completed by two laminated C cores with ideally zero losses and zero MMF drop. The power core loss is then measured by subtracting the resistive drop from the input power.

For the stator core sample, the core is setup with two set of coils as shown in Figure 6-2. The black coil is used as the primary circuit (excitation) and the red coil is used as the secondary circuit (measurement). The primary and the secondary number of turns are selected to be 21 and 10, respectively.

6.3.4 Magnetic measurements

For the magnetic material properties investigation, the slotted stator core and the lamination sheet are exposed to a defined magnetic field, in which the required primary excitation current is supplied by the high power amplifier and the maximum value of the magnetic field strength can be computed using (6-1), where N_1 is the number of

turns of the primary coil, i_1 is the primary current and L_{core} is the mean length of the magnetic flux path.

$$H_{(t)} = \frac{N_1 \cdot i_1}{L_{core}} \quad 6-1$$

The induced voltage in the secondary coil is continuously recorded to determine the magnetic flux density as given by (6-2), where, N_2 is the number of turns of the secondary coil and A_{core} is the cross-sectional area of the sample.

$$B_{(t)} = \frac{1}{N_2 A_{core}} \int v_2 dt \quad 6-2$$

For the stator core set-up, it is assumed that the flux leakage into the tooth area near to the back iron is negligible. In other words, it is assumed that all the flux crossing the core back iron is equal to the flux linkage to the secondary winding. This assumption is verified in Section 6.4.3.

According to the assumption of negligible leakage flux the L_{core} core can be determined by $L_{core} = \pi * (R_{OD} + R_{ID})$, where R_{OD} and R_{ID} are the outer and inner radius of the stator yoke. The cross-sectional area A_{core} of the samples can be determined by (6-3), where M the total is mass of the stator yoke and ρ_{co} is the mass density of the material.

$$A_{core} = \frac{M}{\rho_{co} * L_{core}} \quad 6-3$$

Finally, to determine the instantaneous value of the core loss, the current in the primary coil and the voltage of the secondary coil are measured simultaneously and multiplied by the coil ratio as given by (6-4),

$$P_{core} = \frac{N_1}{N_2} v_2 i_1 \quad 6-4$$

The core loss density (W/kg) can be calculated using (6-5), where S_l is the stator core length.

$$P_{specific} = \frac{P_{core}}{\rho_{co} \cdot S_l \cdot \pi \cdot (R_{OD}^2 - R_{ID}^2)} \quad 6-5$$

6.4 Measurement results

By post processing of the measured data according to the procedures explained in Section 6.3 and using the geometrical specifications of the test samples, the required material properties, such as the BH magnetization curve, the DC relative permeability, the AC BH magnetizing curves and specific core loss can be determined [135].

6.4.1 Basic magnetic properties

First of all, the BH magnetization curve for both the slotted stator core (DS3) and the lamination sheet (DS2) are calculated. The obtained results are compared to the BH magnetization curve provided by the manufacturer (DS1), as shown in Figure 6-4. Analysing the results of Figure 6-4, the BH curve as derived with data provided by the manufacturer shows higher relative permeability, which results in higher magnetic flux density.

The DC relative permeability for all the three different data scenarios is calculated and the results are shown in Figure 6-5. It is clear that the relative permeability of the CoFe material is higher in the range of magnetic field strength 40 A/m to 100 A/m, while the relative permeability goes lower as the core material tends to saturate with higher magnetic field strength. Nevertheless, in the range from 1.5T to 2T, the relative permeability of the stator core is decreased by 67.5% and 47.6% compared to the slotted stator core data and lamination data respectively. However, the mismatch among the relative permeability values is negligible for magnetic field strength higher than 600 A/m.

This indicates that the process using the data derived from the DS3 process results in a 50.3% difference in terms of stator flux density when compared to the other methodologies and thus already highlights the significant change that the manufacturing processes can inflict upon the stator material.

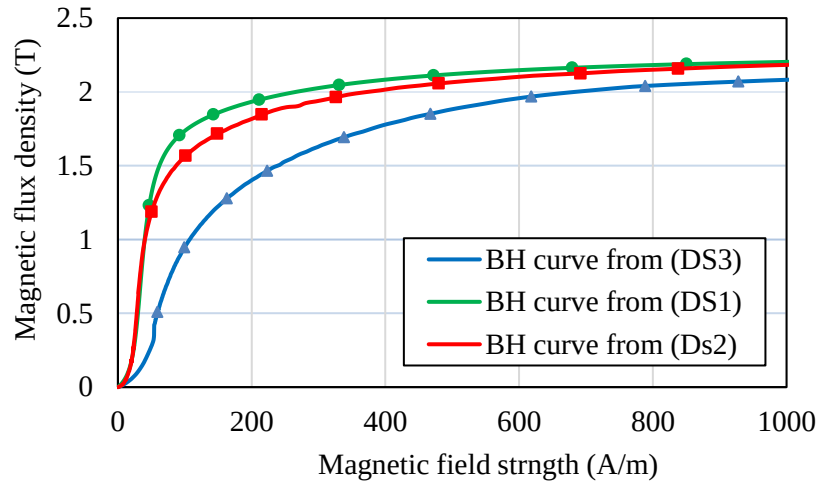


Figure 6-4: Comparison among BH magnetization curves

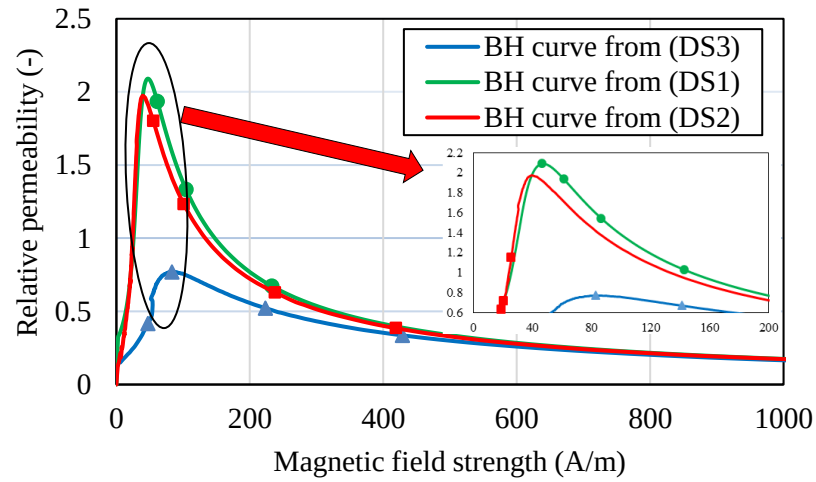


Figure 6-5: Comparison among DC relative permeability

6.4.2 Core losses

As the shape and size of the BH curve (hysteresis loop) depends on the frequency and the field intensity [146]. For the same flux density, the area of the AC hysteresis loop which represents the energy loss per cycle is higher with an increase of frequency and therefore, higher core loss per unit volume. To investigate the variation in the hysteresis loop, and hence variation in specific core loss, another AC test is carried out on the stator core and the lamination sheet with a flux density peak value of 2 T and frequency of 1000 Hz. The results are shown in Figure 6-6. The increase of the hysteresis loop area is more visible in the case of the slotted stator core. In ideal condition, the BH

curve should be as that shown in (Label A Figure 6-6). However, this only counts for the laminated sample which does not go through any manufacturing process (only machining process). The actual stator core was manufactured by gluing the laminations together under pressure and then machining using electrical discharge machining (EDM). The surface is heat treated while machining and the grain size is affected by stacking pressure.

Thus, the manufacturing process affects the performance of material (Label B Figure 6-6), bringing the permeability down for the same magnetic field strength [15]. Also, there is a negligible effect of machining on laminated sheet as the strip is significantly wider 30 mm compared to the stator back iron which is 3.5 mm [15]. After investigating the hysteresis loops, the specific core losses were also determined in order to quantify the influence of machining processes on them.

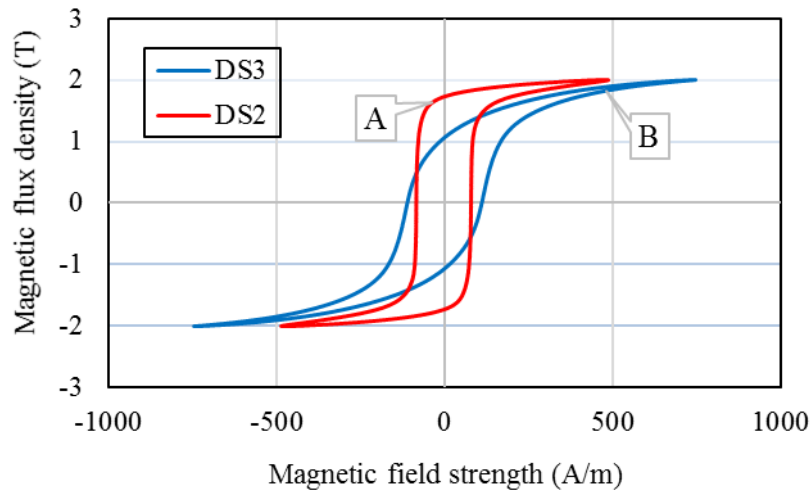


Figure 6-6: AC hysteresis loop at 2 T and 1000 Hz

The specific core loss as a function of different frequencies and flux densities are shown in Figure 6-7. It is found that the core losses are increased in the range of flux density from 1 to 2 T, and the increase in the core loss becomes higher for higher frequencies. The measurements were carried out on both the sheet lamination and slotted stator core samples. The measurements on the stator core showed an increase in the core losses and change in the anhysteretic BH curve and the relative permeability due to the manufacturing process. Even if the flux density is lower for stator core DS3 compared to DS1 and DS2 as shown in Figure 6-4. The losses are higher for DS3 compared to

DS1 and DS2 as shown in Figure 6-7. This is due to fact that, the area of the hysteresis curves for the manufactured stator core DS3 have changed and become bigger because of the manufacturing issue as shown in Figure 6-6 and this reflected in the iron losses measurements. Same thing happens with the eddy current losses as well, because the manufacturing process affects the material properties.

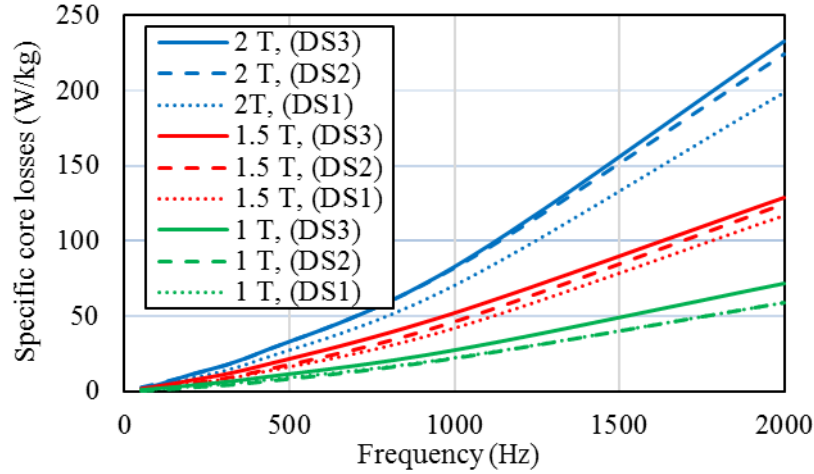


Figure 6-7: Specific core loss comparisons utilizing the three DSs

6.4.3 Validating the negligible leakage assumption

In order to verify the assumptions made at the beginning of the analysis (negligible fringing flux inside the teeth), FE simulations on a slotted stator core, with the same geometrical dimensions and same number of turns as used in the experimental tests conducted on the manufactured stator, are carried out over a range of flux density levels. The results highlight a reduction in the flux density in the back iron of the slotted stator core due to the fringing flux into the teeth. Figure 6-8 shows the magnetic flux lines in the stator core for a maximum current of 41A, where the flux density at the top of the teeth is reduced due to the flux fringing. By following the procedure presented in [14, 63], for different excitation currents, a correction factor has been introduced. The corrected flux density in the stator core is then calculated considering the fringing factor and the results are shown in Figure 6-9. As can be observed, the effect of the fringing flux on the BH curve is smaller than 5%. However, this has been considered in the computation of the motor performance.

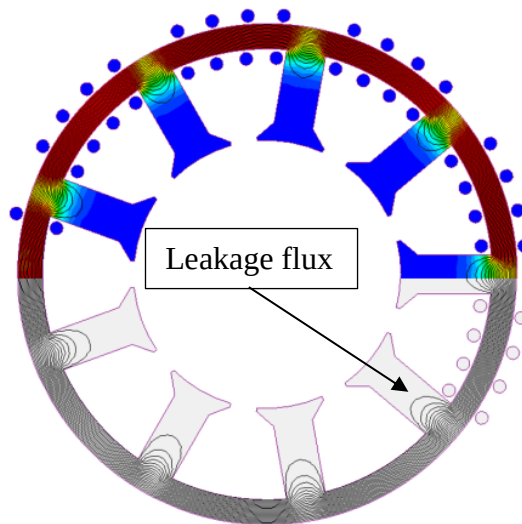


Figure 6-8: Field flux lines in the stator core

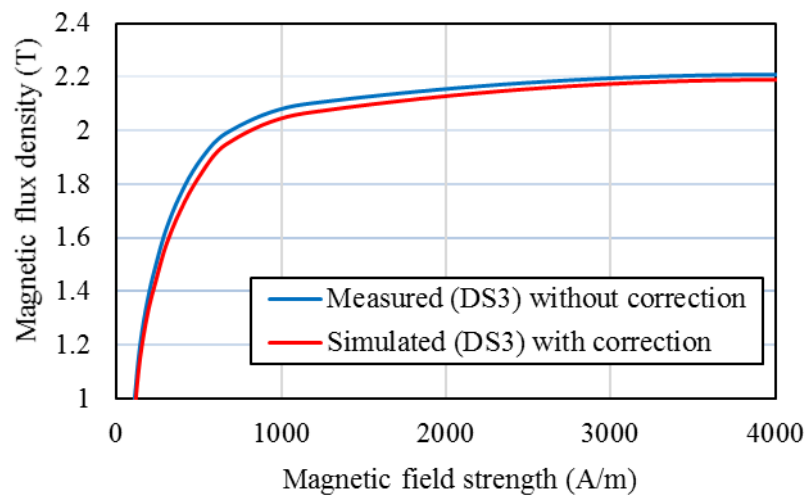


Figure 6-9: BH magnetization curves for the measured stator core

6.5 The influence of material properties on machine efficiency performance

Having identified and quantified the differences in the material properties that result from machining processes, it is then possible to use this information to assess the overall influence on the machine performance. In this section, the main consideration is done on the efficiency maps of the machine, which for this particular (relatively high speed) application are highly dependent on the specific core losses of the laminated material. Traditionally, core loss in electrical machines is always severely underestimated [138]. By combining numerical simulations with the measured losses as explained above then more accurate representations of this important factor can be achieved.

Also, this process permits the user to quantify the effect of the manufacturing process on the stator core loss increase for different machine operating conditions. To assess these effects, the machine is simulated with the three DSs of material data and the results compared for different operating conditions. Under no load condition, the maximum peak value of the back EMF at different operating speeds of 4000, 8700, 13000 and 19000 rpm are reported in Table 6-3. It can be seen that for different operating speeds the change in the no load voltage is negligible (<1%). This is due to the fact that the magnetic flux density is mainly imposed by the magnets and maintained below the saturation level.

Table 6-3: Comparison of the Back-EMF for different data sheets

		DS1	DS2	DS3
back-EMF V (peak)	4000 rpm	36.29	36.29	36.21
	8700 rpm	78.91	78.9	78.73
	13000 rpm	117.89	117.88	117.62
	19000 rpm	174.41	174.38	173.98

In Figure 6-10 (a) the core loss in the stator core at no load condition and for different operating speeds are presented. The results show that at low operating speed of 4000 rpm, the core loss using the measured data of the stator core is 1.98 W higher than the data provided by the manufacturer. As expected, the difference in the core loss increases towards the higher operating speeds of 19000 rpm with 16.28% difference between the measured data of stator core and data sheet of the manufacturer and 8% between the measured data of stator core and lamination.

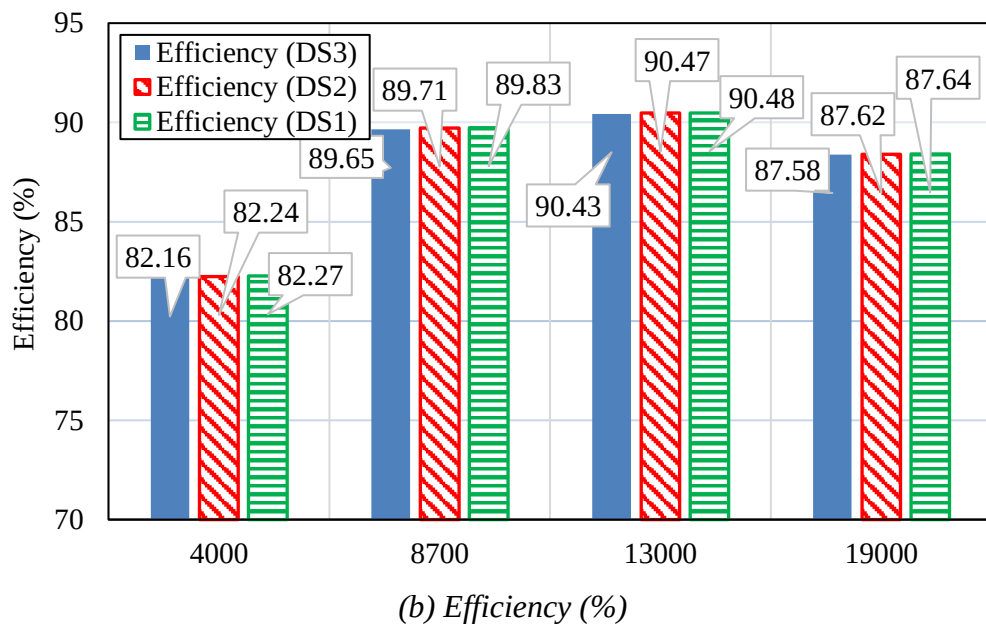
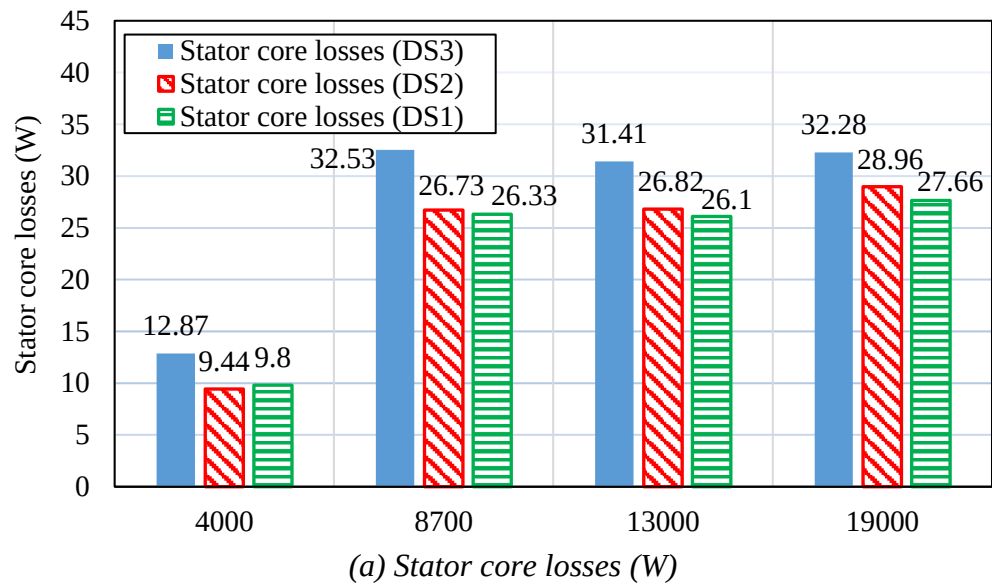


Figure 6-10: Stator core losses and efficiency at rated torque and different speeds

To complete the machine characterization analysis, the performance under load conditions is also determined. By using the data from DS1, DS2 and DS3, the torque-speed curve is calculated for different operating speeds at rated torque condition. The results of machine performance at different operating points are presented in Figure 6-10. In accordance with the simulation results presented in Figure 6-10, a variation in the core loss and efficiency can be observed between the measured data and data sheet of the manufacturer. This discrepancy is due to the deterioration of the magnetic material properties and the increased core loss as shown in Figure 6-4 and Figure 6-5.

In order to evaluate the effect of machining and manufacturing processes of the lamination sheet and stator core on the machine performance, different operational maps for the PMSM are described in detail. Figure 6-11 shows the efficiency map calculated based on DS3 data, DS2 data and DS1 data respectively. The efficiency maps calculations are carried out over the entire speed and torque range and are including all the losses components (copper, stator core, permanent magnet, sleeve, rotor shaft and windage). It can be seen that the variation in efficiency increases towards lower operating points. Maximum reduction in efficiency is about 1% in Region A. This variation in efficiency is mainly caused by increased core loss and decrease in output torque.

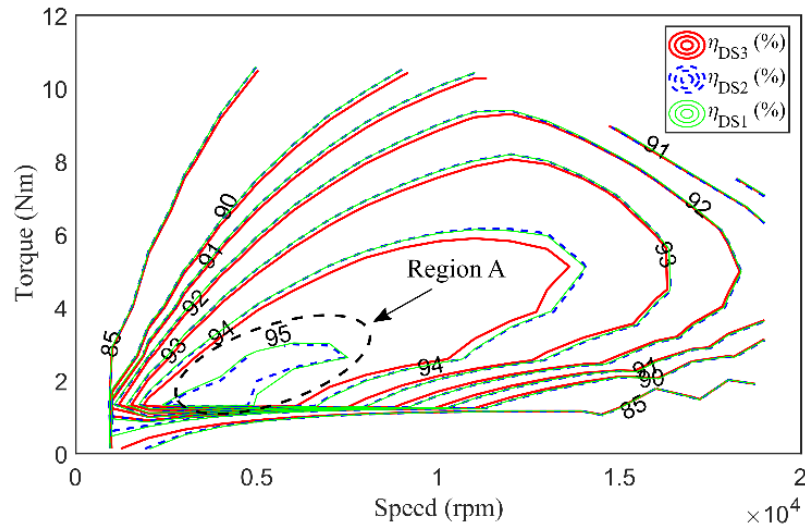


Figure 6-11: Efficiency map of the PMSM

Figure 6-12 and Figure 6-13 show the influence of the frequency and flux density on stator core loss and output torque, respectively. Figure 6-12 shows the core loss in the operational map where C_{DS3} , C_{DS2} and C_{DS1} are the core loss calculated as based on DS3, DS2 and DS1 respectively. It can be seen that results coming from FE simulation when the DS3 data is used, have the highest core loss because of the manufacturing processes on the stator core. At low rotational speeds and with different operating points, the variation in core loss is negligible (around 1 W) and its difference can increase up to 23% (rated operating point). These are caused by increased flux density and operating frequency at these operating points. While, within the area of the flux weakening between (9000 – 19000) rpm, the core loss increases until a certain operating point and then start to decrease due to the reduction of the flux density in the teeth and

back iron. Also, it is found that the difference in core loss between C_{DS2} and C_{DS1} is not significant due to the small variation in the specific core loss obtained from the tested laminated sheet and manufacturer's datasheet values. This is true because no manufacturing processes were applied to achieve the DS2 data.

Figure 6-13 shows the percentage of reduction in the output torque. This is given as $(T_{DS3} - T_{DS2}) / T_{DS3}$ and as $(T_{DS3} - T_{DS1}) / T_{DS3}$, where T_{DS3} , T_{DS2} and T_{DS1} are the output torque calculated based on DS3, DS2 and DS1 data, respectively. The influence of the manufacturing process on the slotted stator core is observed to be most dominant at lower magnetic field strength and tend to decrease towards the saturation region as shown in Figure 6-4.

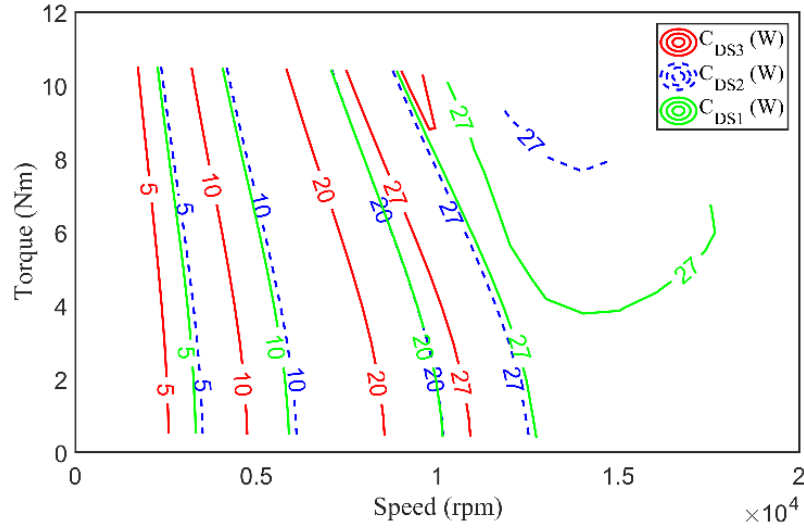


Figure 6-12: Stator core losses map of the PMSM

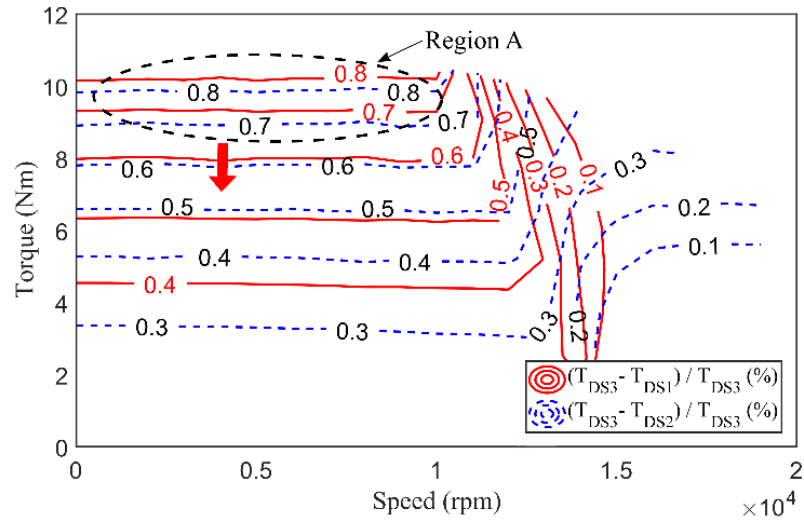


Figure 6-13: Torque reduction map of the PMSM

This results in flux density reduction at lower magnetic field strength, leading to a lower output torque of the designed machine. The reduction in the output torque is noticeable in (Region A in Figure 6-13) with high torque and low rotational speed. A maximum reduction of 0.8% in output torque was observed when experimental data from slotted stator core is used compared to manufacturers datasheet and lamination sheet values (Region A in Figure 6-13). This reduction was caused due to the variation in the flux density (i.e. variation in anhysteretic BH curve) due to the different DSs. While, this variation in output torque was dropped to 0.3% at lower operating points. In particular, for the flux weakening region, it is important to note that the difference in the output torque decreases with the operating speed gradually increasing. Since, higher operating speeds required more negative d-axis current and therefore further reduction in the flux density and output torque.

6.6 Highlights of the chapter and contribution

The work related to this chapter has resulted in a journal publication [147] and a conference paper [49]. The key observations from this work can be summarized as:

- 1) There are negligible variations on the machine performance when a design based on the properties of DS1 is compared to a design based on DS2. This is as expected as in reality the processed used for DS1 and DS2 are basically the same with the only difference being that the DS2 data was produced in house.
- 2) There are significant variations in the machine performance when the machine design is performed using the data in DS3, i.e. when the material data does consider the manufacturing processes required to produce a stator stack. These can be categorized as follows
 - a) A drastic increase in specific core loss.
 - ii. In general, the core losses increase drastically across the full operating range which results in lower efficiencies of the designed machine when considering data from DS3.

- iii. In particular, a 23.5% increase in the stator core loss was observed at the rated operating point of 8700rpm, relative to when the information from DS1 and DS2 is used.
- iv. A 16.7% increase in stator core loss was observed at the maximum operational speed of 19000rpm.

The above is mainly due to the lower flux density in both the teeth and the back iron (flux weakening region).

- b) A significant reduction in magnetic field strength.
 - i. It's observed that at lower magnetic field strength the reduction in the flux density was significant and this variation tend to decrease towards the saturation region. This reduction in the magnetic flux density results in a lower output torque of the designed machine.
 - ii. For such a machine, the reduction in the output torque at full load condition was 0.8%. While, small effect was noted beyond the base speed due to the increase in the negative d-axis current and therefore small variation in the flux density.
- c) Reduced efficiency performances
 - i. An overall reduction in machine efficiency of 1% was observed when the data from DS3 was considered and compared to the designs that utilized data from DS1 and DS2. This variation is related to the reduction in the output torque and the increased core loss.
 - ii. For this particular application, where the highest speed is 19000rpm, a 1% efficiency reduction is not considered to be a major issue. However, it is important to note that for higher speed applications, this can and will have a much more negative impact.

6.7 Conclusion

In this chapter, the effects that variations in material properties (due to manufacturing processes) can have on the general performance characteristics of electrical machines have been investigated and considered. In order to do this, an accurate characterization process that considers three different DSs was used in order to be able to compare how

the overall machine performance is affected according to which DS, the machine designer chooses to use during his development and design of the machine. All existing sources in literature mainly limit themselves to highlighting the differences in the materials before and after specific, ‘standard’ machining processes. This chapter however takes the analysis the full way and considers and compares the full machine performance in terms not only of material properties but also in terms of operational performance. The comparisons are done between the cases of 1) when ‘standard’ material properties are considered and 2) when ‘modified due to machining’ material properties are considered.

In this work, special attention was focused on the efficiency maps of the machine as this aspect is highly dependent on the material properties that are the most sensitive to manufacturing processes such as the BH curve and the specific core losses

This work has thus indicated and confirmed that the often-overlooked effects of manufacturing processes can have a significant and important effect on the overall machine performance. While in the particular case considered here the described effects were not of particularly high values, however it is clear that in the quest for higher performance and higher speed applications, then the above aspects need to be given careful attention.

As a final remark, it is worth noting that for the PMSM of this case study, the main resulting consideration was the 1% reduction in the general efficiency. For a high performance aerospace application, a 1% efficiency reduction cannot be neglected and in fact for the particular application at hand, the problem was solved by using ultra-thin laminations at the cost of more expensive material and less mechanical strength.

While more accurate results were achieved in Chapter 6 by considering the effect of manufacturing processes on the stator core. However, there is a difference between the measured torque value (shown in Chapter 5) and the predicted value (presented in Chapter 6). Therefore, in order to improve the machine performance estimation and to identify the reduction in the output torque. The rotor endcaps will be taken into account in the following chapter so as to achieve more reliable and accurate modelling results.

Chapter 7. INFLUENCE OF ROTOR ENDCAPS ON MACHINE PERFORMANCE

This chapter deals with the influence of the leakage flux on the machine performance that has been developed in Chapter 4. Mainly the work focuses on an area which is not commonly investigated at all, namely understanding the effects of leakage fluxes in inactive components of the machine such as in the permeable rotor endcaps. Generally, the magnetic characteristics of the rotor endcaps affects the distribution of end leakage magnetic field, and thus could have a significant impact on machine performance.

Therefore, it is necessary to investigate the influence of the permeable rotor endcaps on the electromagnetic field and their eddy current losses to improve the accuracy of the FE models for an improved design. Since the original FE model developed in Chapter 4 does not have rotor endcaps, an updated 3D model with endcaps has been proposed in this chapter.

For the proposed model, the electromagnetic performance characteristics, such as the open circuit results, electromagnetic torque and torque-speed characteristics, as well as the electromagnetic losses and efficiency are investigated and compared to the original models of Chapter 4.

7.1 Introduction

Various works machines [148-156] describe the influence of the end-effect on the electromagnetic performance of electrical machines. In [152, 155], a lumped parameter circuit was used for modelling the flux leakage paths in the end-region of a flux switching permanent magnet (FSPM) motor. The above method can predict results relatively quickly, but with some limitation in terms of accuracy due to the assumptions on which they are built. The influence of end-effect on the parameters and torque-speed

characteristics of FSPM machines has also been analysed using 2D-FEA and 3D-FEA, and validated by experiments in [153, 154, 156]. The 3D-FEA and 2D-FEA results of a doubly salient machine are compared in order to examine the end-effect influence in [150, 151]. In this work, an end leakage factor for the end-effect has been defined to improve the accuracy of the 2D model. In [17], the 3D end effects in the flux mapping of synchronous reluctance machine was investigated. The results of the research showed that the inaccuracies on the estimation of the flux mapping, and then on the calculated maximum torque per ampere strategy can be significantly reduced by using 3D FE models. Also, the influence of end-effects on the performance prediction of PMSMs with different non-overlap windings have been analysed and investigated in [18]. This work has shown that despite the short end windings of PMSM, these end-effects have a significant influence on the performance parameters of the machine. In [157], an end-effect factor has been proposed for the back-EMF of PMSM to consider the leakage flux in the end region of the stator and airgap. The proposed equivalent 2D FE method was a very precise with small error when compared to the measurements and 3D FE method.

All the above works focus a lot of effort on the end effect of the electrical machines. However, electrical machines also comprise other permeable parts which increase the leakage flux and losses [18] but which are usually ignored, especially if these components are considered to be inactive material, i.e. not part of the main torque-producing magnetic circuit. In [158], a PMSM with metallic and non-metallic rotor clamping rings was investigated. The study focused on the eddy current losses at the maximum speed. In [159], the additional losses in the stator end regions of PM wind generator are analyzed using a 3D FE model. The results show that failure to consider the leakage fluxes in the inactive material such as the supporting arrangement of the stator and the clamping rings can result in additional losses and therefore affect the overall efficiency.

Considering all the above, it was concluded that while for years and decades, the effects of fluxes in the inactive material have always been ignored, however with today's push towards maximizing performance and efficiency, there is scope to consider and investigate these effects.

In order to do this, three FE models, labelled as M1, M2 and M3 are defined, where M1 is the original 2D-FE model developed in Chapter 4.

M2 is the 3D version of M1 but without including the rotor end-caps.

M3 is the same as M2 but includes the rotor end-caps.

Figure 7-1 shows the models used in the analysis. Endcaps are shown on both ends of the machine rotor shaft, as depicted in Figure 7-1 (c).

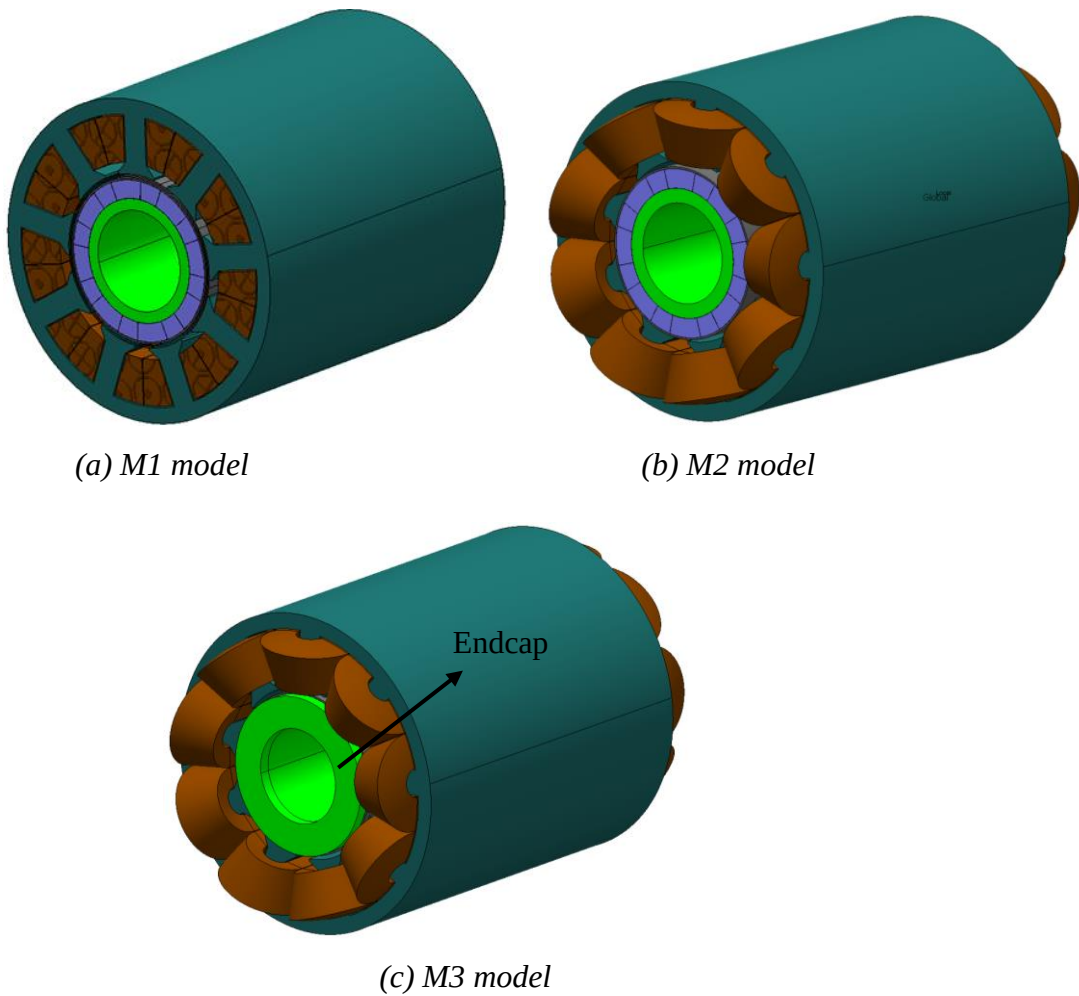


Figure 7-1: Models of the PMSM

7.2 Design consideration

For high-speed applications, a retaining sleeve and rotor endcaps are common measures, employed to guarantee the rotor mechanical integrity. The retaining sleeve however does lead to a larger magnetic airgap, while the presence of permeable rotor endcaps increases the leakage flux at the front and back ends of the rotor [157].

Sleeves are normally applied with a large interference fit to achieve the required retention. However, when the rotor is in standstill or at low speeds, the high interference fitted sleeve results in a high compression, which axially stresses the PMs. To overcome this effect, assemblies with endcaps are used to eliminate any axial strains due to the interference fitted sleeve [120].

For high-speed, low weight PMSMs a hollow shaft is often proposed to reduce the overall machine weight and rotor inertia. As the machine being considered in this work is for an aerospace application, weight, inertia and efficiency are very important aspects. To achieve these requirements, a rotor with a permeable hollow shaft was proposed in Chapter 3. The permeable material used for the shaft performs the same magnetic function of the rotor back-iron, ensures the appropriate mechanical strength and also enhances the magnetic flux production of the quasi-Halbach permanent magnets.

7.2.1 The use of a non-permeable rotor shaft

Despite the large axial length to outer diameter ratio of the motor, an end leakage flux arises, due to the permeable endcaps of the shaft [156]. To understand whether this could be solved through the use of non-permeable shafts, the machine of Chapter 3 was also modelled with a non-permeable shaft material, which is 304 Stainless-steel and rotor endcaps.

Figure 7-2 (a) shows the flux density map of the developed M3 model, while the results of both variations (i.e. the different shaft materials) are compared in Table 7-1. As can be seen in Table 7-1, when adopting a non-permeable shaft, the average torque is 6.64% lower compared to permeable shaft. This is mainly due to the removal of the magnetic rotor back iron and thus no path for the magnetic flux produced by the PM.

This indicates that a considerable reduction in the overall performance of the machine is achieved with the non-permeable shaft solution. Therefore, this solution is considered as not suitable for this work. For the permeable shaft solution, even when considering the end-leakage in the rotor endcaps, the results show the benefit of using a permeable shaft.

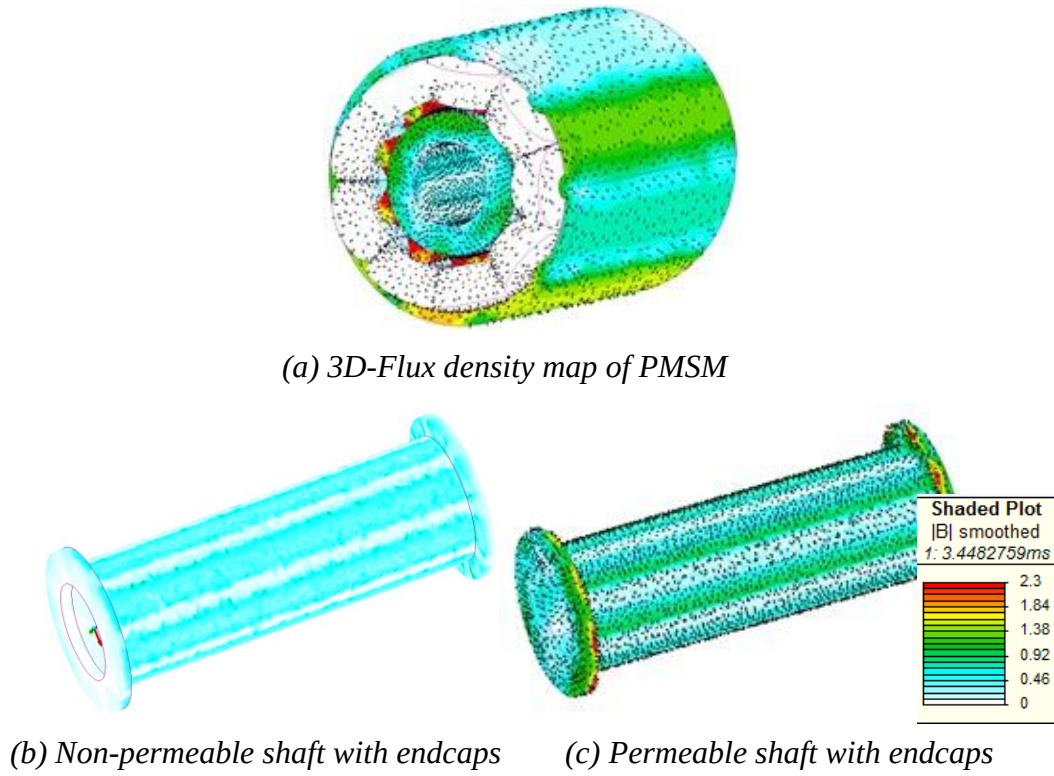


Figure 7-2: Flux density distribution in the M3 model (80A, 8700rpm)

Table 7-1: Average torque of the investigated machine

	Permeable shaft	Non-permeable shaft
Average Torque (Nm), 3D-FEA	10.08	9.41

7.2.2 Mechanical benefits of the end-caps

To demonstrate the benefits of the rotor endcaps, axisymmetric FEA structural analyses have been conducted on rotor with and without endcaps. The PMs are placed around the shaft, whereas the sleeve is applied with 0.004 mm interference fit. Figure 7-3 presents the axial deformation in the rotor with and without endcaps at 20krpm rotor speed. Axial deformation is avoided, when endcaps are included.

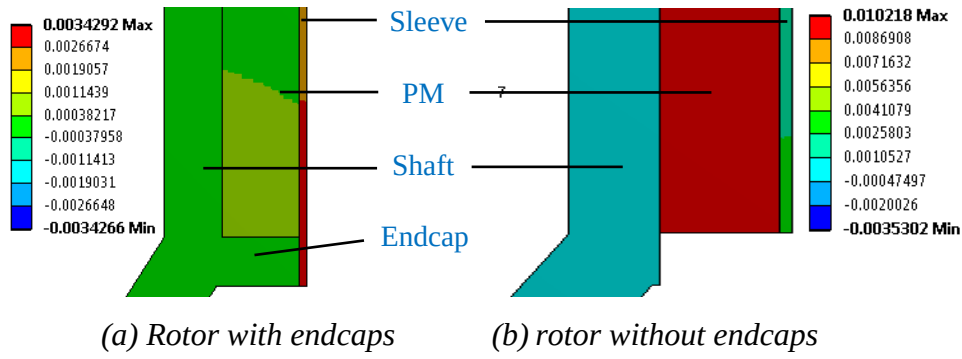


Figure 7-3: Axial deformation (mm) in the PMSM rotor with and without endcaps

Hence, it can be perceived that endcaps prevent potential elastic and/or plastic deformations, which might lead to the PMs breakage during rotor assembly. It is worth noting the thickness of the endcaps does not affect the axial deformation since the mechanically stiff shaft steel provides support for the magnets even with very thin endcaps. However, relatively thick endcaps are normally selected due to manufacturability aspects.

7.2.3 Electro-magnetic aspects of the end-caps

To understand the electromagnetic aspect of the endcaps, three different magnetic steels are investigated and compared in section. To ensure accuracy in terms of the considered material properties, material characterization was applied and the actual BH curves shown in Figure 7-4 (a) were used in the analysis. According to the results presented in Figure 7-4 (b), there is a small change in the output torque when the different materials are used for the rotor shaft. This is mainly because the shaft provides enough magnetic path for the Quasi-Halbach magnets. While, a significant difference in losses can be observed in Figure 7-4 (b). This is due to the difference in the material properties (i.e. electrical resistivity) as presented in Table 7-2. In addition, different endcap thicknesses are used to understand their influence on the electromagnetic performance. For different materials, the torque decreases with the thickness of the endcaps while the losses increase as shown in Figure 7-4 (b). This mainly due to the increase in the leakage flux at the rotor endcaps as well as due to the increase in the eddy current paths.

From the above analysis it was clear that a magnetic stainless steel (17-4 PH) was a preferred solution for the hollow rotor shaft due to its high electrical resistivity. Since the difference in torque is $\leq 1\%$ between the 1.5 and 2.5 mm endcap thicknesses, 2.5 mm has been selected in order to simplify the manufacturing process.

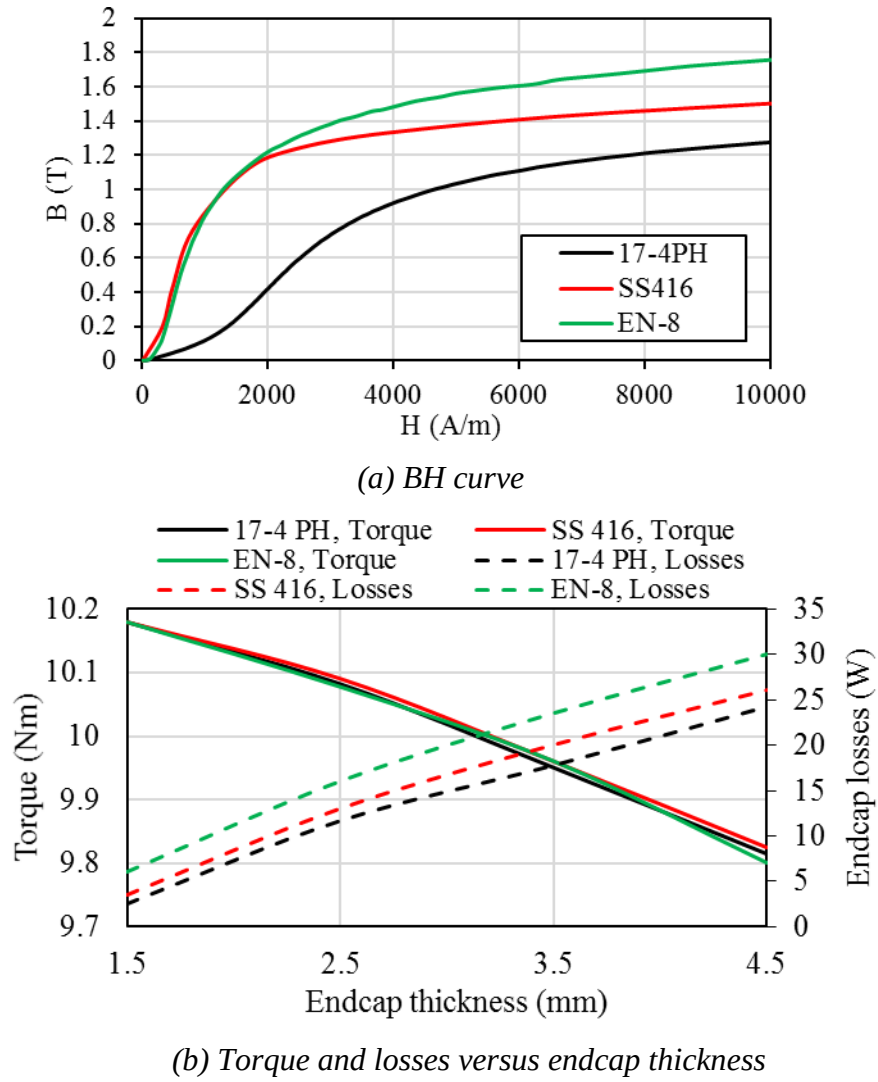


Figure 7-4: The influence of different endcap material

Table 7-2: Material properties

Parameters	17-4 PH	SS-146	EN-8
Density (kg/m^3)	7820	7750	7780
Electrical resistivity ($\Omega\cdot\text{m}$)	7.7×10^{-7}	5.7×10^{-7}	2.2×10^{-7}
Thermal conductivity ($\text{W}/(\text{m}\cdot\text{C})$)	19.55	24.9	33
Yield/tensile stress (MPa)	1379	700	850

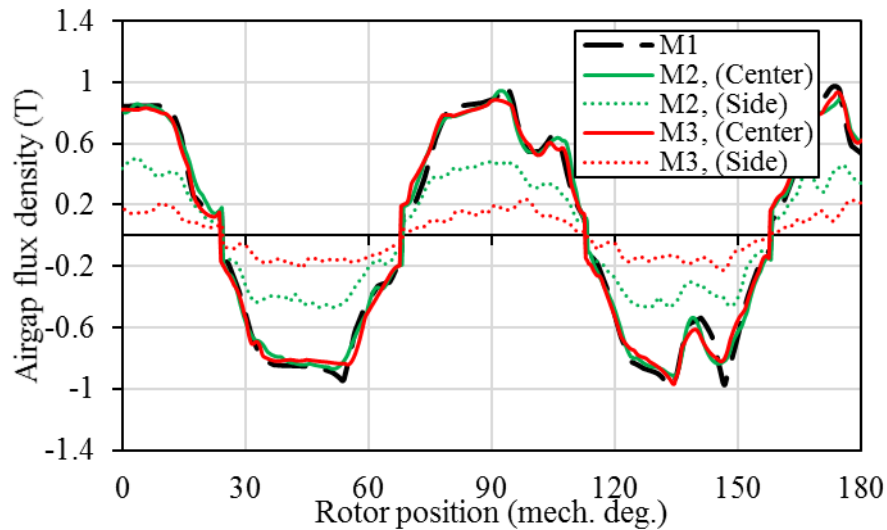
7.3 Influence of endcaps on electromagnetic performance

In order to fully understand the significance of the proposed model M3, a comparison between the resulting airgap flux density, back-EMF and output torque from the different models is presented in this section.

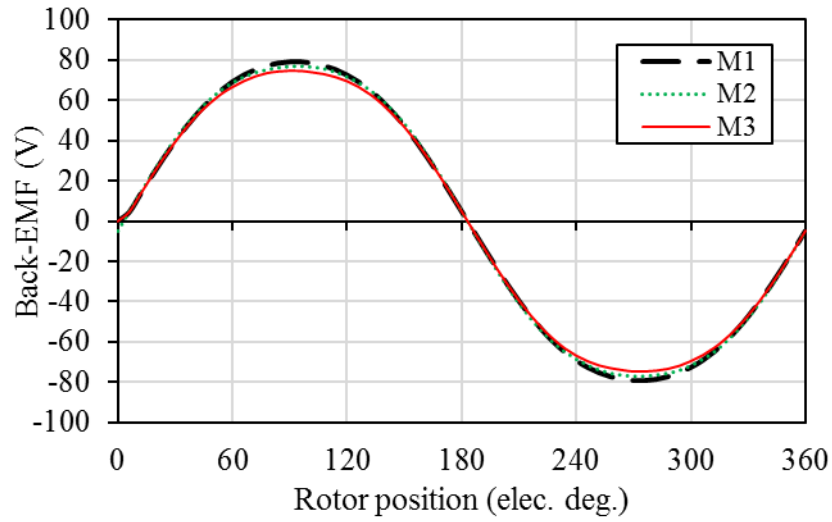
7.3.1 Open circuit

The airgap flux density for all the three different models is calculated and the results are shown in Figure 7-5 (a). In M2 and M3 models, airgap flux density has been detected in two regions: a) at the middle of axial length and b) at one rotor end (inner edge of the front endcap). Lower amplitude and higher distortion of the airgap flux density is observed at the end of the machine, due to the influence of the end leakage. Considering the airgap flux density at the middle of the axial length, a similar distribution is obtained for all the models, although the M3 distribution reveals lower peaks, this is due to axial leakage in the 3D-FEA.

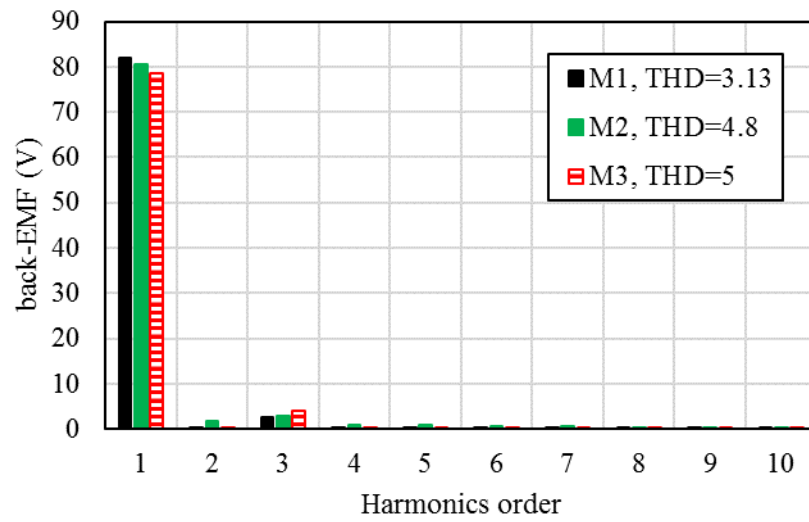
Likewise, a lower back-EMF amplitude is found when M3 is considered as shown in Figure 7-5 (b) and (d). A slightly higher total harmonic distortion (THD) is observed, as reported in Figure 7-5 (c) and (e). The 1st harmonic reduction is a consequence of the end-leakage, due to the endcaps, while the higher THD is mainly given by the increase of the 3rd harmonic. In general, a double layer concentrated winding PMSM with a 9/8 slot/pole combination inherently has a high 3rd harmonic in the back-EMF [160]. In the M1 analysis done here, the tooth tips of the machine are quite saturated, and this has the effect of partially suppressing the 3rd harmonic. However, when the machine is analysed with a M3 model which also considers the end-effects, then this 2D effect is reduced and the 3rd harmonic in the back-EMF becomes more visible.



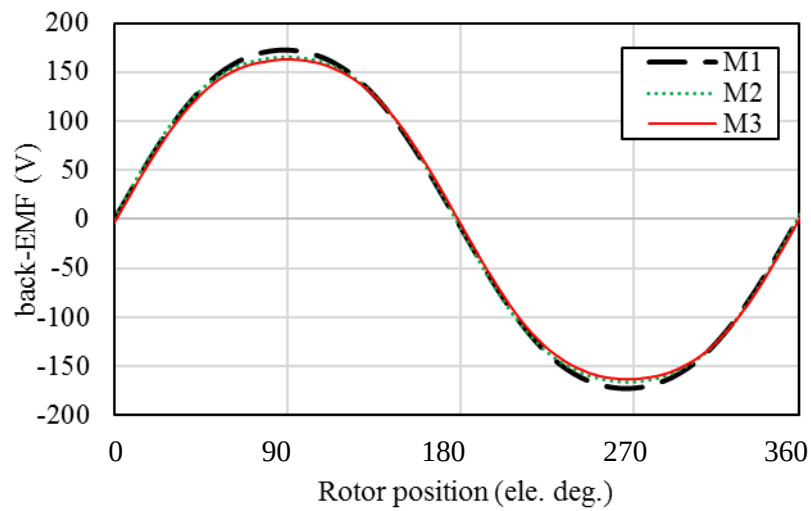
(a) Airgap flux density distribution



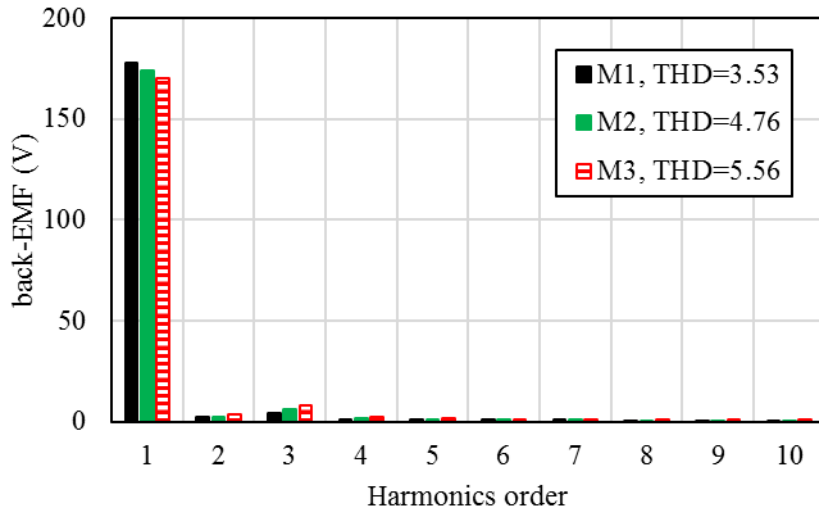
(b) back-EMF waveform at 8700 rpm



(c) back-EMF waveform harmonics at 8700 rpm



(d) back-EMF waveform at 19000 rpm



(e) back-EMF waveform harmonics at 19000 rpm

Figure 7-5: No load performance of the PMSM with / without endcaps

7.3.2 Electromagnetic torque

Figure 7-6 presents the electromagnetic torque of the PMSM predicted by M1, M2 and M3, when a current of 80 A (full load condition) is applied along the q-axis, while the d-axis current is equal to zero. Since the M1 and M2 models do not take into account the influence of the endcaps, the torque obtained from the M3 is 4.18% and 2.42% lower than the one calculated using M1 and M2, respectively. Considering the torque ripple, its value increases in the M3, due to the higher back EMF harmonic content.

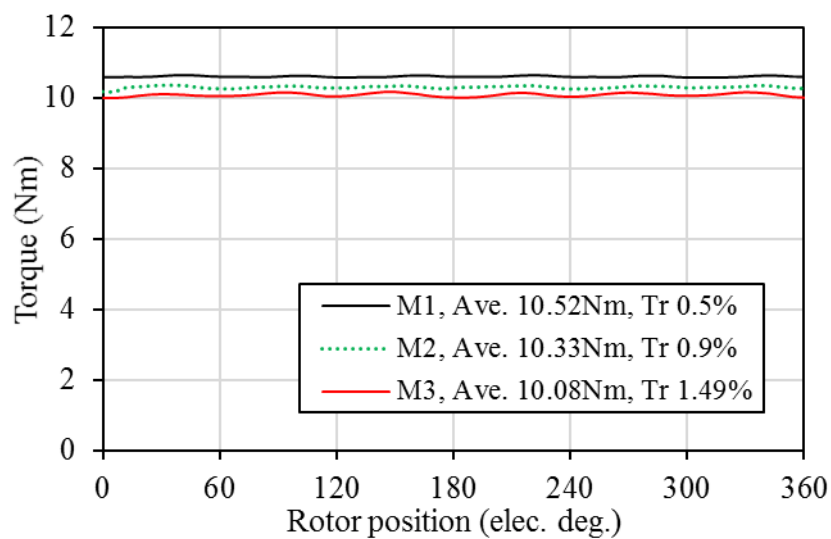


Figure 7-6: Torque waveform of the PMSM with and without endcaps at 8700 rpm

7.4 Influence of endcaps on torque-speed characteristics

In order to increase the prediction accuracy of the torque-speed characteristics, the end-effect influence needs to be considered. To determine the torque-speed characteristics the PM flux-linkage, the d-axis and q-axis inductances are predicted and analyzed.

7.4.1 PM flux linkage

The influence of the magnetic saturation on the PM flux-linkage is investigated in this section. The magnetic flux along the d-axis (Ψ_d) is sum of two terms, the PM flux (Ψ_{pm}) and the flux produced by the d-axis current ($L_d * I_d$).

$$\Psi_d = \Psi_{pm} + L_d * I_d \quad 7-1$$

The variation of the d-axis flux (Ψ_d) as function of the q-axis current is determined by running several FE simulations at different q-axis current values, while the d-axis current is equal to zero. Since the d-axis current is null, its correspondent flux term will result equal to zero.

$$\Psi_d(I_d=0, I_q) = \Psi_{pm}(I_q) \quad 7-2$$

Hence, the variation of the PM-flux with q-axis current is obtained. The described procedure has been repeated using M1, M2 and M3 and the results are shown in Figure 7-7. Due to the magnetic saturation, all the PM flux-linkage values decrease as the q-axis current increases. Although this trend is common for all the models, the PM flux-linkage predicated via M3 reduces at faster rate compared with the M1 and M2, due to the endcaps influence. Since, the presence of permeable rotor endcaps increases the leakage flux at the front and back ends of the rotor. This effect leads to a reduction in the flux-linkage.

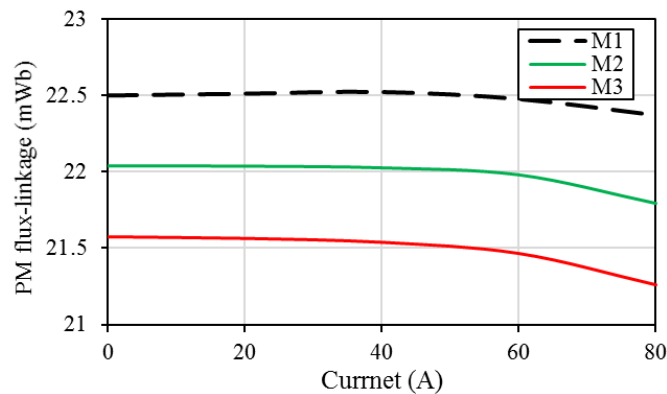


Figure 7-7: PM flux-linkage at different q-axis current levels

7.4.2 Inductance

Using M1, M2 and M3 the d-axis and q-axis inductances are calculated and presented in Figure 7-8. Generally, including the end-effects in the calculation of the inductances lead to higher values compared to the M1 predicted results. This is due to the end leakage inductance, which is quite high in M3 model. It can also be seen that the magnetic saturation is higher, therefore at high current levels lower inductance values are observed due to the high stator leakage similar to the PM flux-linkage

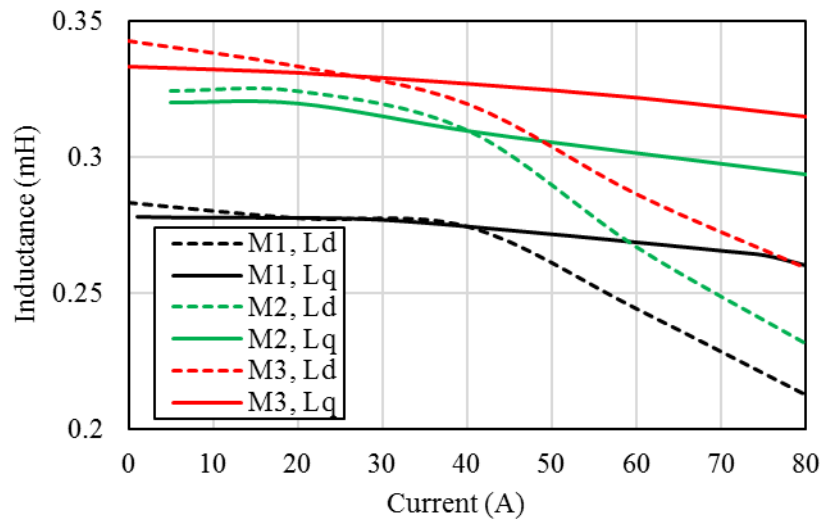


Figure 7-8: D-axis and q-axis inductances

7.4.3 Torque and power speed curves

The PM flux-linkage and inductances values obtained through the FE models have been used for plotting both torque-speed and power-speed curves, as shown in Figure 7-9.

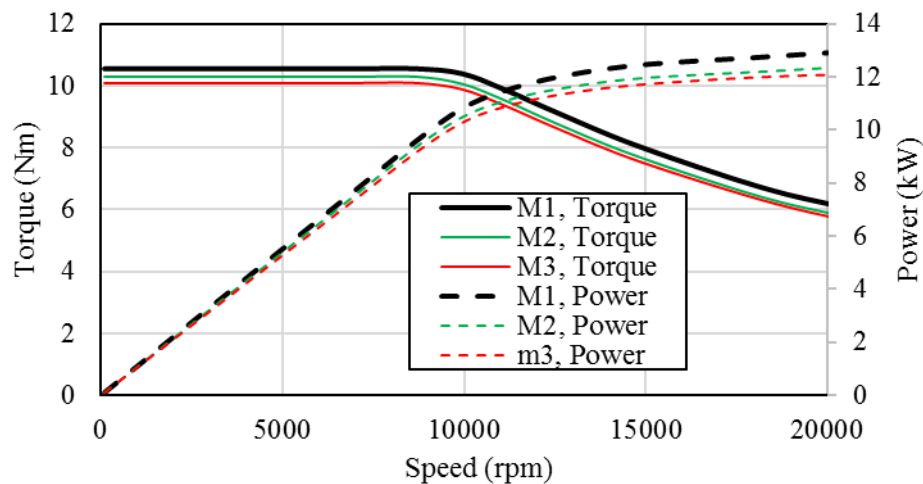


Figure 7-9: Torque-speed characteristics

Considering Figure 7-9, the M3 results provide lower torque and base speed values for the constant torque region. This outcome was expected since the PM flux-linkage predicted by M3 is lower. Additionally, a reduced power and torque can be observed in the flux weakening region. The flux weakening factor (K_{fw}) has been calculated by using (7-3), where I_{max} is the maximum phase current [161].

$$K_{fw} = \frac{L_d I_{max}}{\psi_{pm}} \quad 7-3$$

When the M1 are used, the K_{fw} is equal to 0.95, while it boosts to 1.23 in case of M3. The significant increase is justified by the double-action due to the low PM flux linkage and the high d-axis inductance.

7.5 Influence of endcaps on core and rotor eddy-current losses and efficiency

The objective of this section is to illustrate the difference in the losses that arise from using different models and their effect on machine efficiency.

7.5.1 Core losses

Figure 7-10 presents the core losses of the PMSM at rated speed of 8700 rpm for several q-axis currents. At low current levels, i.e. below 60 A, the core losses predicted by M3 are lower than that of the M1 and M2 due to the leakage effect, which results in lower magnetic flux density in the machine core. Conversely, at higher current levels, i.e. above 60 A, the M2 predicted core losses exceed that of the M1 and M3. This is due to the flux produced by the current flowing inside the end-windings. While the M3 experiences a high leakage flux in the stator core due to the permeable endcaps, resulting in lower core losses compared to the M2.

7.5.2 Rotor losses

For each rotor component, the eddy-current losses have been determined by using M1, M2 and M3. Two 3D FE models, i.e. with and without endcaps, have been considered. Figure 7-11 shows the trend of the total eddy-current losses for the three cases under study

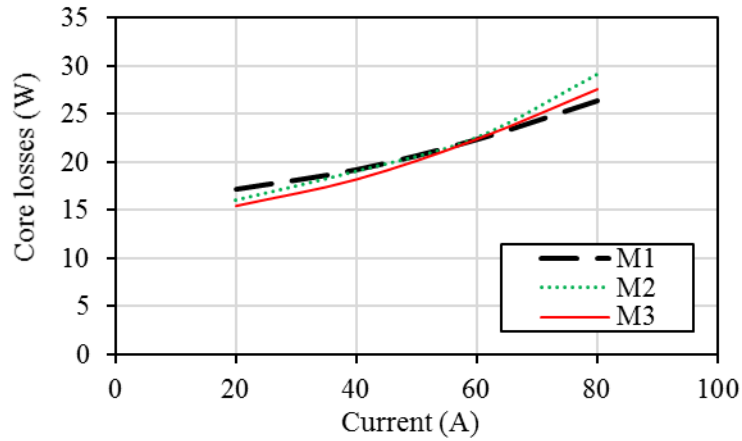


Figure 7-10: Core loss at different q-axis currents

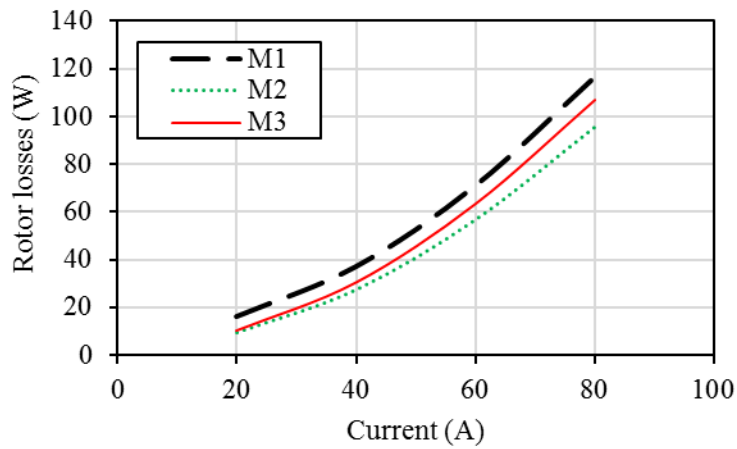


Figure 7-11: Rotor loss at different q-axis currents

Considering the same load condition, lower eddy current losses are observed in both M2 and M3 compared to the M1 (see Figure 7-11). This is the result of lower leakage flux in M1, which does not model end-paths and restrains the flux lines on a plan. Since a higher amount of “active” flux is considered by the M1, the resulting eddy current losses are higher. In M2, lower eddy losses are obtained when the endcaps are neglected. Indeed, by including the endcaps within the model, a further circulating path for the current is added. Although it is stated that computing the eddy-current loss in M2 and M3 might cause an increase in the rotor losses compared to the M1 since the induced current paths in the rotor will close at the end region leading to longer current path. Lower eddy current losses are observed in the M2 and M3 when compared to the M1, since the rotor loss is mainly due to the high order space harmonics which has small end path compared to the rotor axial length, shown in Figure 7-12.

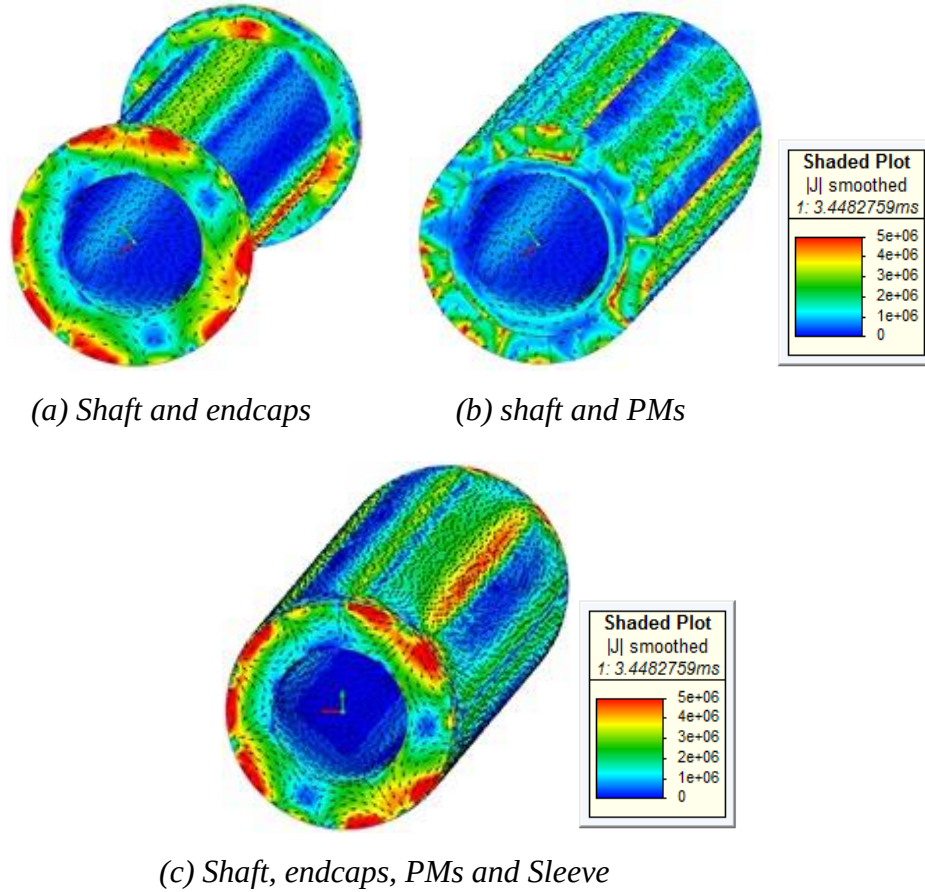


Figure 7-12: Eddy-current density distribution in rotor components of M3 model

Additionally, the PMs span is relatively small compared to the eddy current path in the axial direction. Therefore, the reduction in the losses computed by M2 and M3 compared to the M1 due to the leaked flux is higher than the additional losses due to the eddy current at end paths of the rotor. Also, the losses in the endcaps increases at high rate with the current level, since the last determine a high stator saturation forcing the magnet flux to pass through the endcaps.

7.5.3 SPM efficiency

The loss terms previously predicted by different models have been employed to determine the machine efficiency at several current and speed levels, using (7-4).

$$\eta = \frac{P_{mech}}{P_{mech} + P_{cu} + P_{core} + P_{eddy}} \quad 7-4$$

The PMSM efficiency results are summarized in Table 7-3. Due to the large amount of copper losses, the influence of the endcaps on the PMSM efficiency can be observed,

however it is clear that is quite low, when compared to the copper loss, even though it does have a reductive effect on the mechanical power and changes in the core and rotor losses.

Table 7-3: Efficiency at different input conditions

Operation condition	M1	M2	M3
8700rpm, 20A peak current, 0 current angle	93.5	93.59	93.66
8700rpm, 40A peak current, 0 current angle	93.59	93.6	93.62
8700rpm, 60A peak current, 0 current angle	92.07	92.05	92.03
8700rpm, 80A peak current, 0 current angle	89.83	90.06	89.77
19000rpm, 80A peak current, 53.5 current angle	87.64	87.93	87.21

7.6 Performance evaluation

In order to illustrates the effectiveness of including the rotor endcaps in the FE model of the PMSM, a comparison between the results obtained from the aforementioned investigational studies is presented in tabulated form in Table 7-4. The aim of this is to show the difference in the performance parameters resulting from the use of an inaccurate FE model that ignored the in-active material (rotor endcaps).

Table 7-4: Performance comparison of PMSM with and without rotor endcaps

	M1	M2	M3
Power (kW) @ 8700 rpm (based speed)	9.57	9.4	9.17
Power (kW) @ 19000 rpm (max speed)	12.46	12.27	11.75
Torque (Nm) @ 8700 rpm (based speed)	10.52	10.33	10.08
Torque (Nm) @ 19000 rpm (max speed)	6.27	6.17	5.91
Efficiency (%) @ 8700 rpm (rated load)	89.83	90.06	89.77
Efficiency (%) @ 19000 rpm (max speed)	87.64	87.93	87.21
Back-EMF (V) @ 8700 rpm (based speed)	78.91	77	74.68
Back-EMF (V) @ 19000 rpm (max speed)	174.41	168	162.95
THD (%) @ 8700 rpm (based speed)	3.13	4.8	5
THD (%) @ 19000 rpm (Max speed)	3.53	4.76	6.15
PM flux linkage (mWb)	22.37	21.7	21.3
L_d (mH) @ ($I_d=80A$, $I_q=0A$)	0.212	0.231	0.259
L_q (mH) @ ($I_d=0A$, $I_q=80A$)	0.259	0.293	0.314

7.7 Highlights of the chapter and contribution

The work related to this chapter has resulted in journal publication [130]. The main highlights and contribution for this chapter can be summarized in the following list:

- 1) A detailed analysis of permeable rotor shaft with endcaps and its effect on the electromagnetic torque of the PMSM considering different magnetic steel materials with different BH curves and different endcap thicknesses.
- 2) A detailed investigation of permeable rotor shaft with endcaps and its influence on the electromagnetic machine performance (airgap flux density, back-EMF, THD and torque).
- 3) Investigation of the influence of rotor endcaps on the torque-speed and power speed characteristics, stator core and rotor losses and machine efficiency.

7.8 Conclusion

In this chapter, a detailed analysis on the effect of a permeable rotor shaft with endcaps on the machine performance has been presented. A 3D-FE model with rotor endcaps was developed so as to improve the accuracy of the modelling processes and thus the performance estimation for machines with inactive permeable material.

The proposed model M3 significantly improves on the conventional models M1 and M2 by its ability to map the end-effect leakage. In fact, this leakage has been shown to be higher than that predicted by the previous M1 and M2 models, proving the need for the extra modelling effort related to M3. The detailed 3D-FEA analysis for the proposed model M3 has also demonstrated how torque and power are negatively affected, due to the inclusion of the end-effects.

In conclusion, this work has shown that for high-speed PMSMs with permeable shaft assemblies, the predicted 2D-FEA results, i.e. not taking the permeable endcaps in consideration, are insufficient to predict accurately the machine performance. For the case study considered here, the difference in torque performance of the machine was

2.42%, i.e. when comparing the modelling results with and without the endcaps. This implies that while a model that does not consider the effects of leakage flux in inactive material is good for the general performance estimation, however it is subject to a performance prediction error. Therefore, in order to include the end effects caused by the rotor endcaps and to improve the accuracy of the machine performance estimation, it is necessary to use the M3 model, over the existing models M1 and M2.

With the knowledge gained in this chapter and as a final step towards a better design for the case study machine developed in Chapter 4, the decision was made to use the M3 model as shown in Chapter 8.

Chapter 8. IMPROVING THE PERFORMANCE OF THE CASE STUDY MACHINE THROUGH MORE ACCURATE MODELLING

8.1 Introduction

For the required industrial application, an electrical machine was designed, analysed and tested in Chapter 3, Chapter 4 and Chapter 5, respectively. For that design process, the used FE models were based on a number of specific assumptions, which ignored the effects of manufacturing processes and rotor endcaps, on the machine performance.

The effects of machining processes and of leakage in inactive material were investigated in the detailed analysis, presented in Chapter 6 and Chapter 7, respectively. These studies confirmed the hypothesis that the effects of manufacturing processes can have a significant impact on the overall machine performance, while more accurate results can be achieved by including the in-active material components (i.e. rotor endcaps) to the 3D-FE modelling.

In this chapter, as a final step towards better and more accurate modelling, the information gathered from Chapter 6 and Chapter 7 is used to update the original machine design of Chapter 4. This updated model allows for more accurate identification of the machine's losses and output torque.

8.2 Implementation of the accurate modelling

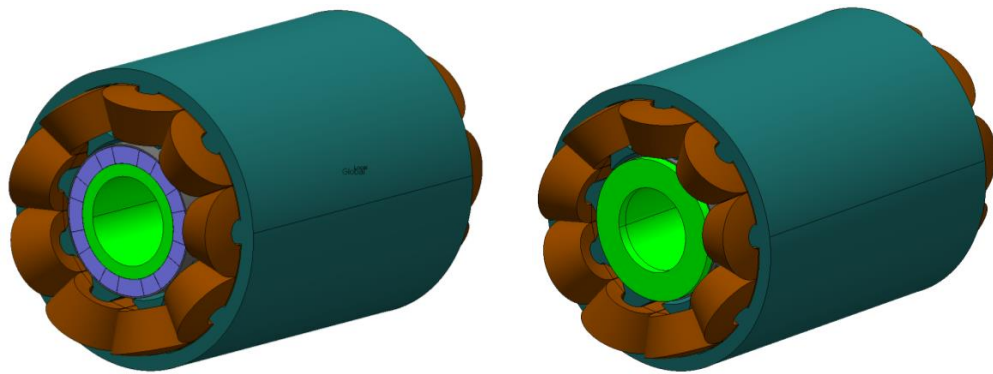
In this section, the material data (DS3) of the machine stator core derived from the material characterisations of Chapter 6 and the rotor endcaps from Chapter 7 will be used to update the original motor designed in Chapter 4. So as to achieve a better understanding of the potential advantages that can be gained by the updated model.

In order to do this, two FE models are used plus the experimental results, which are listed as follows

- 1) F1 – Original 3D-FE model (with DS1 + without endcaps).
- 2) F2 – Updated 3D-FE model (with DS3 + endcaps).
- 3) Exp. – Experimental

For F1 above, the model comes from Chapter 4. F2 is an updated model from Chapter 4 (i.e. including DS3 + rotor endcaps) for more accurate results. Exp. represents the experimental results of the prototype machine from Chapter 5.

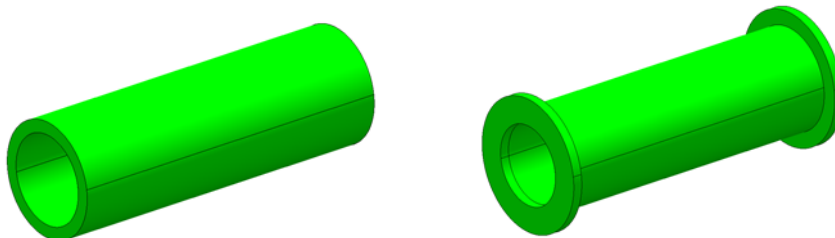
The models used in the analysis are shown in Figure 8-1. While the specifications for each FE model are shown Figure 8-2 and Figure 8-3.



(a) F1 model

(b) F2 model

Figure 8-1: PMSM: final FE models



(a) Without endcaps

(b) With endcaps

Figure 8-2: Permeable rotor shaft (from Chapter 7)

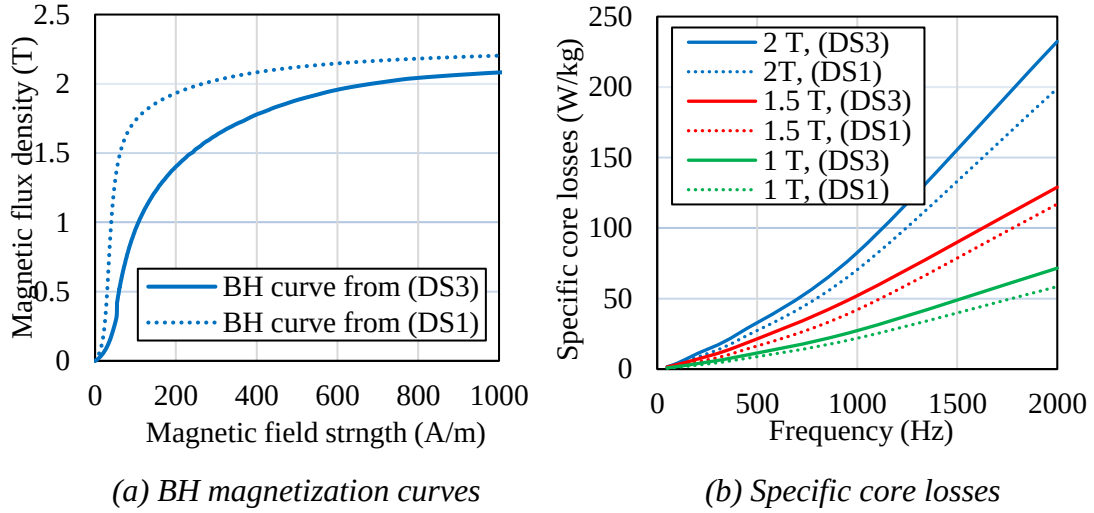


Figure 8-3: Comparison among the material properties (from Chapter 6)

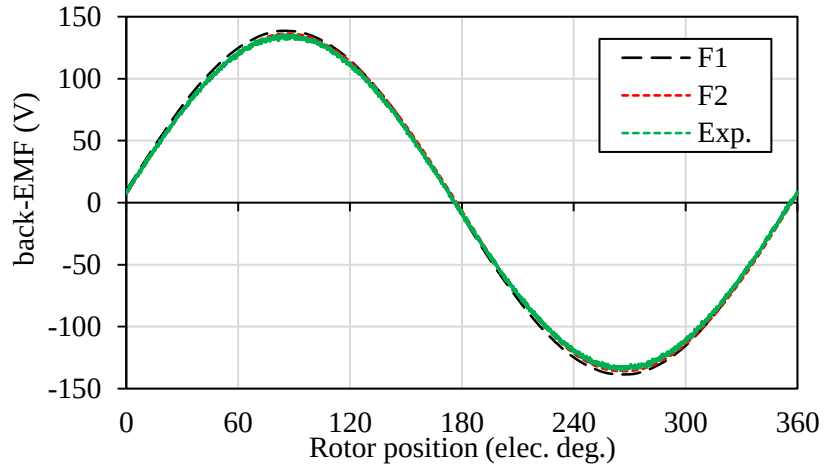
8.3 Results and discussion

Figure 8-4 to Figure 8-8 compares the performance of the PMSM with experimental results, using both F1 and F2 models.

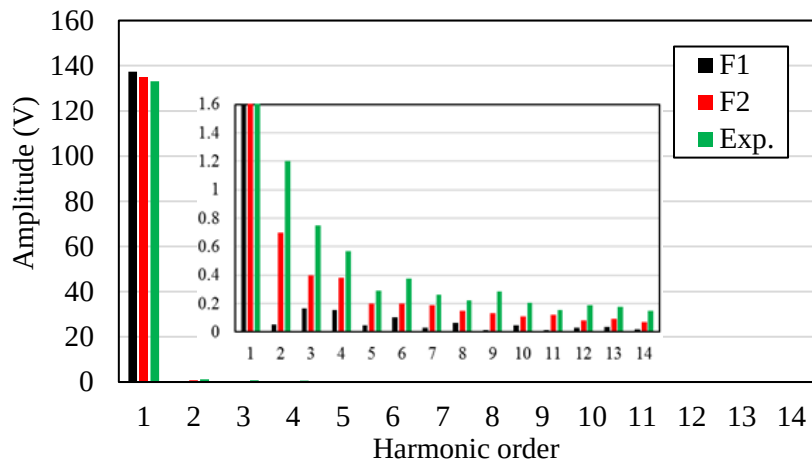
The machine line to line back-EMF waveform and its harmonics at 8700 rpm (based speed) and 19000 rpm (maximum speed) are reported in Figure 8-4 and Figure 8-5, respectively.

It can be clearly seen that the implementation of the F2 model only has a small effect on the back-EMF, mainly due to the influence of the permeable rotor endcaps. While including the DS3 has a negligible effect on the back-EMF as shown in (Table 6-3), mainly because the magnetic flux density is imposed by the magnets and maintained below the saturation level.

Although the measured values of the Back-EMF are slightly low at the maximum speeds with K_v 0.0363 Vs/rad. The obtained K_v from measurement is 1.1% lower than the one calculated using the F2 model.

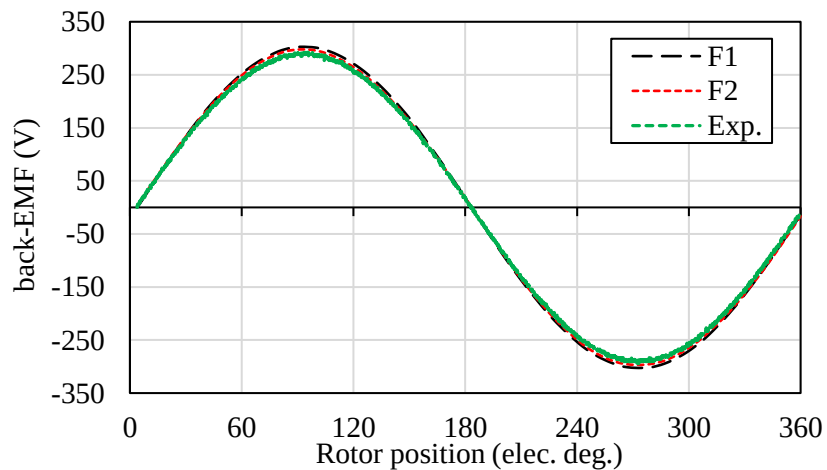


(a) Back-EMF waveform at 8700 rpm

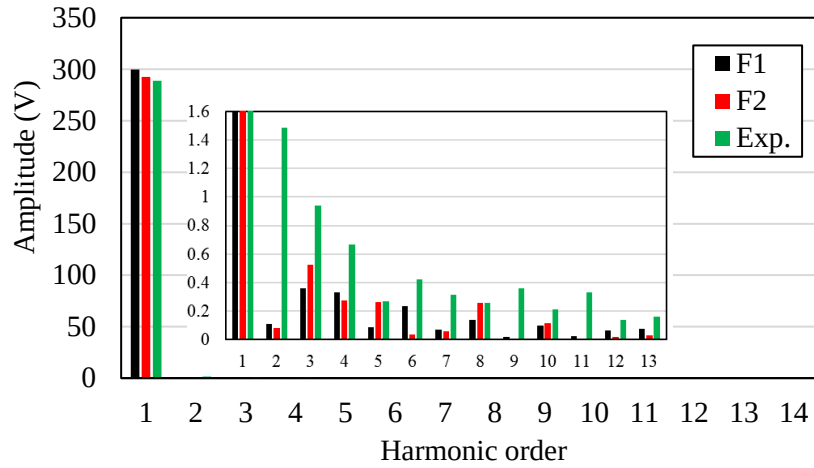


(b) Back-EMF harmonics at 8700 rpm

Figure 8-4: Comparison of the back-EMF at 8700 rpm



(a) Back-EMF waveform at 19000 rpm



(b) Back-EMF harmonics at 19000 rpm

Figure 8-5: Comparison of the back-EMF at 19000 rpm

As certain aspects were neglected (i.e. rotor endcaps and manufacturing processes), the F1 model overestimates the average torque by 4.37% (at rated current condition 80 A), as shown in Figure 8-6. The effect of these aspects on the output torque can be easily observed in the F2 model with more accurate results as found from Figure 8-6.

This variation becomes very small (negligible) at a low operating current of 20 A, as shown in Figure 8-6. This is due to a reduction in the saturation of the stator laminations, which results in a lower leakage flux.

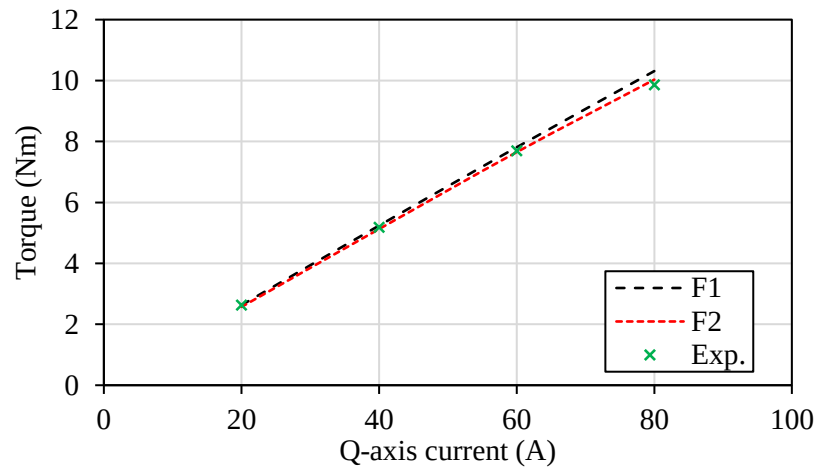


Figure 8-6: The measured and calculated torque versus current

Figure 8-7 shows a slightly higher losses for the F2 model which is mainly due to the increase in the additional losses caused by the machining processes of the stator core and the eddy currents induced in the rotor endcaps. For all the considered operating points, the F2 model allows better estimation of losses.

Once the reduction in the output torque and the increase in the total losses are defined, the efficiency of the case study machine can be determined.

According to the simulation results (i.e. F1 and F2 models), the inclusion of DS3 data and rotor endcaps in the F2 model reduces efficiency by 0.46% and 1.84 % at the base speed of 910 rad/s (8700 rpm) and the maximum speed of 1988.6 rad/s (19000 rpm), as shown in Figure 8-8.

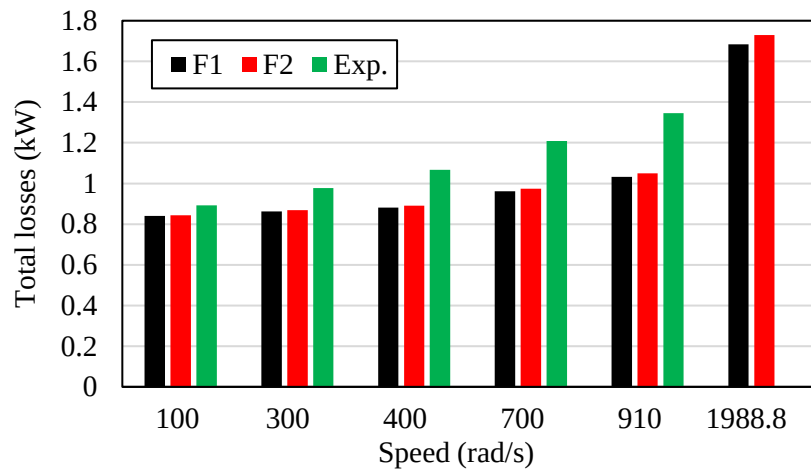


Figure 8-7: Losses at different operating conditions

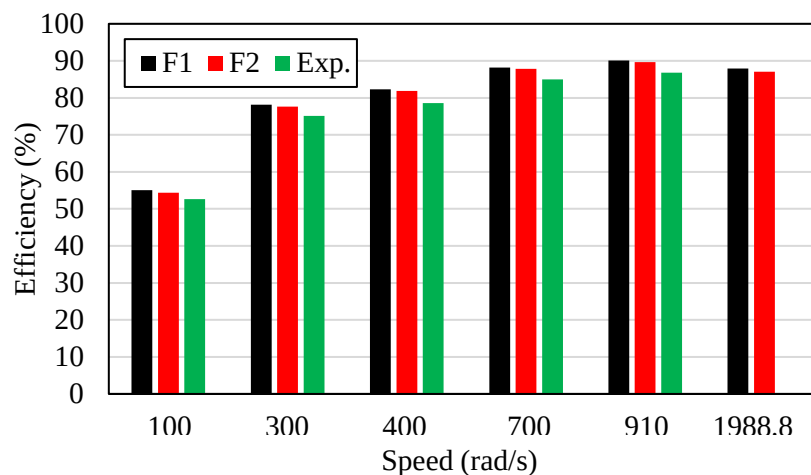


Figure 8-8: Efficiency at different operating conditions

Finally, the overall performance capabilities of the PMSM with F1 and F2 models are presented in Table 8-1, which clearly illustrates the advantages gained by including the effect of permeable rotor endcaps (3D effect) and stator core manufacturing processes in the FE model.

Table 8-1: Performance comparison of PMSM with different models

	F1	F2	Exp.
Power (kW) @ 8700 rpm (based speed)	9.4	9.13	8.96
Power (kW) @ 19000 rpm (max speed)	12.27	11.67	-
Torque (Nm) @ 8700 rpm (based speed)	10.33	10.03	9.84
Torque (Nm) @ 19000 rpm (max speed)	6.17	5.87	-
Efficiency (%) @ 8700 rpm (rated load)	90.06	89.67	86.83
Efficiency (%) @ 19000 rpm (max speed)	87.93	87.09	-
Back-EMF (V) @ 8700 rpm (based speed)	140.6	136.2	133.06
Back-EMF (V) @ 19000 rpm (max speed)	299.84	295.9	289.41

8.4 Considerations on the updated model

In order to fully understand and appreciate the importance of the updated model, a comparison of the performance parameters resulting from the above-mentioned models is illustrated in Table 8-2. The objective of this is to show how much extra current is needed to meet the torque requirement of 10.5 Nm and how this affects the machine efficiency and therefore temperature.

As shown in Table 8-2 to achieve the torque specification for the F1 and F2 models, the maximum phase current needs to be increased by 1.62 % (81.3 A) and 4.75% (83.8 A) respectively. As expected, a higher current was required for the F2 model, mainly due to the reduction in the output torque caused by the leakage flux of the rotor endcaps and the variation of the material properties of the stator core (i.e. DS3). This in turn indicates an increase in the machine losses of the updated F2 model, which in turn results in a variation in the efficiency and temperature of the electrical machine, as shown in Table 8-2. While the original F1 model shows that the machine can satisfy the required torque at a maximum winding temperature of 180.41°C with safety margin of 14%. With the updated F2 model, the temperature rises to 187.93°C.

Based on the results of this section, it can easily estimate the importance of considering these effects on the machines performance. Obviously, such an error in the estimation of the machine performance can have a severe impact on the electrical machine. Especially for the high performance aerospace application, where the power density of the machine is pushed to the limit to reduce the size and mass of the electrical machine.

Table 8-2: Performance of PMSM with new torque limit (10.5 Nm)

FEA model	Current		Temperature		Efficiency	
	Value (A)	Increase (%)	Value (°C)	Increase (%)	Value (%)	Decrease (%)
F1	81.3	3.07	180.41	4.16	89.92	0.64
F2	83.8		187.93		89.34	

8.5 Final recommendations for the next generation design of case study machine

Based on the updated FE model described above (which includes all the advantages coming from the previous chapters), this section is aimed to provide recommendations that may potentially allow for machine efficiency and power density improvements. Especially for high performance applications where low rotor weight and inertia are essential requirements, the following points are considered extremely important.

1. The use of thinner laminations for the stator core:

According to the simulation results (see Figure 6-10 (b)), a loss that reduces the machine efficiency by 0.2% (at the base speed of 8700 rpm) takes place in the stator core when using a DS3 material. In fact, this loss value can be reduced by using ultra-thin laminations at the cost of more expensive material and less mechanical strength.

2. The use of non-permeable material for the rotor shaft with endcaps:

According to the results presented in Table 7-1, the use of non-permeable material has have negative impact on the machine performance, mainly due to the absence of the magnetic circuit (i.e. rotor back iron). Nevertheless, the

feasibility of using a non-permeable material for the rotor shaft becomes possible by the use of Halbach array with fully controlled magnetization for the PMs. Thus, no leakage flux through the endcaps and better air gap flux density with a more sinusoidal waveform. However, the main disadvantages associated with this configuration are its higher price and complexity of manufacturing.

3. The use of PMs overhang:

This solution involves increasing the magnet length in the axial direction, thereby increasing the magnetic flux density at the ends of the stack length.

4. The elimination of rotor endcaps:

This solution is possible when there are no mechanical advantages for the rotor endcaps.

Based on the above, the performance of the original machine can be improved without compromising the other specifications (i.e. overall machine weight and rotor inertia).

8.6 Highlights of the chapter and contribution

The main highlights and contribution for this chapter can be summarized in the following list:

- 1) Improvements of the machine performance prediction using the updated FE model.
- 2) A detailed investigation of the general performance characteristics (i.e. back-EMF, torque, losses, efficiency and temperature) of the case study machine with the updated FE model.
- 3) Numerical and experimental validation of the updated FE model.
- 4) Proposed recommendations for the design of PMSMs with a hollow rotor shaft.

8.7 Conclusion

In this chapter, based on the knowledge gained from using the actual material data (presented in Chapter 6) and rotor endcaps (presented in Chapter 7), a reliable FE model for the estimation of the machine performance has been presented and verified with experimental measurements at different operating points. This updated model was implemented and compared with original model from Chapter 4, with the aim of showing how the failure to accurately evaluate the machine's performance at the design stage could lead to a further reduction in the output torque and unwanted additional losses.

More specifically, in Section 8.3, the validity of the updated model was proven through the analysis of the no-load voltage, torque, losses and efficiency, which showed precise results with small errors compared to the experimental measurements (see Table 8-1). In Section 8.4, the performance of the original model and the updated model were compared again, but for the same output torque (i.e. 10.5 Nm). In this section, the key area of interest was highlighted in Table 8-2, which shows that the updated FE model required an additional 4.75% of the current to satisfy 10.5 Nm compared to 1.62% for the original model. This increase in the current requirements, led to a 0.64% decrease in the efficiency for the updated model, which in turn led to a 4.16% increase in the winding temperature. Finally, recommendations for an improved machine design are presented in Section 8.5.

Overall, the results presented in this chapter confirm the need for considering the effect of the manufacturing processes of the stator core and the effect of the supporting structure with permeable material (i.e. rotor endcaps) on the performance of the electrical machine to have more realistic values that could help to improve the machine design.

In conclusion, it should be noted that even a small improvement in the estimation of machine performance, as is found in this work, is important in the pursuit of highly efficient electrical machines.

Chapter 9. CONCLUSION AND FUTURE WORK

The work presented in this thesis has been focused on development of a high performance electrical machine for direct coupled pump application. This chapter presents the general conclusions and highlights the main outcomes that were observed throughout all the stage of the work. Additionally, some ideas for future work are also included to extend the finding of this thesis.

9.1 General conclusion

In general, it can be concluded that the design methodologies presented in this thesis have resulted in a machine design which satisfies the project requirements in terms of geometrical dimensions, weight and inertia and respects thermal limitations. Having a detailed knowledge of the power losses and the manufacturing aspects, was shown to be extremely beneficial in terms of improving the electrical machine performance through better machine design processes.

To this end, a ‘journey’ related to machine design and optimisation, core material machining and permeable rotor endcaps has been undertaken and presented in this thesis. Discussions on these aspects are given below.

9.1.1 Machine design and optimization

From a comprehensive literature survey on electric motors for aerospace actuators, it is inferred that surface mounted PMSM is better for the considered application due to its numerous advantages. While the selection of the electrical machine topology for the aerospace actuators is of critical importance, there is also a need to understand the power losses of such machine. Therefore, a detailed literature survey on the most commonly used methods for modelling power losses in PMSM was presented and discussed in Chapter 2.

Considering the main requirements and challenges associated with industrial application (see Table 3-1). Chapter 3 and Chapter 4 have focused a lot of effort on the

analysis, design and optimization of a PMSM. In particular this chapter has investigated a feasible methodology to improve the capabilities of electrical machine, especially in terms of power density and efficiency capabilities. This methodology involved the use of several techniques such as the use of hollow rotor shaft, the use of CoFe laminations, the use of Halbach arrays and the use of completely flooded cooling system (i.e. oil directly pushed through the windings and airgap). The successful implementation of these solutions has resulted in a high machine performance that satisfied the project requirements with power density of 4.87 kW/kg (i.e. including active and non-active material) at the maximum speed of 19000 rpm.

Finally, the experimental results on the manufactured prototype were presented in Chapter 5 to validate the machine design of Chapter 4. The motor showed excellent performance and was able to meet the required specifications, despite the 6.46% reduction in the output torque and the 3.25% reduction in efficiency at the base speed condition of 8700 rpm. The prototype has shown an output torque equal to 9.86 Nm, which is below the set objective of 10.5 Nm, and an efficiency of 86.93%. The results shown in Chapter 5 have then confirmed the validity of the proposed methods to improve the machine performance.

However, contrary to initial perceptions, it has been shown that ignoring the impact of the manufacturing processes of the stator core and the use of permeable material for the rotor shaft can also have a negative effect on machine performance. To validate all this a detailed research based around the developed machine was presented in Chapter 6 and Chapter 7.

9.1.2 Core material machining

Since the stator requires several machining processes, then its magnetic properties, such as the magnetization BH curve and the specific core loss, will change and this requires special attention. In this part of the thesis (i.e. Chapter 6) an investigation related to the effect of such processes on the machine performance was done. To consider the variations in the material properties due to the manufacturing process, the stator core has been experimentally characterized. It was shown that the BH curve and the specific core loss changed drastically when considering data from the measured core as opposed

to the material data from its datasheet and/or lamination sheet. This is mainly due to the degradation of the magnetic material properties. Although its influence was neglected at the early stage of the design, Section 6.5 shows the motor performance with more realistic results that take into account manufacturing processes. In particular, an increase of 23.5% and 16.7% in the stator core losses was observed at 8700 rpm and 19000 rpm respectively. Also, for such a machine a reduction of 0.8% in the output torque was noted.

In relation to all this, it can be concluded that if these variations are not taken into account, then these can have a significant effect in terms of the general performance characteristics of a prototyped machine functioning in a similar environment such as described in this thesis.

9.1.3 Permeable rotor endcaps

For an electrical machine that comprises permeable rotor shaft with endcaps, then it is clear that the rotor endcaps will be a major source of leakage flux and losses. Therefore, Chapter 7 has focused a lot of effort on understanding these effects. For the machine at hand, it was clear that ignoring the rotor endcaps led to significant error in the calculations of the main flux, inductance, output torque and losses. Mainly due to the increase in the leakage flux at the rotor ends. Thus, to have an accurate result, an updated FE model with rotor endcaps was presented, based on the original model developed in Chapter 4. The updated FE model significantly improved on the conventional FE model by its ability to map the end-effect leakage, which was higher than predicted by the conventional FE model.

The updated FE model was shown to provide more accurate results for the torque, losses and efficiency. It is therefore ideal for the electrical machine designer to include all the non-active components to have a better estimation of machine performance.

In this chapter, it was shown how the “endcaps” structure adopted for the 3D FEA of the developed machine, resulted in more accurate estimation for the overall machine performance. This would confirm that when the 3D-FE model equipped with non-active permeable component (e.g. rotor endcaps), the prediction errors of the model are minimal when compared to experimental results.

With the knowledge gained from the previous chapters, an updated version of the developed machine through complete modelling was presented and discussed in Chapter 8. It has been shown that the failure to accurately evaluate the machine's performance at the design stage (i.e. ignoring core machining and rotor endcaps) led to a further reduction in the output torque and unwanted additional losses.

Finally, based on the updated version of the developed machine, recommendations were proposed for the next generation machine design, which can lead to improvements in machine efficiency and power density.

In general, it can be concluded that this thesis has shown how the design processes of electrical machines can be improved by considering effects of machining processes of the stator core and the rotor endcaps. In fact, the accurate estimation of the machine performance can help the designer to minimize the error between the measured and the simulated results and thus make it possible to improve the performance of the machine by addressing a certain aspect without compromising overall performance.

9.2 Future work

Although much has been achieved in terms of concepts and their related tasks, there are several interesting research areas that can be investigated for future work:

9.2.1 Experimental tests

Apart from the 'standard' tests on the prototype machine as shown in Chapter 5, more realistic and missing-centred testing regime would allow for a complete characterisation of the machine. This includes

- Testing the developed machine in realistic environment, where the machine would be assembled in its intended application platform. This push towards higher technology readiness levels (TRL), goes beyond the scope of this PhD thesis, as this is the remit of the industrial partner. However, it would be a natural next step in terms of testing. This would require the design and

development of the whole EHA system including the pump plus the PMSM system described here.

- An extensive testing regime for all the mechanical aspects would be highly beneficial as it would investigate all the mechanical vibrations and acoustic noise that the developed machine is expected to be exposed to. Considering the relatively high speed of the system, i.e. 19000 rpm, then this represents an important step towards achieving a higher TRL for the machine. For this test, a sophisticated, high-resolution monitoring system needs to be implemented on each part of the motor components.

9.2.2 Machine design

The performance of the electrical machine can be improved through detailed optimisation on several parameter of the machine. To this end, further steps include

- Adoption of a multi-objective optimization for alternative PMSM designs, with different slot/pole combinations, winding configurations and different materials. This work will contain a guidance that will contribute to the design of PMSM for flooded pump applications.
- An investigation of the overhang structure for the machine developed in Chapter 4 in an effort to improve the power density performance. This solution considers the increase in the length of the permanent magnet in the axial direction while keeping the stator core stack length fixed. In PMSMs, the magnetic flux density at the ends of the stack length is low due to the leakage flux. Therefore, by increasing the rotor length compared to stator length in the axial direction, the magnetic flux density in the airgap can be increased and higher torque and power density can be achieved.

9.2.3 Modelling processes

In order to improve the modelling processes of the electrical machine with fast and more accurate results, the following aspects need to be considered in the future:

- An experimental investigation on the effects of different rotational speeds, airgap widths, rotor sizes, flow rate, fluid type and properties on the windage losses. This would help to provide a better understanding for the windage losses mechanism and identify the most effective parameters on the machine performance in case of flooded airgap machines especially at low temperature where the viscosity of the fluid increases significantly. This work would help to propose an empirical coefficient that can be used for the windage losses calculation.
- An investigation on the thermal effects on the material properties and the specific core losses in the stator core of PMSM and how this would affect the work presented in Chapter 6.
- An investigation on the effects of different geometrical dimension and material properties on the rotor endcaps (mentioned in Chapter 7) in an effort to develop an empirical coefficient that could help in the design stage of the electrical machine.
- Development of a complete analytical model (i.e. includes losses and saturation) to accurately evaluate the machine performance.

Appendix A. TECHNICAL SOFTWARE

A.1 Infolytica Magnet [162]

The MagNet is a FEA software which allows for accurate modelling of electromagnetics even when time variable fields with motion are involved and when non-homogenous, anisotropic and non-linear material are used. The FEA is a numerical technique that is based on the determination of the electromagnetic field distribution through the solution of Maxwell's equations. MagNet offers both two-dimensional (2D) and three-dimensional (3D) electromagnetic field simulations, all within the same user-friendly interface. This was the main commercial software used in this project, producing detailed simulations of electrical machine models.

A.2 MotorCAD [163]

The commercial package MotorCAD was mainly used for thermal analysis. From machine geometry, cooling method and power losses inputs MotorCAD builds a thermal network which is then used to predict the temperatures that the machine will reach. The thermal model in MotorCAD is based upon analytical lumped-circuit analysis making it extremely fast and efficient to calculate. All the thermal resistances and capacitances in the MotorCAD model are automatically calculated from geometric dimensions and material properties where the user does not need to be familiar with complex heat transfer phenomena such as dimensionless correlations for convection.

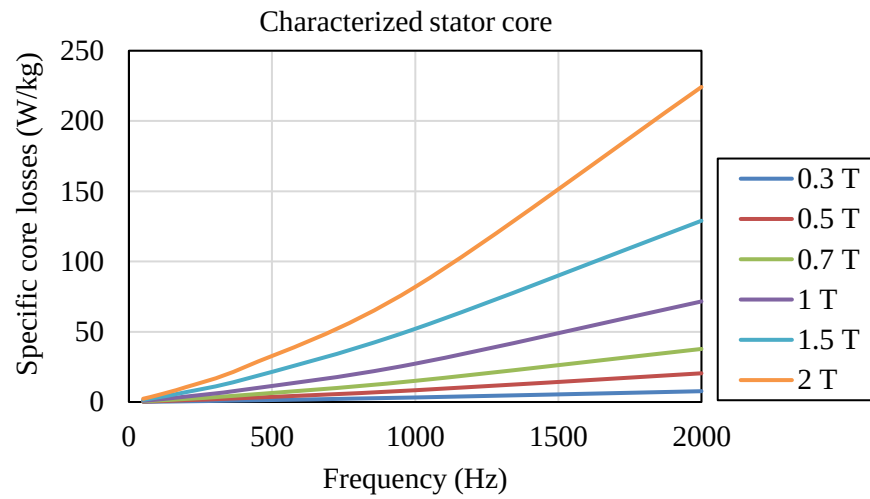
A.3 ANSYS [164]

To ensure the mechanical integrity of the rotor, mechanical stresses and dynamics were evaluated using a commercial software ANSYS Workbench. FEA software ANSYS was used in this study due to its friendly user-oriented environment. ANSYS offers the most accurate predictions of stresses and critical speeds, particularly for shafts with complex shapes and additional features, but at an increased computational time. It has also the ability to account for stress caused by variations in temperature and physical loads such as pressure, forces, accelerations and displacements.

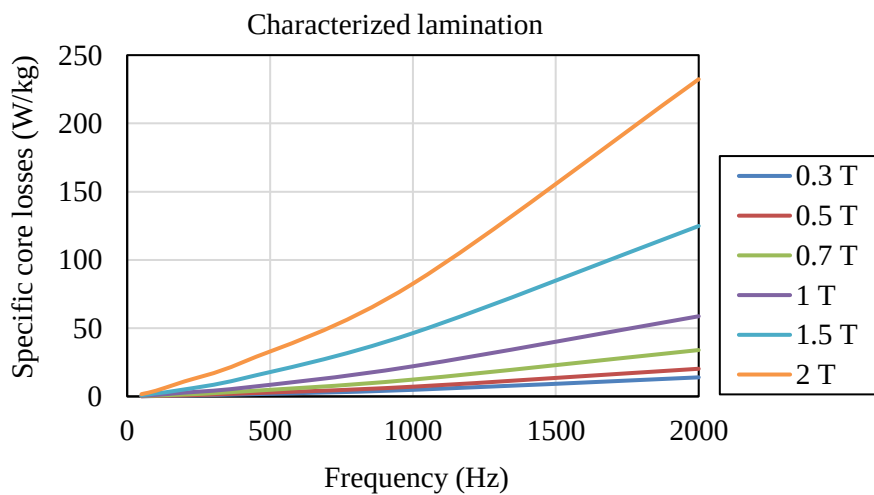
Appendix B. STATOR CORE MATERIAL CHARACTERIZATION

B.1 Coefficients extraction from the characterized stator core and lamination sheet.

The coefficients were used for the Modified Steinmetz equation in order to calculate the total iron losses in the stator core. Figure B-1 shows the specific core losses of Vacoflux 48 CoFe material at different values of frequency and flux density.



(a) Stator core



(a) Lamination sheet

Figure B-1: Specific core losses

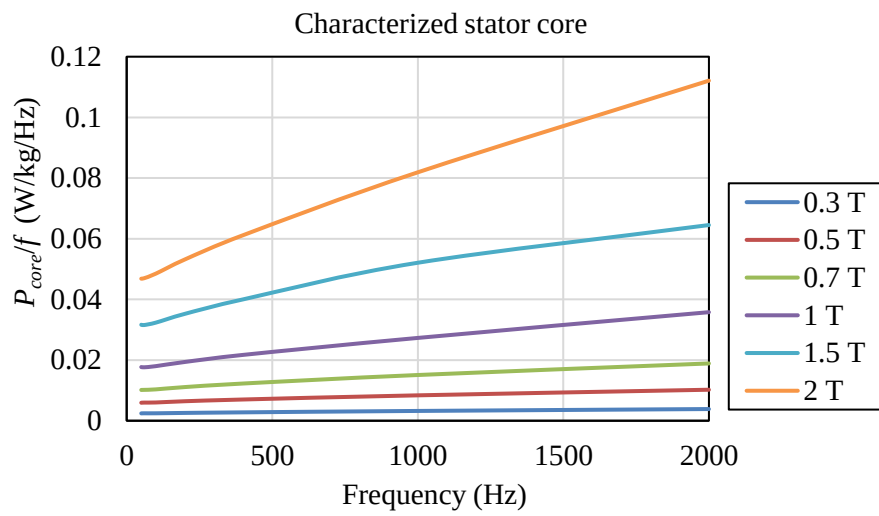
The Modified Steinmetz equation is given by

$$P_{core} = K_h f B^\beta + K_e f^2 B^2 \quad \text{B-1}$$

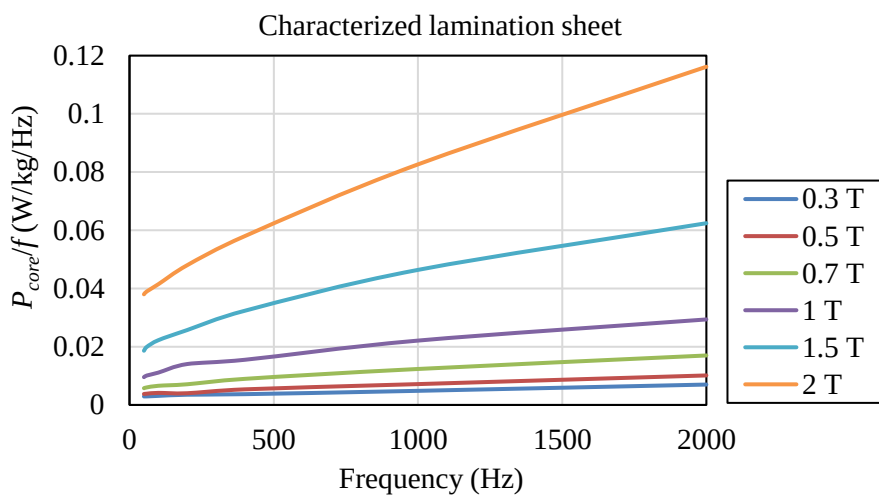
To get the values of β , K_h and K_e , the above equation is divided by the frequency f as follow:

$$P_{core}/f = a + bf \quad \text{B-2}$$

With the specific core loss data provided in Figure B-1, the (P_{core}/f) can be drawn versus different frequencies as shown in Figure B-2.



(a) Stator core



(a) Lamination sheet

Figure B-2: The relation between the P/f with different values of B

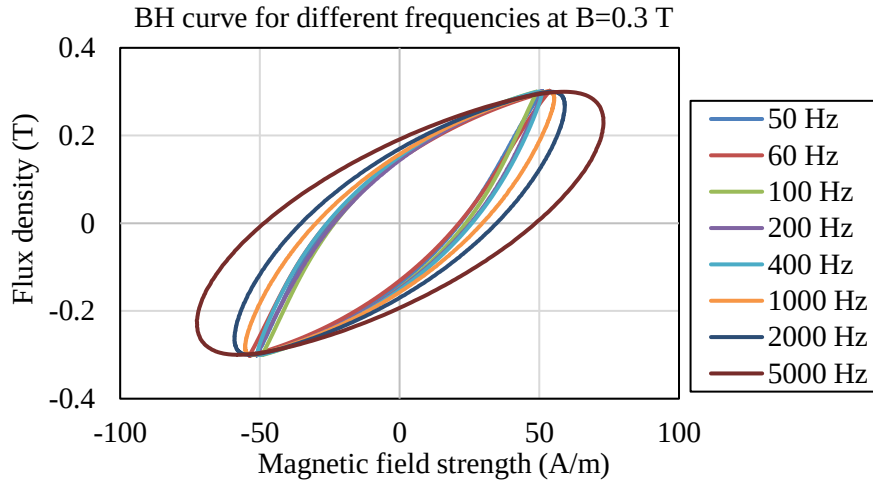
From Figure B-2, the interception of the P_{core}/f on the y-axis is $K_h B^\beta$ and the slope is identifying the $K_e B^2$.

The extracted coefficients from the manufacture datasheet and the characterized datasheet are shown in Table B-1.

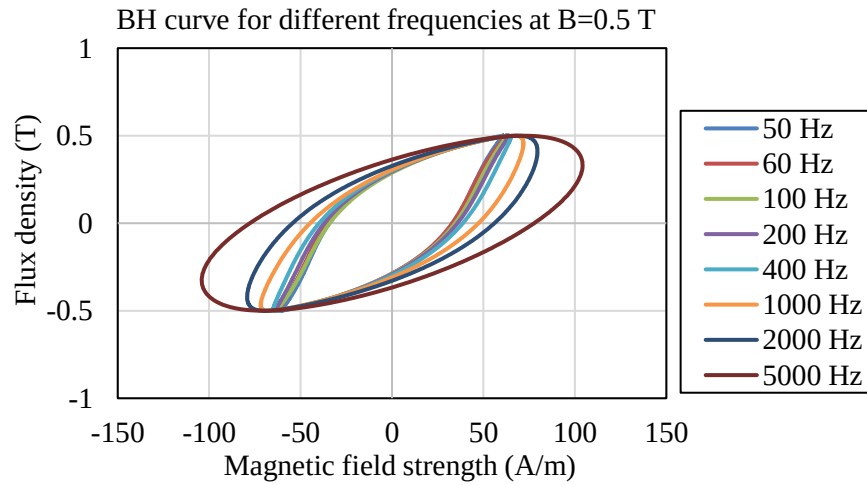
Table B-1: The coefficients of iron losses calculation

	K_h	K_e	β
Manufacture datasheet	8.476 e-3	6.344e-6	1.50253
Lamination sheet	6.2803	8.167e-6	1.32571
Characterized stator core	12.7677	5.514e-6	1.56153

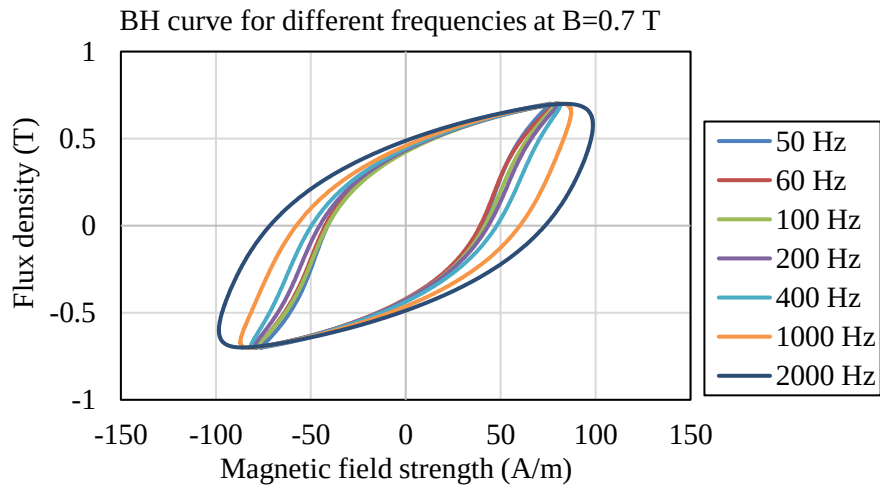
B.2 AC hysteresis loops at (50 Hz, 60 Hz, 100 Hz, 200 Hz, 400 Hz, 1000 Hz, 2000 Hz, and 5000 Hz).



(a)



(b)



(c)

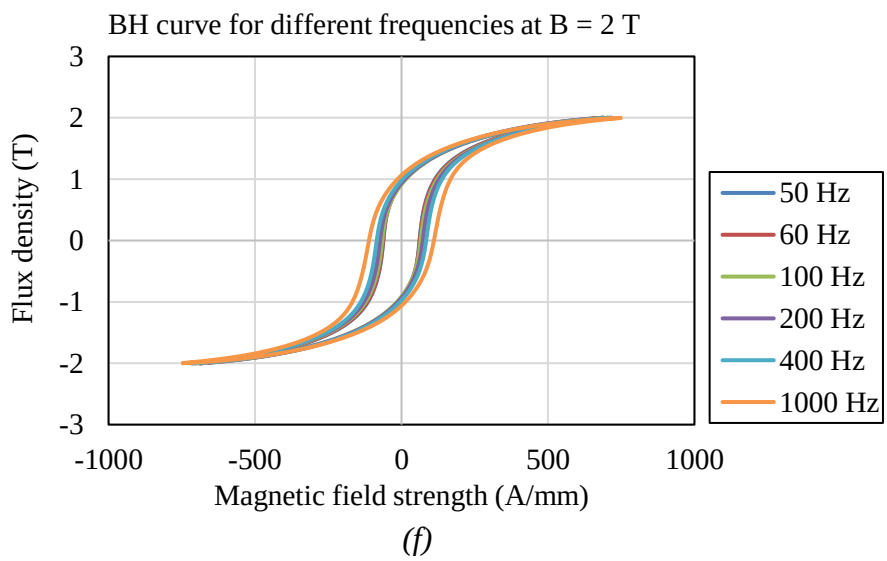
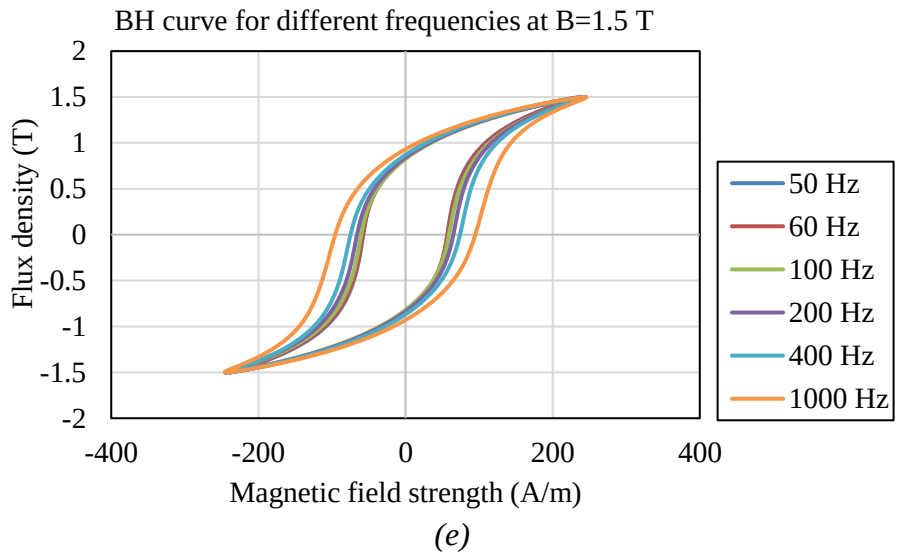
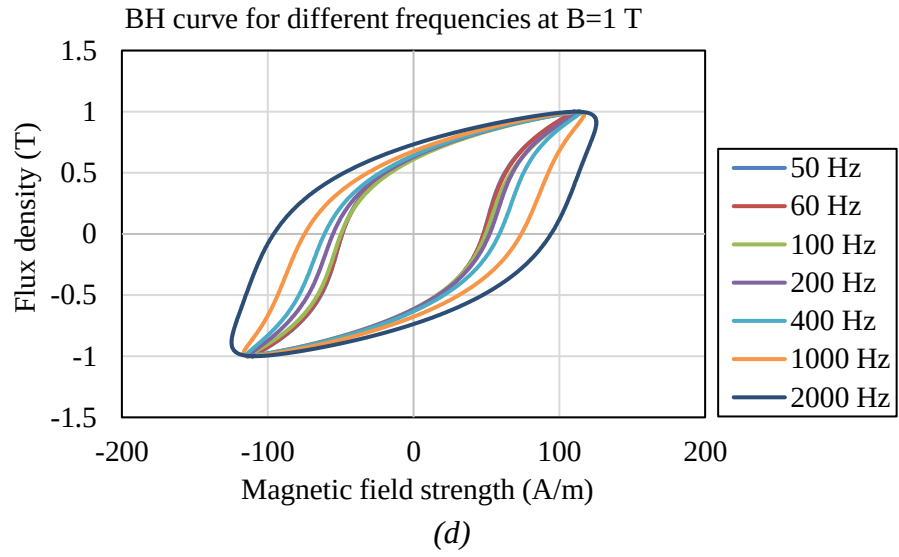


Figure B-3: AC hysteresis at different for the characterized core

Appendix C. THERMAL MODELLING

The thermal network of the PMSM used in Section 3.6 are detailed in this appendix.

Figure C-1 shows the schematic of the thermal resistance network. While, the key to symbols used in Figure C-1 are tabulated in Table C-1:

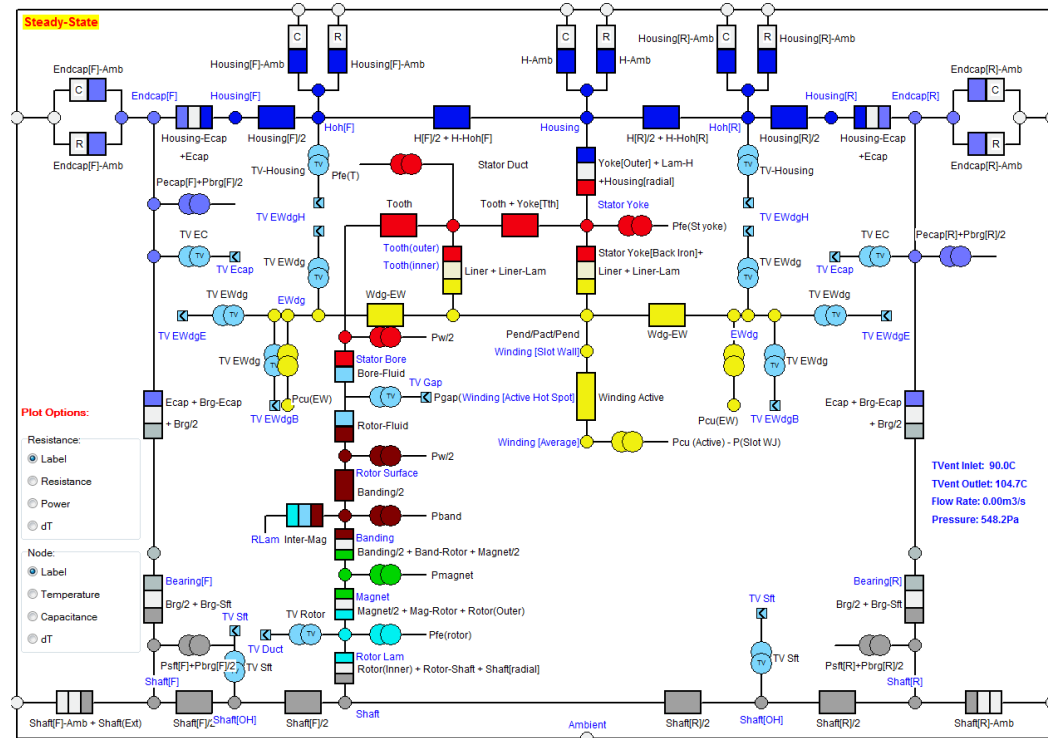


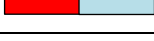
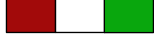
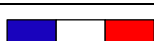
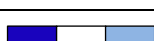
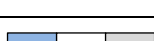
Figure C-1: Schematic of a PMSM thermal network in MotorCAD

Table C-1: Symbols description for the schematic of thermal network

Component	Description
	Power (Loss) Source
	Conduction Thermal Resistance
	Interface Thermal Resistance
	Convection Thermal Resistance
	Radiation Thermal Resistance
	Node
	Through Ventilation Cooling

Table C-2 defines the colour of the components represented by thermal resistance in Figure C-1.

Table C-2: Colour description for the thermal resistances of the LPN (MotorCAD)

Colour	Description
	Housing
	Stator Lamination
	Winding
	Magnet
	Rotor Sleeve or Banding
	Shaft
	Convection/Radiation from Housing
	Convection from Stator to Airgap Fluid
	Convection from Sleeve to Airgap Fluid
	Convection/Radiation from Endcaps
	Interface Gap Sleeve - Magnet
	Interface Gap Rotor Lamination - Shaft
	Interface Gap Magnet - Rotor Lamination
	Interface Gap Stator Lamination - Liner
	Interface Gap Housing - Stator Lamination
	Interface Gap Housing - Endcaps
	Interface Gap Endcaps - Bearing
	Interface Gap Bearing - Shaft

C.1 Thermal nodes

The nodes at different locations of the PMSM are shown in Figure C-2.

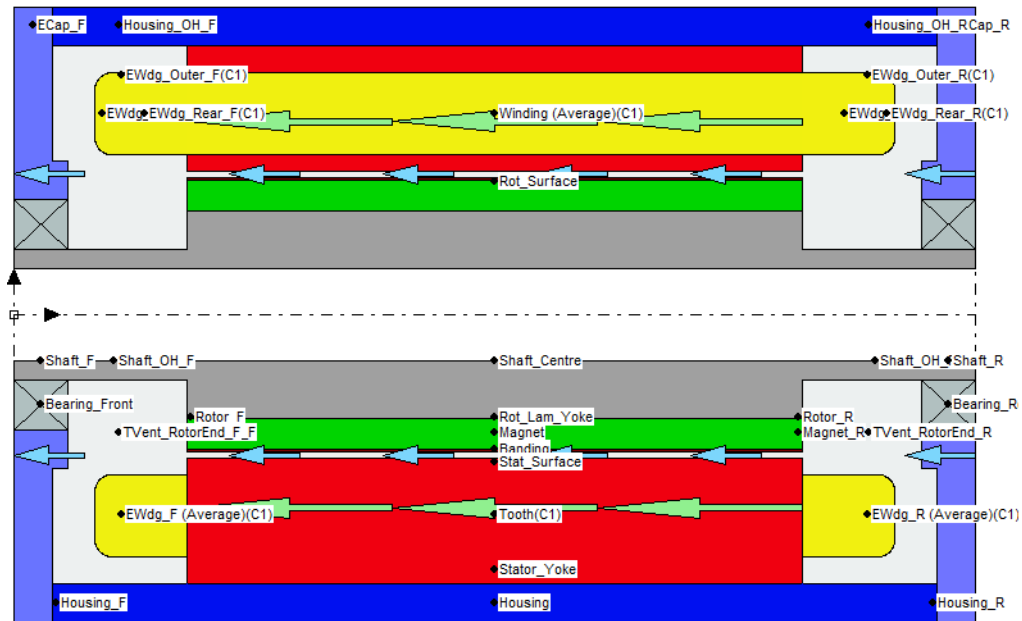


Figure C-2: Thermal nodes location for the PMSM

Thermal nodes from Figure C-2 are defined as:

- Shaft: Shaft Centre, Shaft (Overhang) Front, Shaft (Overhang) Rear, Shaft Front and Shaft Rear.
- Rotor: Rotor Lamination Yoke, Rotor Surface, Rotor Front and Rotor Rear.
- Magnet: Magnet Centre, Magnet Front and Magnet Rear.
- Sleeve or Banding: Banding Centre.
- Bearing: Bearing Front and Bearing Rear.
- Winding: End Winding Front (EWdg_Front), End Winding Rear (EWdg_Rear), End Winding Outer Front (EWdg_Outer_F), End Winding Outer Rear (EWdg_Outer_R), End Winding Front Average (EWdg_F Average), End Winding Rear Average (EWdg_R_Average), End Winding Rear Front (EWdg_Rear_F), End Winding Rear Rear (EWdg_Rear_R) and Winding Average.
- Stator: Stator Surface, Stator Yoke and Tooth.

- Housing: Housing, Housing Front, Housing Rear, Housing (Overhang) Front and Housing (Overhang) Rear.
- Endcap: Endcaps Front (ECap_F) and Endcap Rear (ECap_R).
- Through Ventilation: Through Ventilation Rotor End Front (TVent_RotorEnd_F) and Through Ventilation Rotor End Rear (TVent_RotorEnd_R)

C.2 Thermal conduction

The general form of a thermal resistance R_{cond} which describing the conduction heat transfer is

$$R_{cond} = \frac{L}{k_m A} \quad \text{C-1}$$

Where, L is the length of the component in the heat flow direction, A is the cross-sectional area and k_m is the thermal conductivity.

The regions where thermal conduction is occurred are represented by thermal resistances (mono-colour) as shown in Figure C-1.

In general, thermal conduction is depends mainly on the thermal conductivity of materials which can be high as 386 (W/m/C) for copper and low as 0.2 (W/m/C) for insulation material. Table C-3 show the material properties for different parts of PMSM.

Table C-3: Mass density, Heat capacity and thermal conductivity for the different parts of the machine

Material	Mass density (kg/m ³)	Specific heat capacity (J/kg/C)	Thermal conductivity (W/m/C)
Cobalt iron (VacoFlux 48)	8120	460	30
Silicon steel (M270-35)	7650	490	25
Copper	8954	383.1	386
Wire insulation	1400	1000	0.2
Sm ₂ Co ₁₇	8300	350	10
Titanium Ti-6AL-4V	4400	526	6.7
17-4 PH Stainless steel	7820	460	19.55
Air	1.2	103	0.026
AEROSHELL FLUID 31	851	2310	0.14
Aluminium	2790	833	168

C.3 Thermal convection

The general form of a thermal resistance R_{conv} which describing the convection heat transfer is

$$R_{conv} = \frac{1}{h_{ag} A} \quad C-2$$

Where, h_{ag} is the convection heat transfer coefficient.

The regions where thermal convection is occurred are:

- The cooling fluid in the airgap between the rotor and stator.
- The cooling fluid in the slot cooling channel.
- The cooling fluid in the end-space of the PMSM such as end-winding, rotor end shaft, magnet, housing and rotor endcaps.

1. The convection heat transfer coefficient for the stator-rotor airgap is the most important coefficient and can be calculated using (C-3):

$$h_{conv_airgap} = \frac{N_u \cdot k_m}{2 L_g} \quad C-3$$

Where L_g is the airgap length and N_u is the Nusselt number and it's a dimensionless parameter to reflect the enhancement of convective heat transfer due to the flow motion inside the airgap and is defined as [165]:

$$N_u = 2 \quad (41 > T_a) \quad C-4$$

$$N_u = 0.212 \cdot (T_a)^{0.63} \cdot (P_r)^{0.27} \quad (41 < T_a < 100) \quad C-5$$

$$N_u = 0.386 \cdot (T_a)^{0.5} \cdot (P_r)^{0.27} \quad (100 < T_a) \quad C-6$$

Where, P_r is the Prandtl number, (0.694 for air and 73.13 for oil) and T_a is the Taylor number.

Generally, the Taylor number is used to estimate whether the flow pattern inside the airgap of the machine is laminar, vortex and/or turbulent and is calculated using:

$$T_a = \frac{u_1 L_g}{\mu} \cdot \left(\frac{L_g}{r} \right)^{0.5} \quad C-7$$

Where, μ is the dynamic viscosity of the fluid and u_1 is the peripheral speed of the rotor. When the flow inside the airgap has both tangential and axial flow components then the effective velocity is used for the Taylor number calculation as shown below.

$$u_f = \sqrt{u_1^2 + u_{axial}^2} \quad \text{C-8}$$

2. The convection heat transfer coefficient h_{conv_ch} , due to the flow of the coolant in the slot cooling channel is described by:

$$h_{conv_ch} = \frac{N_{u_ch} \cdot k_m}{D_h} \quad \text{C-9}$$

Where D_h is the hydraulic diameter of the slot cooling channel and can be determined using (C-10).

$$D_h = \frac{4 \cdot A_{ch}}{P_{ch}} \quad \text{C-10}$$

Where A_{ch} is the channel cross sectional area and P_{ch} is the perimeter of the channel in contact with fluid.

Assuming the flow is laminar, the Nusselt number N_{u_sch} of the enclosed cooling channel can be calculated from (C-11), where H/W is the channel height to width ratio, L_{ch} is the axial length of the cooling channel and R_e is the Reynold number [166].

$$N_{u_ch} = 7.49 - 17.02 * \left(\frac{H}{W}\right) + 22.43 * \left(\frac{H}{W}\right)^2 - 9.94 * \left(\frac{H}{W}\right)^3 \left[\frac{0.065 * \left(\frac{D_h}{L_{ch}}\right) * R_e * P_r}{1 + 0.04 * \left(\left(\frac{D_h}{L_{ch}}\right) * R_e * P_r\right)^{2/3}} \right] \quad \text{C-11}$$

3. For the end-space region, the convection heat transfer coefficient h_{conv_end} , can be derived assuming turbulent flow forced convection as shown below [107, 163]:

$$h_{conv_end} = k_m \cdot R_{e_end}^{0.8} \cdot P_r^{0.33} \quad \text{C-12}$$

Where R_{e_end} is the Reynold number for the flow in the end-space region.

C.4 Thermal interface gap

The general form of a thermal resistance R_{cont} which describing the contact or interface gap heat transfer is [163]:

$$R_{cont} = \frac{L_{gap}}{k_{air} A} \quad C-13$$

Where, L_{gap} is the interface gap length and k_{air} is the thermal conductivity of air.

The regions where thermal contact resistance is occurred are:

- The interface gap between housing and stator lamination.
- The interface gap between stator lamination and liner.
- The interface gap between rotor sleeve and magnet.
- The interface gap between magnet and rotor shaft.
- The interface gap between housing and endcaps.
- The interface gap between endcaps and bearing.
- The interface gap between bearing and rotor shaft.

The correct estimation of thermal contact resistance is quite difficult to calculate, as it depends upon the contact surfaces, surface roughness, shrink-fit pressures and materials used [163].

C.5 Thermal radiation

The general form of a thermal resistance R_{rad} which describing the radiation heat transfer is [85]:

$$R_{rad} = \frac{1}{h_{rad} A} \quad C-14$$

Where, h_{rad} is the radiation heat transfer coefficient and can be calculated by using the following formula, where ε_m is the surface emissivity, and T_1 and T_2 are the surface temperatures.

$$h_{rad} = 5.67 \cdot 10^{-8} * \varepsilon_m * \frac{[T_1^4 - T_2^4]}{T_1 - T_2} \quad C-15$$

The regions where thermal radiation is occurred are:

- Housing to ambient.
- Endcaps to ambient.

Appendix D. MECHANICAL STRESS MODELLING

The analytical expressions for the mechanical stresses on the PMSM rotor (magnet and sleeve) used in Section 3.10 are detailed in this appendix.

The rotor model is shown in Figure D-1. While, the material characteristics used for the stress analysis are presented in Table D-1. For this specific geometry which represent the rotor as rotating cylinders, there are two sources of stress [118]:

- Stress due to the interference fit (compression force).
- Stress due to rotation (centrifugal force).

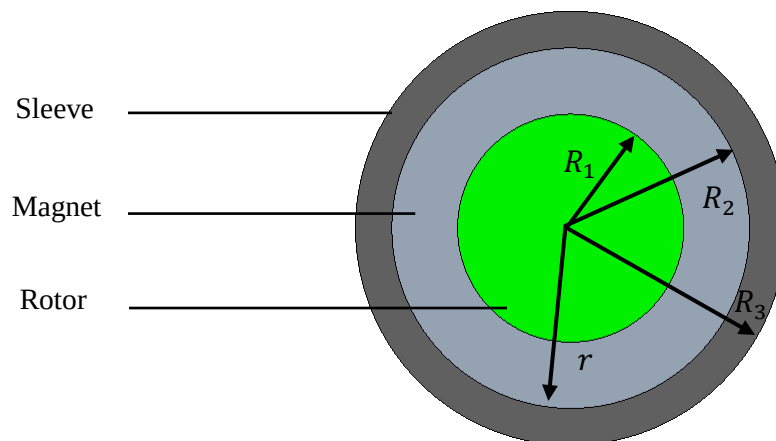


Figure D-1: Rotor model

Table D-1: Material properties

	Titanium Ti-6AL-4V	PM SM ₂ Co ₁₇
Density (kg/m ³)	4400	8300
Young's Modulus (GPa)	110	190
Poisson's ratio	0.31	0.24
Thermal Conductivity (W/m/C)	6.7	10
Permitted stress (MPa)	550	23
Yield /tensile Strength (MPa)	825	35
Electrical Resistivity (Ω.m)	1.78e-6	0.9e-6

D.1 Rotor under standstill condition

The sleeve is normally applied with large interference fit to assure PM retention at high speed. However, when the rotor is in standstill, the high interference fitted sleeve results in high compression force, which stresses the PMs. Also, the equivalent pressure of the magnet body resulted from the interference fitted sleeve may lead to significant stress on the inner surface of the sleeve [119].

Thus, the radial $\sigma_{r,0}$ and circumferential $\sigma_{\theta,0}$ stresses resulting from the interference between magnet and sleeve can be computed from [120] as follows: (the subscript s and m refer to the sleeve and magnet, respectively):

$$\sigma_{rm,0} = \frac{S_0 R_2^2}{R_1^2 - R_2^2} \left(1 - \frac{R_1^2}{r^2} \right) \quad \text{D-1}$$

$$\sigma_{\theta m,0} = \frac{S_0 R_2^2}{R_1^2 - R_2^2} \left(1 + \frac{R_1^2}{r^2} \right) \quad \text{D-2}$$

$$\sigma_{rs,0} = \frac{S_0 R_2^2}{R_3^2 - R_2^2} \left(1 - \frac{R_3^2}{r^2} \right) \quad \text{D-3}$$

$$\sigma_{\theta s,0} = \frac{S_0 R_2^2}{R_3^2 - R_2^2} \left(1 + \frac{R_3^2}{r^2} \right) \quad \text{D-4}$$

The contact pressure S_0 between the sleeve and the magnet can be determined as (D-5) in which δ_0 is the interference fit, E and ν are the Young's Modulus and Poisson's ratio respectively:

$$S_0 = \frac{\delta_0}{\frac{R_2 \{ R_3^2 (1 + \nu_s) + R_2^2 (1 - \nu_s) \}}{E_s (R_3^2 - R_2^2)} + \frac{R_2 \{ R_2^2 (1 - \nu_m) + R_1^2 (1 + \nu_m) \}}{E_m (R_2^2 - R_1^2)}} \quad \text{D-5}$$

The equivalent stress at any point in either the magnets or the sleeve can then be calculated using equation (D-6)

$$\sigma_{equ} = \sqrt{\sigma_{r,0}^2 + \sigma_{\theta,0}^2 - \sigma_{r,0} \sigma_{\theta,0}} \quad \text{D-6}$$

D.2 Rotor under maximum speed condition

When the rotor operates at high speed, the retaining sleeve and the magnet will experience a mechanical stress due to the centrifugal force.

Hence, the radial, $\sigma_{r,\omega}$ and circumferential, $\sigma_{\theta,\omega}$ stresses at a radius r caused by the rotation only are given in equations (D-7) and (D-10) as [120]:

$$\sigma_{rm,\omega} = \frac{3+\nu}{8} \rho \omega^2 \left\{ R_2^2 + R_1^2 - \frac{R_2^2 R_1^2}{r^2} - r^2 \right\} \quad D-7$$

$$\sigma_{\theta m,\omega} = \frac{\rho \omega^2}{8} \left\{ (3+\nu) \left(R_2^2 + R_1^2 + \frac{R_2^2 R_1^2}{r^2} \right) - (1+3\nu)r^2 \right\} \quad D-8$$

$$\sigma_{rs,\omega} = \frac{3+\nu}{8} \rho \omega^2 \left\{ R_3^2 + R_2^2 - \frac{R_3^2 R_2^2}{r^2} - r^2 \right\} \quad D-9$$

$$\sigma_{\theta s,\omega} = \frac{\rho \omega^2}{8} \left\{ (3+\nu) \left(R_3^2 + R_2^2 + \frac{R_3^2 R_2^2}{r^2} \right) - (1+3\nu)r^2 \right\} \quad D-10$$

Where ρ is the density of the material and ω is the speed of rotation.

Finally, the radial σ_r and circumferential σ_θ stresses due to the rotation speed and interference fit can be calculated as:

$$\sigma_{rm} = \sigma_{rm,\omega} + \sigma_{rm,0} \quad D-11$$

$$\sigma_{\theta m} = \sigma_{\theta m,\omega} + \sigma_{\theta m,0} \quad D-12$$

$$\sigma_{rs} = \sigma_{rs,\omega} + \sigma_{rs,0} \quad D-13$$

$$\sigma_{\theta s} = \sigma_{\theta s,\omega} + \sigma_{\theta s,0} \quad D-14$$

Thus, the equivalent (von Mises) stresses of the magnet and sleeve, $\sigma_{equ,m}$ and $\sigma_{equ,s}$, respectively, are described as [119]:

$$\sigma_{equ,m} = \sqrt{\sigma_{rm}^2 + \sigma_{\theta m}^2 - \sigma_{rm}\sigma_{\theta m}} \quad D-15$$

$$\sigma_{equ,s} = \sqrt{\sigma_{rs}^2 + \sigma_{\theta s}^2 - \sigma_{rs}\sigma_{\theta s}} \quad D-16$$

REFERENCES

- [1] G. Qiao, G. Liu, Z. Shi, Y. Wang, S. Ma, and T. C. Lim, "A review of electromechanical actuators for More/All Electric aircraft systems," *Proceedings of the Institution of Mechanical Engineers, Part C: Journal of Mechanical Engineering Science*, p. 0954406217749869, 2017.
- [2] W. Cao, B. C. Mecrow, G. J. Atkinson, J. W. Bennett, and D. J. Atkinson, "Overview of Electric Motor Technologies Used for More Electric Aircraft (MEA)," *IEEE transactions on industrial electronics*, vol. 59, pp. 3523-3531, 2012.
- [3] M. Villani, M. Tursini, G. Fabri, and L. Castellini, "Electromechanical actuator for helicopter rotor damper application," *IEEE Transactions on Industry Applications*, vol. 50, pp. 1007-1014, 2014.
- [4] M. Khatre, "Reduced DC-link capacitor drives for more-electric aircraft applications," 2010.
- [5] N. Alle, S. S. Hiremath, S. Makaram, K. Subramaniam, and A. Talukdar, "Review on electro hydrostatic actuator for flight control," *International Journal of Fluid Power*, vol. 17, pp. 125-145, 2016.
- [6] J. Zhang, C. Qun, and X. Bing, "Analysis of the cylinder block tilting inertia moment and its effect on the performance of high-speed electro-hydrostatic actuator pumps of aircraft," *Chinese Journal of Aeronautics*, vol. 31, pp. 169-177, 2018.
- [7] G. J. Atkinson, B. C. Mecrow, A. G. Jack, D. J. Atkinson, P. Sangha, and M. Benarous, "The Analysis of Losses in High-Power Fault-Tolerant Machines for Aerospace Applications," *IEEE Transactions on Industry Applications*, vol. 42, pp. 1162-1170, 2006.
- [8] B. C. Mecrow, A. G. Jack, D. J. Atkinson, S. R. Green, G. J. Atkinson, A. King, *et al.*, "Design and testing of a four-phase fault-tolerant permanent-magnet machine for an engine fuel pump," *IEEE Transactions on Energy Conversion*, vol. 19, pp. 671-678, 2004.
- [9] X. Huang, A. Goodman, C. Gerada, Y. Fang, and Q. Lu, "Design of a five-phase brushless DC motor for a safety critical aerospace application," *IEEE transactions on industrial electronics*, vol. 59, pp. 3532-3541, 2012.
- [10] C. Gerada and K. J. Bradley, "Integrated PM Machine Design for an Aircraft EMA," *IEEE Transactions on Industrial Electronics*, vol. 55, pp. 3300-3306, 2008.
- [11] M. Villani, M. Tursini, G. Fabri, and L. Castellini, "Multi-phase fault tolerant drives for aircraft applications," *Electrical Systems for Aircraft, Railway and Ship Propulsion (ESARS 2010)*, 2010.
- [12] J. W. Bennett, B. C. Mecrow, A. G. Jack, and D. J. Atkinson, "A prototype electrical actuator for aircraft flaps," *IEEE Transactions on Industry Applications*, vol. 46, pp. 915-921, 2010.
- [13] S. Elfgen, S. Steentjes, S. Böhmer, D. Franck, and K. Hameyer, "Influences of Material Degradation Due to Laser Cutting on the Operating Behavior of PMSM Using a Continuous Local Material Model," *IEEE Transactions on Industry Applications*, vol. 53, pp. 1978-1984, 2017.

- [14] A. Krings, M. Cossale, J. Soulard, A. Boglietti, and A. Cavagnino, "Manufacturing influence on the magnetic properties and iron losses in cobalt-iron stator cores for electrical machines," in *2014 IEEE Energy Conversion Congress and Exposition (ECCE)*, 2014, pp. 5595-5601.
- [15] A. Krings, S. Nategh, O. Wallmark, and J. Soulard, "Influence of the Welding Process on the Performance of Slotless PM Motors With SiFe and NiFe Stator Laminations," *IEEE Transactions on Industry Applications*, vol. 50, pp. 296-306, 2014.
- [16] D. G. Ahn, M. H. Yoon, S. I. Kim, and J. P. Hong, "Welding loss modeling and evaluation of electric motors using laminated cores," in *2016 IEEE Conference on Electromagnetic Field Computation (CEFC)*, 2016, pp. 1-1.
- [17] M. Gamba, G. Pellegrino, A. Cavagnino, Z. Gmyrek, and M. Lefik, "Impact of Rotor End Effects on FEM-based Flux Mapping of Synchronous Reluctance Motors," *IEEE Transactions on Industry Applications*, 2018.
- [18] J. H. Potgieter and M. J. Kamper, "Calculation methods and effects of end-winding inductance and permanent-magnet end flux on performance prediction of nonoverlap winding permanent-magnet machines," *IEEE Transactions on Industry Applications*, vol. 50, pp. 2458-2466, 2014.
- [19] J. Fan, D. Ling, Z. Tang, and Z. Pei, "Study on nonlinear PID control for electro-hydrostatic actuator," in *Guidance, Navigation and Control Conference (CGNCC), 2014 IEEE Chinese*, 2014, pp. 1317-1320.
- [20] P. A. Robson, K. J. Bradley, P. Wheeler, J. Clare, L. d. Lillo, C. Gerada, *et al.*, "The impact of matrix converter technology on motor design for an integrated flight control surface actuation system," in *Electric Machines and Drives Conference, 2003. IEMDC'03. IEEE International*, 2003, pp. 1321-1327 vol.2.
- [21] M. Hartmann, "Ultra-compact and ultra-efficient three-phase PWM rectifier systems for more electric aircraft," 2011.
- [22] M. A. Ismail, E. Balaban, and H. Spangenberg, "Fault detection and classification for flight control electromechanical actuators," in *Aerospace Conference, 2016 IEEE*, 2016, pp. 1-10.
- [23] J.-C. Derrien and S. D. Sécurité, "Electromechanical actuator (EMA) advanced technologies for flight controls," in *International Congress of the Aeronautical Sciences*, 2012, pp. 1-10.
- [24] J. Claeysse and L. NKERP, "New actuators for aircraft, space and military applications," *12th International Conference on New Actuators, Bremen, Germany*, pp. 324-330, 2010.
- [25] P. Ponomarev, M. Polikarpova, O. Heinikainen, and J. Pyrhönen, "Design of integrated electro-hydraulic power unit for hybrid mobile working machines," in *Proceedings of the 2011 14th European Conference on Power Electronics and Applications*, 2011, pp. 1-10.
- [26] D. Van Den Bossche, "The A380 flight control electrohydrostatic actuators, achievements and lessons learnt," in *25th international congress of the aeronautical sciences*, 2006, pp. 1-8.
- [27] D. J. Powell, K. Atallah, and G. Jewell, "Thermal Modeling of Flooded Rotor Electrical Machines for Electro-Hydrostatic Actuators," in *Electric Machines & Drives Conference, 2007. IEMDC'07. IEEE International*, 2007, pp. 1632-1637.

- [28] Z. Xu, A. Al-Timimy, M. Degano, P. Giangrande, G. L. Calzo, H. Zhang, *et al.*, "Thermal management of a permanent magnet motor for an directly coupled pump," in *2016 XXII International Conference on Electrical Machines (ICEM)*, 2016, pp. 2738-2744.
- [29] G. Hong, T. Wei, X. Ding, and C. Duan, "Multi-Objective Optimal Design of Electro-Hydrostatic Actuator Driving Motors for Low Temperature Rise and High Power Weight Ratio," *Energies*, vol. 11, p. 1173, 2018.
- [30] K. Atallah, F. Caparrelli, C. Bingham, N. Schofield, D. Howe, P. Mellor, *et al.*, "Permanent magnet brushless drives for aircraft flight control surface actuation," in *Electrical Machines and Systems for the More Electric Aircraft (Ref. No. 1999/180), IEE Colloquium on*, 1999, pp. 8/1-8/5.
- [31] T. Minav, C. Bonato, P. Sainio, and M. Pietola, "Efficiency of direct driven hydraulic drive for non-road mobile working machines," in *Electrical Machines (ICEM), 2014 International Conference on*, 2014, pp. 2431-2435.
- [32] T. A. Minav, J. J. Pyrhönen, and L. I. Laurila, "Permanent magnet synchronous machine sizing: effect on the energy efficiency of an electro-hydraulic forklift," *IEEE Transactions on Industrial Electronics*, vol. 59, p. 2466, 2012.
- [33] M. Jean-Charles and F. Jian, "Review on signal-by-wire and power-by-wire actuation for more electric aircraft," *Chinese Journal of Aeronautics*, vol. 30, pp. 857-870, 2017.
- [34] L. Zhihui, Y. SHANG, J. Zongxia, L. Yan, W. Shuai, and L. Xiaobin, "Analysis of the dynamic performance of an electro-hydrostatic actuator and improvement methods," *Chinese Journal of Aeronautics*, 2018.
- [35] C. Cossar, L. Kelly, T. Miller, C. Whitley, C. Maxwell, and D. Moorhouse, "The design of a switched reluctance drive for aircraft flight control surface actuation," 1999.
- [36] R. Krishnan, D. Blanding, A. Bhanot, A. Staley, and N. Lobo, "High reliability SRM drive system for aerospace applications," in *Industrial Electronics Society, 2003. IECON'03. The 29th Annual Conference of the IEEE*, 2003, pp. 1110-1115.
- [37] P. Churn, C. Maxwell, N. Schofield, D. Howe, and D. Powell, "Electro-hydraulic actuation of primary flight control surfaces," 1998.
- [38] A. Boglietti, A. Cavagnino, A. Tenconi, S. Vaschetto, and P. di Torino, "The safety critical electric machines and drives in the more electric aircraft: A survey," in *2009 35th Annual Conference of IEEE Industrial Electronics*, 2009, pp. 2587-2594.
- [39] P. E. Kakosimos, A. G. Sarigiannidis, M. E. Beniakar, A. G. Kladas, and C. Gerada, "Induction Motors Versus Permanent-Magnet Actuators for Aerospace Applications," *IEEE Transactions on Industrial Electronics*, vol. 61, pp. 4315-4325, 2014.
- [40] A. M. El-Refaie and T. M. Jahns, "Comparison of synchronous PM machine types for wide constant-power speed range operation," *COMPEL-The international journal for computation and mathematics in electrical and electronic engineering*, vol. 27, pp. 967-984, 2008.
- [41] A. Vagati, G. Pellegrino, and P. Guglielmi, "Comparison between SPM and IPM motor drives for EV application," in *The XIX International Conference on Electrical Machines - ICEM 2010*, 2010, pp. 1-6.

- [42] F. Charih, F. Dubas, C. Espanet, and D. Chamagne, "Performances comparison of PM machines with different rotor topologies and similar slot and pole numbers," in *Power Electronics, Electrical Drives, Automation and Motion (SPEEDAM), 2012 International Symposium on*, 2012, pp. 56-59.
- [43] A. R. Munoz, F. Liang, and M. W. Degner, "Evaluation of Interior PM and Surface PM Synchronous machines with distributed and concentrated windings," in *Industrial Electronics, 2008. IECON 2008. 34th Annual Conference of IEEE*, 2008, pp. 1189-1193.
- [44] C. Gerada, K. Bradley, C. Whitley, and G. Towers, "High Torque Density PM Machines for High Performance Operation," in *IECON 2007 - 33rd Annual Conference of the IEEE Industrial Electronics Society*, 2007, pp. 210-215.
- [45] J. B. Bartolo, H. Zhang, D. Gerada, L. D. Lillo, and C. Gerada, "High speed electrical generators, application, materials and design," in *2013 IEEE Workshop on Electrical Machines Design, Control and Diagnosis (WEMDCD)*, 2013, pp. 47-59.
- [46] P. Pillay and R. Krishnan, "Application characteristics of permanent magnet synchronous and brushless DC motors for servo drives," *IEEE Transactions on industry applications*, vol. 27, pp. 986-996, 1991.
- [47] P. Arumugam, Z. Xu, A. L. Rocca, G. Vakil, M. Dickinson, E. Amankwah, *et al.*, "High-Speed Solid Rotor Permanent Magnet Machines: Concept and Design," *IEEE Transactions on Transportation Electrification*, vol. 2, pp. 391-400, 2016.
- [48] A. Al-Timimy, M. Degano, P. Giangrande, G. L. Calzo, Z. X. M. Galea, C. Gerada, *et al.*, "Design and optimization of a high power density machine for flooded industrial pump," in *Electrical Machines (ICEM), 2016 XXII International Conference on*, 2016, pp. 1480-1486.
- [49] A. Al-Timimy, P. Giangrande, M. Degano, M. Galea, and C. Gerada, "Investigation of AC Copper and Iron Losses in High-Speed High-Power Density PMSM," in *2018 XIII International Conference on Electrical Machines (ICEM)*, 2018, pp. 263-269.
- [50] F. Jiancheng, L. Xiquan, B. Han, and K. Wang, "Analysis of circulating current loss for high-speed permanent magnet motor," *IEEE Transactions on Magnetics*, vol. 51, pp. 1-13, 2015.
- [51] M. van der Geest, H. Polinder, J. A. Ferreira, and D. Zeilstra, "Stator winding proximity loss reduction techniques in high speed electrical machines," in *Electric Machines & Drives Conference (IEMDC), 2013 IEEE International*, 2013, pp. 340-346.
- [52] S. Iwasaki, R. Deodhar, Y. Liu, A. Pride, Z. Zhu, and J. Bremner, "Influence of PWM on the proximity loss in permanent magnet brushless AC machines," in *Industry Applications Society Annual Meeting, 2008. IAS'08. IEEE*, 2008, pp. 1-8.
- [53] Y. Amara, P. Reghem, and G. Barakat, "Analytical prediction of eddy-current loss in armature windings of permanent magnet brushless AC machines," *IEEE Transactions on Magnetics*, vol. 46, pp. 3481-3484, 2010.
- [54] A. Thomas, Z. Zhu, and G. Jewell, "Proximity loss study in high speed flux-switching permanent magnet machine," *IEEE Transactions on Magnetics*, vol. 45, pp. 4748-4751, 2009.

- [55] P. Reddy, Z. Zhu, S.-H. Han, and T. Jahns, "Strand-level proximity losses in PM machines designed for high-speed operation," in *Electrical Machines, 2008. ICEM 2008. 18th International Conference on*, 2008, pp. 1-6.
- [56] D. Zarko, D. Ban, and T. A. Lipo, "Analytical calculation of magnetic field distribution in the slotted air gap of a surface permanent-magnet motor using complex relative air-gap permeance," *IEEE Transactions on Magnetics*, vol. 42, pp. 1828-1837, 2006.
- [57] P. B. Reddy, T. Jahns, and T. P. Bohn, "Modeling and analysis of proximity losses in high-speed surface permanent magnet machines with concentrated windings," in *Energy Conversion Congress and Exposition (ECCE), 2010 IEEE*, 2010, pp. 996-1003.
- [58] L. Wu and Z. Zhu, "Analytical investigation of open-circuit eddy current loss in windings of PM machines," in *Electrical Machines (ICEM), 2012 XXth International Conference on*, 2012, pp. 2759-2765.
- [59] L. Wu, Z. Zhu, D. Staton, M. Popescu, and D. Hawkins, "Analytical model of eddy current loss in windings of permanent-magnet machines accounting for load," *IEEE Transactions on Magnetics*, vol. 48, pp. 2138-2151, 2012.
- [60] R. Wrobel, A. Mlot, and P. H. Mellor, "Contribution of end-winding proximity losses to temperature variation in electromagnetic devices," *IEEE Transactions on Industrial Electronics*, vol. 59, pp. 848-857, 2012.
- [61] A. Mlot, M. Lukaniszyn, and M. Korkosz, "Influence of an end-winding size on proximity losses in a high-speed PM synchronous motor," in *Selected Problems of Electrical Engineering and Electronics (WZEE), 2015*, 2015, pp. 1-6.
- [62] D. Bauer, P. Mamuschkin, H.-C. Reuss, and E. Nolle, "Influence of parallel wire placement on the AC copper losses in electrical machines," in *Electric Machines & Drives Conference (IEMDC), 2015 IEEE International*, 2015, pp. 1247-1253.
- [63] M. Cossale, A. Krings, J. Souldard, A. Boglietti, and A. Cavagnino, "Practical Investigations on Cobalt–Iron Laminations for Electrical Machines," *IEEE Transactions on Industry Applications*, vol. 51, pp. 2933-2939, 2015.
- [64] P.-K. Lee, K.-C. Kuo, C.-J. Wu, Z.-T. Wong, and J.-Y. Yen, "Prediction of iron losses using the modified Steinmetz equation under the sinusoidal waveform," in *Control Conference (ASCC), 2011 8th Asian*, 2011, pp. 579-584.
- [65] Z. Zhu, S. Xue, W. Chu, J. Feng, S. Guo, Z. Chen, *et al.*, "Evaluation of iron loss models in electrical machines," *IEEE Transactions on Industry Applications*, 2018.
- [66] C. Huynh, L. Zheng, and D. Acharya, "Losses in high speed permanent magnet machines used in microturbine applications," *Journal of Engineering for Gas Turbines and Power*, vol. 131, p. 022301, 2009.
- [67] V. Leivsdottir, "Investigation of Loss Calculation Methods for PMSMs and Implementation of Loss Functionality on a Developed FEM Model," NTNU, 2016.
- [68] F. Magnussen, "On design and analysis of synchronous permanent magnet machines for field-weakening operation in hybrid electric vehicles," *Elektrotekniska system*, 2004.
- [69] Z. Zhu, K. Ng, N. Schofield, and D. Howe, "Improved analytical modelling of rotor eddy current loss in brushless machines equipped with surface-mounted permanent magnets," *IEE Proceedings-Electric Power Applications*, vol. 151, pp. 641-650, 2004.

- [70] D. Ishak, Z. Zhu, and D. Howe, "Eddy-current loss in the rotor magnets of permanent-magnet brushless machines having a fractional number of slots per pole," *IEEE Transactions on magnetics*, vol. 41, pp. 2462-2469, 2005.
- [71] L. Wu, Z. Zhu, D. Staton, M. Popescu, and D. Hawkins, "Analytical modeling and analysis of open-circuit magnet loss in surface-mounted permanent-magnet machines," *IEEE transactions on magnetics*, vol. 48, pp. 1234-1247, 2012.
- [72] L. Wu, Z. Zhu, D. Staton, M. Popescu, and D. Hawkins, "Analytical model for predicting magnet loss of surface-mounted permanent magnet machines accounting for slotting effect and load," *IEEE Transactions on Magnetics*, vol. 48, pp. 107-117, 2012.
- [73] K. Atallah, D. Howe, P. H. Mellor, and D. A. Stone, "Rotor loss in permanent-magnet brushless AC machines," *IEEE Transactions on Industry Applications*, vol. 36, pp. 1612-1618, 2000.
- [74] H. V. Xuan, D. Lahaye, H. Polinder, and J. A. Ferreira, "Influence of stator slotting on the performance of permanent-magnet machines with concentrated windings," *IEEE Transactions on Magnetics*, vol. 49, pp. 929-938, 2013.
- [75] N. Zhao, Z. Zhu, and W. Liu, "Rotor eddy current loss calculation and thermal analysis of permanent magnet motor and generator," *IEEE Transactions on Magnetics*, vol. 47, pp. 4199-4202, 2011.
- [76] L. Wu, R. Qu, and D. Li, "Reduction of rotor eddy-current losses for surface PM machines with fractional slot concentrated windings and retaining sleeve," *IEEE Transactions on Magnetics*, vol. 50, pp. 1-4, 2014.
- [77] G. Choi and T. M. Jahns, "Reduction of eddy-current losses in fractional-slot concentrated-winding synchronous PM machines," *IEEE Transactions on Magnetics*, vol. 52, pp. 1-4, 2016.
- [78] K. Yamazaki, Y. Kanou, Y. Fukushima, S. Ohki, A. Nezu, T. Ikemi, *et al.*, "Reduction of magnet eddy-current loss in interior permanent-magnet motors with concentrated windings," *IEEE Transactions on Industry Applications*, vol. 46, pp. 2434-2441, 2010.
- [79] K. Yamazaki, Y. Fukushima, and M. Sato, "Loss analysis of permanent magnet motors with concentrated windings-variation of magnet eddy current loss due to stator and rotor shapes," in *Industry Applications Society Annual Meeting, 2008. IAS'08. IEEE*, 2008, pp. 1-8.
- [80] M.-M. Koo, J.-Y. Choi, K. Hong, and K. Lee, "Comparative analysis of eddy-current loss in permanent magnet synchronous machine considering PM shape and skew effect using 3-D FEA," *IEEE Transactions on Magnetics*, vol. 51, pp. 1-4, 2015.
- [81] J. Li, Y. Xu, J. Zou, Q. Wang, and W. Liang, "Analysis and reduction of magnet loss by deepening magnets in interior permanent-magnet machines with a pole/slot ratio of 2/3," *IEEE Transactions on Magnetics*, vol. 51, pp. 1-4, 2015.
- [82] A. Krings, "Iron losses in electrical machines-Influence of material properties, manufacturing processes, and inverter operation," KTH Royal Institute of Technology, 2014.
- [83] R. J. Bruckner, "Windage Power Loss in Gas Foil Bearings and the Rotor-Stator Clearance of High Speed Generators Operating in High Pressure Environments," in *ASME Turbo Expo 2009: Power for Land, Sea, and Air*, 2009, pp. 263-270.

- [84] J. Saari, *Thermal analysis of high-speed induction machines*: Helsinki University of Technology, 1998.
- [85] J. Nerg, M. Rilla, and J. Pyrhonen, "Thermal Analysis of Radial-Flux Electrical Machines With a High Power Density," *IEEE Transactions on Industrial Electronics*, vol. 55, pp. 3543-3554, 2008.
- [86] A. Nachouane, A. Abdelli, G. Friedrich, and S. Vivier, "Estimation of windage losses inside very narrow air gaps of high speed electrical machines without an internal ventilation using CFD methods," in *Electrical Machines (ICEM), 2016 XXII International Conference on*, 2016, pp. 2704-2710.
- [87] P. Sangha, T. Sawata, J. Yon, and P. Mellor, "Assessment of fluid drag loss in a flooded rotor electro-hydrostatic actuator motor," in *Electric Machines & Drives Conference (IEMDC), 2015 IEEE International*, 2015, pp. 139-142.
- [88] D. Zhang, F. Wang, and X. Kong, "Air friction loss calculation of high speed permanent magnet machines," in *Electrical Machines and Systems, 2008. ICEMS 2008. International Conference on*, 2008, pp. 320-323.
- [89] C. Zwyssig, S. Round, and J. W. Kolar, "Analytical and experimental investigation of a low torque, ultra-high speed drive system," in *Industry Applications Conference, 2006. 41st IAS Annual Meeting. Conference Record of the 2006 IEEE*, 2006, pp. 1507-1513.
- [90] M. N. B. Othman, "Design and evaluation of high power density brushless DC permanent magnet machines," University of Nottingham, 2012.
- [91] J. F. Gieras, *Permanent magnet motor technology: design and applications*: CRC press, 2009.
- [92] J. Pyrhonen, T. Jokinen, and V. Hrabovcova, *Design of rotating electrical machines*: John Wiley & Sons, 2013.
- [93] S.K.F. (03-June-2016). *Bearing friction and power losses*. Available: <http://www.skf.com/group/products/bearings-units-housings/principles/bearing-selection-process/operating-temperature-and-speed/friction-powerloss-startingtorque/index.html>
- [94] S. o. Kwon, S. i. Kim, P. Zhang, and J. p. Hong, "Performance comparison of IPMSM with distributed and concentrated windings," in *Conference Record of the 2006 IEEE Industry Applications Conference Forty-First IAS Annual Meeting*, 2006, pp. 1984-1988.
- [95] J. Gao, Y. Yu, and S. Huang, "Winding layers and slot/pole combination in fractional slot/pole PMSM—Effects on motor performance," in *2009 International Conference on Electrical Machines and Systems*, 2009, pp. 1-4.
- [96] H. Dogan, F. Wurtz, A. Foggia, and L. Garbuio, "Analysis of slot-pole combination of fractional-slots PMSM for embedded applications," in *International Aegean Conference on Electrical Machines and Power Electronics and Electromotion, Joint Conference*, 2011, pp. 611-615.
- [97] P. Ponomarev, P. Lindh, and J. Pyrhönen, "Effect of slot-and-pole combination on the leakage inductance and the performance of tooth-coil permanent-magnet synchronous machines," *IEEE Transactions on Industrial Electronics*, vol. 60, pp. 4310-4317, 2013.
- [98] Z.-Q. Zhu, "Fractional slot permanent magnet brushless machines and drives for electric and hybrid propulsion systems," *COMPEL-The international journal for computation and mathematics in electrical and electronic engineering*, vol. 30, pp. 9-31, 2011.

- [99] P. Ponomarev, Y. Alexandrova, I. Petrov, P. Lindh, E. Lomonova, and J. Pyrhönen, "Inductance calculation of tooth-coil permanent-magnet synchronous machines," *IEEE Transactions on Industrial Electronics*, vol. 61, pp. 5966-5973, 2014.
- [100] A. Krings, A. Boglietti, A. Cavagnino, and S. Sprague, "Soft Magnetic Material Status and Trends in Electric Machines," *IEEE Transactions on Industrial Electronics*, vol. 64, pp. 2405-2414, 2017.
- [101] M. Galea, Z. Xu, C. Tighe, T. Hamiti, C. Gerada, and S. Pickering, "Development of an aircraft wheel actuator for green taxiing," in *2014 International Conference on Electrical Machines (ICEM)*, 2014, pp. 2492-2498.
- [102] *Material properties data for M270-35A*. Available: https://cogent-power.com/cms-data/downloads/m270-35a_1.pdf
- [103] A. M. Mohammed, "Considerations on the use of solid material for high performance, tubular linear motors," University of Nottingham, 2017.
- [104] A.T.M. Arnold Magnetics Technologies, *Technical Datasheet*. Available: <https://www.arnoldmagnetics.com/wp-content/uploads/2017/10/Recoma-Combined-160301.pdf>
- [105] H. Fang, R. Qu, J. Li, P. Zheng, and X. Fan, "Rotor Design for High-Speed High-Power Permanent-Magnet Synchronous Machines," *IEEE Transactions on Industry Applications*, vol. 53, pp. 3411-3419, 2017.
- [106] S. Ahmed and H. A. Toliyat, "Coupled field analysis needs in the design of submersible electric motors," in *Electric Ship Technologies Symposium, 2007. ESTS'07. IEEE*, 2007, pp. 231-237.
- [107] B. Zhang, R. Qu, J. Wang, W. Xu, X. Fan, and Y. Chen, "Thermal Model of Totally Enclosed Water-Cooled Permanent-Magnet Synchronous Machines for Electric Vehicle Application," *IEEE Transactions on Industry Applications*, vol. 51, pp. 3020-3029, 2015.
- [108] C. Kral, A. Haumer, and T. Bauml, "Thermal Model and Behavior of a Totally-Enclosed-Water-Cooled Squirrel-Cage Induction Machine for Traction Applications," *IEEE Transactions on Industrial Electronics*, vol. 55, pp. 3555-3565, 2008.
- [109] A. Boglietti, A. Cavagnino, M. Lazzari, and M. Pastorelli, "A simplified thermal model for variable-speed self-cooled industrial induction motor," *IEEE Transactions on Industry Applications*, vol. 39, pp. 945-952, 2003.
- [110] R. Wrobel and P. H. Mellor, "A General Cuboidal Element for Three-Dimensional Thermal Modelling," *IEEE Transactions on Magnetics*, vol. 46, pp. 3197-3200, 2010.
- [111] A. L. Rocca, S. J. Pickering, C. Eastwick, and C. Gerada, "Enhanced cooling for an electric starter-generator for aerospace application," in *7th IET International Conference on Power Electronics, Machines and Drives (PEMD 2014)*, 2014, pp. 1-7.
- [112] Z. Xu, C. Tighe, M. Galea, T. Hamiti, C. Gerada, and S. J. Pickering, "Thermal design of a permanent magnetic motor for direct drive wheel actuator," in *2014 International Conference on Electrical Machines (ICEM)*, 2014, pp. 2186-2192.
- [113] Z. Xu, A. La Rocca, P. Arumugam, S. Pickering, C. Gerada, S. Bozhko, *et al.*, "A semi-flooded cooling for a high speed machine: Concept, design and practice

- of an oil sleeve," in *Industrial Electronics Society, IECON 2017-43rd Annual Conference of the IEEE*, 2017, pp. 8557-8562.
- [114] T. Sun, J.-M. Kim, G.-H. Lee, J.-P. Hong, and M.-R. Choi, "Effect of pole and slot combination on noise and vibration in permanent magnet synchronous motor," *IEEE Transactions on magnetics*, vol. 47, pp. 1038-1041, 2011.
 - [115] N. Uzhegov, E. Kurvinen, J. Nerg, J. Pyrhönen, J. T. Sopanen, and S. Shirinskii, "Multidisciplinary design process of a 6-slot 2-pole high-speed permanent-magnet synchronous machine," *IEEE Transactions on Industrial Electronics*, vol. 63, pp. 784-795, 2016.
 - [116] M. Galea, C. Gerada, T. Raminosa, and P. Wheeler, "Design of a high force density tubular permanent magnet motor," in *Electrical Machines (ICEM), 2010 XIX International Conference on*, 2010, pp. 1-6.
 - [117] F. Zhang, G. Du, T. Wang, G. Liu, and W. Cao, "Rotor retaining sleeve design for a 1.12-MW high-speed PM machine," *IEEE Transactions on Industry Applications*, vol. 51, pp. 3675-3685, 2015.
 - [118] J.-h. Ahn, C. Han, C.-w. Kim, and J.-y. Choi, "Rotor Design of High-Speed Permanent Magnet Synchronous Motors Considering Rotor Magnet and Sleeve Materials," *IEEE Transactions on Applied Superconductivity*, vol. 28, pp. 1-4, 2018.
 - [119] A. Damiano, A. Floris, G. Fois, M. Porru, and A. Serpi, "Modelling and design of PM retention sleeves for High-Speed PM Synchronous Machines," in *Electric Drives Production Conference (EDPC), 2016 6th International*, 2016, pp. 118-125.
 - [120] S. M. Barrans, M. M. Al-Ani, and J. Carter, "Mechanical design of rotors for permanent magnet high-speed electric motors for turbocharger applications," *IET Electrical Systems in Transportation*, vol. 7, pp. 278-286, 2017.
 - [121] R. D. Bremner, "Rapid optimization of interior permanent magnet (IPM) machines using the response surface method and dimensionless parameters," 2010.
 - [122] M. Galea, L. Papini, H. Zhang, C. Gerada, and T. Hamiti, "Demagnetization analysis for Halbach array configurations in electrical machines," *IEEE Transactions on Magnetism*, vol. 51, pp. 1-9, 2015.
 - [123] G. Qi, J. Chen, Z. Zhu, D. Howe, L. Zhou, and C. Gu, "Influence of skew and cross-coupling on flux-weakening performance of permanent-magnet brushless AC machines," *IEEE Transactions on magnetics*, vol. 45, pp. 2110-2117, 2009.
 - [124] A. Al-Timimy, P. Giangrande, M. Degano, Z. Xu, M. Galea, C. Gerada, *et al.*, "Design and losses analysis of a high power density machine for flooded pump applications," *IEEE Transactions on Industry Applications*, 2018.
 - [125] A. Al-Timimy, M. Degano, Z. Xu, G. L. Calzo, P. Giangrande, M. Galea, *et al.*, "Trade-off analysis and design of a high power density PM machine for flooded industrial pump," in *Industrial Electronics Society, IECON 2016-42nd Annual Conference of the IEEE*, 2016, pp. 1749-1754.
 - [126] A. F. 31. (15-October-2015). *Aeroshell Hydraulic Fluids, About Oil Properties*. Available: https://www.shell.com/business-customers/aviation/aeroshell/knowledge-centre/the-aeroshell-book/_jcr_content/par/textimage_1433441235.stream/1445042942924/2dc948cd5b705e5aba8766802e749f9aa44c0a59c9d156505f4fd5bff842b968/aeroshell-book-6hydraulics.pdf

- [127] W. Soong, "Inductance measurements for synchronous machines," *Power Engineering Briefing Note Series*, vol. 2, 2008.
- [128] S. Mukundan, H. Dhulipati, K. L. V. Iyer, N. C. Kar, and K. Mukherjee, "Comparison of inductance determination methods of PMSMs for EV application," in *Electrical and Computer Engineering (CCECE), 2017 IEEE 30th Canadian Conference on*, 2017, pp. 1-4.
- [129] D. Steinert, I. Kovacevic-Badstübner, T. Nussbaumer, and J. W. Kolar, "Loss investigation of slotless bearingless disk drives," in *Energy Conversion Congress and Exposition (ECCE), 2015 IEEE*, 2015, pp. 4418-4424.
- [130] A. Al-Timimy, M. Al-Ani, M. Degano, P. Giangrande, C. Gerada, and M. Galea, "Influence of rotor endcaps on the electromagnetic performance of high-speed PM machine," *IET Electric Power Applications*, 2018.
- [131] M. Cossale, A. Krings, J. Souldard, A. Boglietti, and A. Cavagnino, "Practical investigations on cobalt-iron laminations for electrical machines," *IEEE Transactions on Industry Applications*, vol. 51, pp. 2933-2939, 2015.
- [132] P. Goes, E. Hoferlin, and M. De Wulf, "Calculating iron losses taking into account effects of manufacturing processes," in *Proc. of COMSOL Conference*, 2008.
- [133] Z. Gmyrek, A. Cavagnino, and L. Ferraris, "Estimation of the magnetic properties of the damaged area resulting from the punching process: Experimental research and FEM modeling," *IEEE Transactions on Industry Applications*, vol. 49, pp. 2069-2077, 2013.
- [134] M. S. Lim, J. H. Kim, and J. P. Hong, "Experimental Characterization of the Slinky-Laminated Core and Iron Loss Analysis of Electrical Machine," *IEEE Transactions on Magnetics*, vol. 51, pp. 1-4, 2015.
- [135] N. Fernando, G. Vakil, P. Arumugam, E. Amankwah, C. Gerada, and S. Bozhko, "Impact of Soft Magnetic Material on Design of High-Speed Permanent-Magnet Machines," *IEEE Transactions on Industrial Electronics*, vol. 64, pp. 2415-2423, 2017.
- [136] M. Dems and K. Komeza, "The Influence of Electrical Sheet on the Core Losses at No-Load and Full-Load of Small Power Induction Motors," *IEEE Transactions on Industrial Electronics*, vol. 64, pp. 2433-2442, 2017.
- [137] M. Dems and K. Komeza, "Performance Characteristics of a High-Speed Energy-Saving Induction Motor With an Amorphous Stator Core," *IEEE Transactions on Industrial Electronics*, vol. 61, pp. 3046-3055, 2014.
- [138] A. Krings, S. Nategh, A. Stening, H. Grop, O. Wallmark, and J. Souldard, "Measurement and modeling of iron losses in electrical machines," in *5th International Conference Magnetism and Metallurgy WMM'12, June 20th to 22nd, 2012, Ghent, Belgium*, 2012, pp. 101-119.
- [139] M. Galea, C. Gerada, T. Raminosoa, and P. Wheeler, "A Thermal Improvement Technique for the Phase Windings of Electrical Machines," *IEEE Transactions on Industry Applications*, vol. 48, pp. 79-87, 2012.
- [140] C. Sciascera, P. Giangrande, L. Papini, C. Gerada, and M. Galea, "Analytical thermal model for fast stator winding temperature prediction," *IEEE Transactions on Industrial Electronics*, vol. 64, pp. 6116-6126, 2017.
- [141] A. M. Mohammed, T. Cox, M. Galea, and C. Gerada, "A New Method for Determining the Magnetic Properties of Solid Materials Employed in

- Unconventional Magnetic Circuits," *IEEE Transactions on Industrial Electronics*, vol. 64, pp. 2468-2475, 2017.
- [142] A. Krings and J. Soulard, "Experimental characterization of magnetic materials for electrical machine applications," in *2015 IEEE Workshop on Electrical Machines Design, Control and Diagnosis (WEMDCD)*, 2015, pp. 85-89.
- [143] A. J. Clerc and A. Muetze, "Measurement of Stator Core Magnetic Degradation During the Manufacturing Process," *IEEE Transactions on Industry Applications*, vol. 48, pp. 1344-1352, 2012.
- [144] N. Volbers and J. Gerster, "High saturation high strength iron-cobalt alloy for electrical machines," *Proceedings of the INDUCTICA, CWIEME Berlin*, 2012.
- [145] B. Measurements, "Manual for software operation of the MPG 200 D," 2014.
- [146] A. Saleem, N. Alatawneh, R. R. Chromik, and D. A. Lowther, "Effect of Shear Cutting on Microstructure and Magnetic Properties of Non-Oriented Electrical Steel," *IEEE Transactions on Magnetics*, vol. 52, pp. 1-4, 2016.
- [147] A. Al-Timimy, G. Vakil, M. Degano, P. Giangrande, C. Gerada, and M. Galea, "Considerations on the effects that core material machining has on an electrical machine's performance," *IEEE Transactions on Energy Conversion*, 2018.
- [148] M. Sanada, S. Morimoto, and Y. Takeda, "Interior permanent magnet linear synchronous motor for high-performance drives," *IEEE transactions on industry applications*, vol. 33, pp. 966-972, 1997.
- [149] H. Polinder, J. G. Sloopweg, M. J. Hoeijmakers, and J. C. Compter, "Modeling of a linear PM machine including magnetic saturation and end effects: Maximum force-to-current ratio," *IEEE Transactions on Industry Applications*, vol. 39, pp. 1681-1688, 2003.
- [150] W. Hua, M. Cheng, X. Zhu, and J. Zhang, "Investigation of end-effect in brushless machines having magnets in the stator with doubly salient structure," in *Magnetics Conference, 2006. INTERMAG 2006. IEEE International*, 2006, pp. 197-197.
- [151] X. Zhu, M. Cheng, W. Hua, and J. Zhang, "Investigation of end-effect and experimental validation for hybrid excited doubly salient machine," in *Electromagnetic Field Computation, 2006 12th Biennial IEEE Conference on*, 2006, pp. 320-320.
- [152] Z. Zhu, J. Chen, Y. Pang, D. Howe, S. Iwasaki, and R. Deodhar, "Modeling of end-effect in flux-switching permanent magnet machines," in *Electrical Machines and Systems, 2007. ICEMS. International Conference on*, 2007, pp. 943-948.
- [153] Z. Zhu, Y. Pang, W. Hua, M. Cheng, and D. Howe, "Investigation of end effect in permanent magnet brushless machines having magnets on the stator," *Journal of Applied Physics*, vol. 99, p. 08R319, 2006.
- [154] Z. Zhu and Z. Azar, "Influence of end-effect and cross-coupling on torque-speed characteristics of switched flux permanent magnet machines," in *Power Electronics and ECCE Asia (ICPE & ECCE), 2011 IEEE 8th International Conference on*, 2011, pp. 145-152.
- [155] Y. Chen, Z. Zhu, and D. Howe, "Three-dimensional lumped-parameter magnetic circuit analysis of single-phase flux-switching permanent-magnet motor," *IEEE Transactions on Industry Applications*, vol. 44, pp. 1701-1710, 2008.

- [156] M. Al-ani, MJ and Z. Zhu, "Influence of end-effect on torque-speed characteristics of various switched flux permanent magnet machine topologies," *COMPEL: The International Journal for Computation and Mathematics in Electrical and Electronic Engineering*, vol. 35, pp. 525-539, 2016.
- [157] K.-Y. Nam, S.-R. Hwang, and P. S. Shin, "An end-effect equivalent factor for back-EMF analysis of PMSM," in *Electrical Machines and Systems (ICEMS), 2017 20th International Conference on*, 2017, pp. 1-4.
- [158] A. M. El-Refaie, M. R. Shah, J. P. Alexander, S. Galioto, K.-K. Huh, and W. D. Gerstler, "Rotor end losses in multiphase fractional-slot concentrated-winding permanent magnet synchronous machines," *IEEE Transactions on Industry Applications*, vol. 47, pp. 2066-2074, 2011.
- [159] H. M. Hamalainen, J. Pyrhonen, J. Nerg, and J. Puranen, "3-D finite element method analysis of additional load losses in the end region of permanent-magnet generators," *IEEE Transactions on Magnetics*, vol. 48, pp. 2352-2357, 2012.
- [160] D. C. Hanselman, *Brushless permanent magnet motor design*: The Writers' Collective, 2003.
- [161] W. L. Soong and T. Miller, "Field-weakening performance of brushless synchronous AC motor drives," *IEE Proceedings-Electric Power Applications*, vol. 141, pp. 331-340, 1994.
- [162] Infolytica. (2017). *MagNet. About Magnetic Field Analysis*. Available: <https://www.mentor.com/products/mechanical/magnet/magnet/>
- [163] M. D. L. (MDL). ((1-January-2016)). *Motor Design Limited, About Motor CAD*. Available: <https://www.motor-design.com/motor-cad-software/therm/>
- [164] ANSYS. (2017). *ANSYS Mechanical. About Mechanical Analysis*. Available: <https://www.ansys.com/products/structures>
- [165] Y. Chin and D. Staton, "Transient thermal analysis using both lumped-circuit approach and finite element method of a permanent magnet traction motor," in *AFRICON, 2004. 7th AFRICON Conference in Africa*, 2004, pp. 1027-1035.
- [166] S. Nategh, "Thermal analysis and management of high-performance electrical machines," KTH Royal Institute of Technology, 2013.