



MODELING AND TOPOLOGY INVESTIGATION
OF MODULAR MACHINES FOR WIND
GENERATOR SYSTEMS

Georgios Alpogiannis
4421574

Supervised By
Dr. Ir. Henk Polinder

A thesis submitted in partial fulfillment of the requirement for the degree of
Master in Electrical Engineering

Electrical Power Processing
Faculty of Electrical Engineering, Mathematics and Computer Science
Delft University of Technology
Delft, The Netherlands

October 2016

ABSTRACT

Wind energy has recently come to constitute more than 10% of the electric power consumed in the EU. As the wind industry is rapidly growing, wind turbines grow larger and new models are soon expected to reach ratings of 10 MW. The development of offshore wind farms is another major trend in the industry. Offshore farms are particularly suitable for installation of large turbines, able to utilize the stronger winds available. However, in case of an offshore failure, it could require weeks to access and repair the turbine. Even in onshore farms, the turbine size cannot allow frequent disconnections. In this context, increasing turbine availability emerges as a critical issue: Modularity is seen as a promising approach to respond to this need.

Modular systems can be decomposed into a number of independent 'modules' or components. Such systems can have their faulted modules bypassed and continue operation after fault, increasing system availability. Less frequent and easier repair can also be achieved. Different levels of modularity have been proposed: Systems with modular converters or systems with modularity both in the converter and machine, with possible levels of modularity in the machine itself. Focusing on the machine-side of modularity, this thesis aims to produce a comparative study of the main modular machine topologies proposed in literature.

The first step to address this problem, is to propose some promising machine topologies, suitable for modular design. Subsequently, a 2-D analytical model is developed, able to account for different types of winding. The model calculates important quantities to evaluate electromechanical performance (Back-EMF, Power, cogging torque) and efficiency (iron loss, copper loss). Analytical modelling results are validated by means of FEM modelling.

The validated analytical model is then used to carry a first level comparison of numerous modular winding machines. The designs which perform best in terms energy yield, efficiency and cost are promoted for a second level comparison, where further modularity is introduced. This time, the machine stator is segmented and different segmentation ideas are applied. The segmented designs are compared by means of FEM. After the optimal design is selected, the flux gaps introduced in the stator core are considered, and the influence of gap width on the machine performance is investigated.

ACKNOWLEDGMENTS

Firstly, I would like to thank Henk Polinder for his help and guidance throughout the duration of this thesis. The enlightening comments and feedback i received in our discussions helped me not to lose sight of the big picture, while struggling with particular details of the thesis.

I would very much like to thank Udai Shipurkar for all the time and effort he devoted for me during the thesis. He has been the main source of support to my various and frequently asked questions. His ideas helped me address problems in a soild way, and his help with FEM models was valuable.

Thanks also to my fellow master students Albert, Deesh, Takis, Alvaro and Chetan, with which i became good friends and worked on common projects during these months. The same holds for all master and PhD students of Henk, who I got to meet during the thesis presentation sessions.

Finally I would like to thank Christina, Dimitris as well as all my big family back home.

Contents

1	Chapter 1 : Introduction	1
1.1	Background	1
1.2	The growing need for modular systems	2
1.3	Thesis Objectives	2
1.3.1	Main Objective	2
1.3.2	Main Steps	2
1.4	Thesis Layout	3
1.5	Contribution	4
2	Modularity in electrical machines and drives	5
2.1	The Concept of Modularity	5
2.2	Applications of Modularity	5
2.3	Forms of Fault-tolerance - Relation to Modularity	6
2.4	Levels of Modularity	7
2.4.1	Modularity in the Power Electronic Converter	7
2.4.2	Modularity in the Machine windings	11
2.4.3	Segmented Machine Topologies	12
2.5	Generator Topologies	15
2.5.1	The type of electrical machine	15
2.5.2	Modular generators with enhanced fault tolerance	16
2.6	Main challenges for Modular Machine topologies	17
2.6.1	Challenges in Control	17
2.6.2	Challenges in machine structure	18
2.7	Summary	18
3	Analytical Modeling of a Radial Flux PM machine	20
3.1	Analytical models	20
3.1.1	Choice of analytical model	20
3.1.2	Analytical model assumptions	21
3.2	Field produced by magnets	22
3.2.1	Magnetization in polar coordinates	22
3.2.2	The Field equations	23
3.3	Magnetic Field due to stator currents	25
3.3.1	Magnetic Field produced due to a stator coil	25
3.3.2	Magnetic Field produced from a 3-phase distributed winding	26
3.3.3	Magnetic Field produced from a 3-phase FSC winding	27
3.4	Effect of stator slotting	28
3.4.1	The relative permeance function	28
3.4.2	Slotted machine results	30
3.5	Adjusting the radius and slot opening size	32
3.6	Back-EMF and Output Power	33
3.7	Torque and Cogging Torque	35
3.7.1	Torque	35

3.7.2	Cogging Torque	36
3.8	Losses	38
3.8.1	Iron Losses	38
3.8.2	Copper Losses	41
3.9	Summary	42
4	Comparison of modular winding machines by means of analytical modeling	44
4.1	Design choices for comparison	44
4.2	Comparison Criteria	45
4.3	Choice of slot pole combinations	46
4.3.1	Annual Energy Yield	47
4.3.2	Efficiency	48
4.3.3	Cost	50
4.3.4	Conclusions - Final choice of topologies	52
4.4	Summary	53
5	Comparison of segmented machine designs	54
5.1	Design choices and Comparison criteria	54
5.2	Comparison and Results	54
5.2.1	Output Power	56
5.2.2	Efficiency	57
5.2.3	Cost	57
5.2.4	Conclusions	58
5.3	Effect of the flux gap size on machine performance	59
5.3.1	Effect on output power	59
5.3.2	Effect on power ripple	60
5.4	Summary	60
6	Conclusion	62
6.1	Conclusions	62
6.1.1	Theoretical work	62
6.1.2	Analytical modelling	62
6.1.3	First-level comparison	62
6.1.4	Second-level comparison	63
6.2	Future Work	63

1 Chapter 1 : Introduction

1.1 Background

With an overall installed power of 128.28 GW by the end of 2014, wind energy has come to constitute more than 10% of the electric power consumed in the EU [1] . Since 2010, significant growth has also been noted outside the traditional markets in Europe and North America, mainly because of large construction developments in China. Recent environmental state commitments on climate change and limiting of fossil fuels, are also expected to create favourable incentives for the wind industry [2].

Figure 1.1 illustrates the net installed power (MW) for different technologies, during the decades 1995-2015:

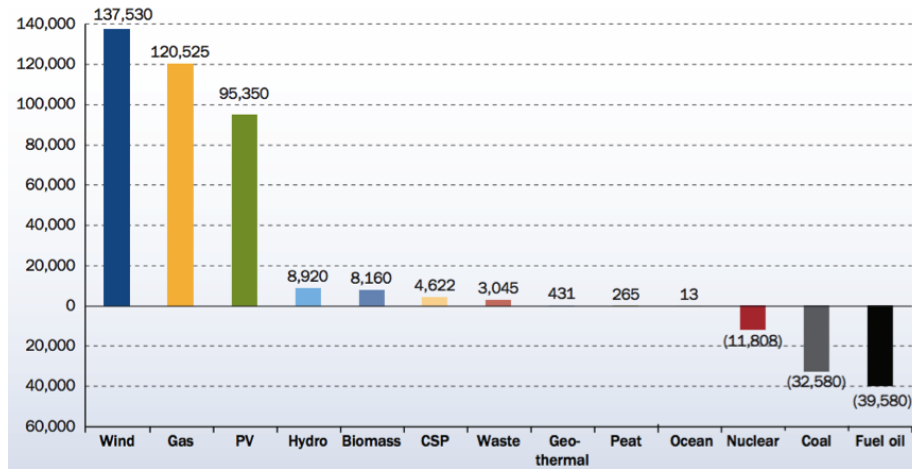


Figure 1.1: Net electricity generating installations in the EU, 1995-2015 (MW)[3]

It can be seen that wind is the first renewable energy technology in installed capacity, with a net growth of 137.530 MW. A net decrease can be observed in conventional sources (coal, oil) as well as nuclear power, over the last twenty years.

As the wind industry is rapidly growing, the turbine size also grows and new models are soon expected to reach ratings of 10 MW. Another important trend in the wind turbine sector concerns the development of large offshore wind farms. Offshore farms are particularly suitable for installation of large turbines, able to utilize the more frequent and stronger winds available. During the first half of 2015, 584 offshore turbines, with an average size that exceeds 4 MW, have been installed in Europe [4].

1.2 The growing need for modular systems

Both of the aforementioned trends, increase the need for reliable wind turbines with high availability: For remote offshore farms, in case of a failure, it could take weeks for the maintenance team to access and repair the turbine [5]. Even in onshore wind farms, the growing turbine size cannot allow frequent disconnections, as the cost of the lost energy would increase severely. Modularity is seen as a promising approach to meet the need for reliable, highly available systems.

Modularity refers to the concept of designing systems that can be decomposed into a number of "modules" or components [6]. Such systems could have their faulted modules bypassed and continue operation after a fault (fault tolerance). The post fault operation can be optimised by appropriately adjusting the control of the drive, so that the level and quality of delivered power is not severely reduced. This would allow repair to be postponed until the next scheduled regular maintenance [5]. Moreover, in a modular design, the faulted module could be more easily replaced or repaired.

Different levels of modularity have been proposed in literature: Systems with modular converters [7, 8, 9] or systems with modularity both in the converter and machine [10], with possible levels of modularity in the machine itself (modular windings, segmentation).

1.3 Thesis Objectives

1.3.1 Main Objective

The primary objective of this thesis is to produce a concentrated comparative study of the main modular machine topologies proposed in literature. Starting from a first level of modularity only in the stator winding, different designs are compared based on electromechanical performance. Each topology promoted from the first level comparison, is simulated with FEM for different types of segmentation. The optimal segmentation strategy for each individual topology is eventually selected.

1.3.2 Main Steps

To address this problem, two different modelling approaches will be adopted. The first level comparison is based on analytical modelling, whereas the different segment designs are compared by means of FEM. The main objective is then broken down into a number of subsidiary steps, which are described subsequently:

1. Propose some promising machine topologies, which are suitable for modular design.
2. Develop an analytical model of the proposed machine topologies, accounting for different types of winding. Calculate important quantities to evalu-

ate electromechanical performance (Power, Torque, cogging torque, THD of back-EMF) and efficiency (iron loss , copper loss).

3. Validate the analytical modelling results by means of FEM modelling. Conclude the first level comparison using the validated analytical model and select the topologies to be promoted for segmentation.
4. For each promoted topology, apply different segmentation ideas. Compare the segmented designs using FEM. Examine the effect of adjusting the flux gap size.

1.4 Thesis Layout

The individual steps needed to accomplish the Thesis's objectives, are implemented in 5 chapters, which are described subsequently:

- Chapter 2 presents a detailed review on the concept of modularity. Firstly, the various possible levels of modularity are presented, and the most promising applications of modular systems are highlighted. Next, generator topologies which are suitable for modular design are discussed. Subsequently, several modular and segment designs proposed in literature, are reviewed. Then, specific pole-slot combinations that result in fault tolerant machines are highlighted. Finally, the future challenges and prospects of modularity are discussed altogether.
- Chapter 3 presents a 2-D analytical modeling of the surface PM machine, accounting for both distributed and modular winding designs. Firstly, the magnetic field due to the PM's alone is calculated in section 3.2. Then, the armature reaction field is calculated and the effect of slotting is included. Subsequently, the induced EMF, power, torque, cogging torque and stator losses are computed. The analytical model results are validated with FEM simulations.
- Chapter 4 presents the first level comparison between different modular machine topologies. Different slot/pole combinations of modular winding designs are compared. A varying axial length is assumed to maintain the same speed and airgap power in all machines. The base case of distributed winding with 1 slot per pole per phase is considered as a reference. The validated analytical model is used to promote the topologies which perform the best, based on energy yield, efficiency and cost .
- Chapter 5 includes the second level comparison which is drawn by means of FEM modelling. Each topology promoted from the first level comparison, is now designed in FEM for different types of segmentation. The optimal segmentation idea is then selected for each specific topology. The flux gaps introduced by segmentation are finally considered, and the effect of different gap size on the machine performance is evaluated.

- Chapter 6 summarizes the obtained results, and presents the main conclusions drawn. Recommendations which may help to enhance the value of the present work are briefly provided. Potential related future work on modular systems is conclusively discussed.

1.5 Contribution

This thesis focuses on the machine-side of modularity. It draws a comparative study of modular designs, which have been individually proposed in literature, but have not been compared in an aggregated manner.

For this, a 2-D analytical model able to account for both distributed and modular windings has been developed. The model is capable of drawing fast comparisons between different machines, which would otherwise require computationally expensive FEM simulations. Therefore, it is a useful and complete tool to address time-demanding problems such as the considered topology investigation.

Regarding the segment topologies, the effects of varying the flux gap size in the segmented machine, is also a topic that has not yet been widely covered.

2 Modularity in electrical machines and drives

2.1 The Concept of Modularity

Modularity refers to the concept of designing systems that can be decomposed into a number of 'modules' or components [6].

One of the best examples of modular design is the computer: A typical computer is assembled from modules such as processors, hard drives, memory units, power supply units, graphics and sound cards. Each of these modules is a replaceable part allowing for easy repair or upgrade. Other recent examples of systems with modular design include modular smartphones[11] and modular, prefabricated buildings.

2.2 Applications of Modularity

In electrical engineering, modularity can be found in the design of PV systems, as well as in electrical machines and drives. In the past, machines with readily assembled modular windings have been used in low power applications due to advantages such as higher slot fill factors, shorter end-windings and simpler manufacture.

The ongoing research on modular windings, revealed that certain modular machine topologies can provide higher physical, magnetic and thermal isolation between phase windings. This means that in case of fault in one phase, the effects on the other phases are minimized and the system is well capable of continuing operation after fault (fault tolerance). This prospect of more reliable, fault-tolerant systems, rendered modularity very attractive for safety-critical applications, such as aircraft fuel pumps and flight control surface actuators [12]. Figure 2.1 depicts a modular, fault-tolerant machine and motor drive, designed in TU Delft to be used in an aircraft application:

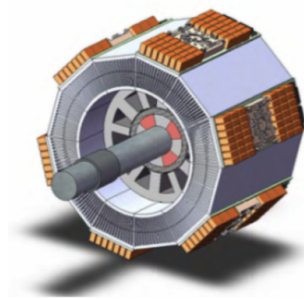


Figure 2.1: Modular machine and power electronics - Integrated concept for aerospace application [10]

Spooner et al. were the first to propose modular machines for wind generator systems [13]. The proposed topology was that of Direct Drive (DD)

systems, which adopt very large machines to eliminate the need for a gearbox. A schematic depiction of a DD Permanent Magnet (PM) generator can be seen in Figure 2.2: As DD systems grow in popularity, and the generator size keeps

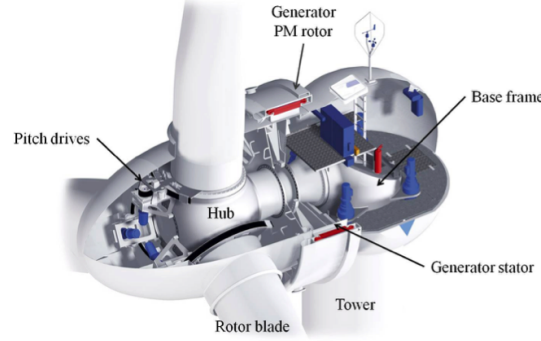


Figure 2.2: Direct-Drive PM generator : A promising candidate for modular systems (%)

increasing, benefits such as easier manufacture and transportation become more and more significant. Moreover, contrary to DFIG configurations, DD machines have no direct connection of the stator winding to the grid. This allows for more freedom when it comes to employing modular winding and converter configurations.

The prospect for enhanced fault-tolerance, is another major driving force towards modular wind generator systems. The aim here is not to ensure uninterrupted operation by all means, but to increase the overall turbine availability [5]. As the rating of turbines is soon expected to reach levels of 10 MW, and the world's dependency on wind power is growing, the currently high downtimes of wind turbines tend to become unacceptable.

A modular wind generator system could have its faulted module bypassed. Moreover, post fault operation can be optimized by appropriately adjusting the control of the drive, so that the level and quality of delivered power is not severely reduced. That could allow repair to be postponed until the next scheduled regular maintenance [5]. This is crucial for remote offshore farms where access for maintenance can be difficult. Finally, maintenance itself is simplified, as the faulted module can be more easily replaced or repaired on site.

2.3 Forms of Fault-tolerance - Relation to Modularity

The next sections present modular concepts, proposed as a means of improving availability or reliability. In this sense, all presented topologies can enhance system fault-tolerance. However, not all possible forms of fault-tolerance are addressed, but only those that contribute to a more modular system. The major

forms of fault tolerance both in the machine and converter are briefly presented subsequently for completeness [5]. Those which will be further addressed in the subsequent sections, are highlighted:

- **Converters with redundant semiconductors.** In this case, additional semiconductors are added to the stack of semiconductors comprising every switch. This solution is more attractive only when voltage levels are high enough, so that all available semiconductors are needed. Otherwise, the rise in number of semiconductors and the respective losses make this solution somewhat unattractive.
- **Fault tolerant converter topologies.** In this case, power semiconductors and fuses are added to make the converter fault tolerant. The switch/double-switch redundant topologies are such proposed examples. Another variation, the phase redundant topology, includes a spare phase leg which is capable to replace a faulty phase leg. Finally, the cascaded converter topology devotes a full bridge converter for each individual phase. This topology will be addressed in the following sections, as it can be combined with various modular concepts.
- **Increasing the number of phases.** Two possible approaches exist here: If one phase fails, the sound phases continue. Alternatively, when one set of three phases fails, the other sets of three phases continue. Both methods are promising and will be discussed in the next sections. Higher phase number systems are well-combined with modular concepts and have already been applied in the wind industry.
- **Design for fault tolerance.** From the beginning, both converter and machine are designed to tolerate a specific range of faults. The design needs to carefully consider interaction between converter and machine: For example in case of a turn-to-turn s/c in a machine coil, the whole coil can be short-circuited by the converter, to maintain nominal s/c current. Design choices for modular, fault-tolerant machines are discussed in the next sections.

2.4 Levels of Modularity

Different levels of modularity have been proposed in literature: Systems with modular converters [7, 8, 9], or systems with modularity both in the converter and machine [10], with possible levels of modularity in the machine itself (modular windings, segmentation). A more detailed description of the levels of modularity will be attempted in the following sections.

2.4.1 Modularity in the Power Electronic Converter

2.4.1.1 Converters connected in Parallel

Modular power electronic converters are already a well-proven, commercialized technique towards enhancing wind turbine system reliability. A common ap-

proach for state-of-the art turbines with power levels that exceed 3MW, is to adopt multiple converter modules operating in parallel. This arrangement gives the opportunity to share the generated power and maintain a low voltage and current rating for each module. The Gamesa G10x 4,5 MW model [8] and the Siemens multimegawatt turbines [14], are both well known examples of this topology. A parallel configuration of two level (2L) back-to-back converters can be seen in Figure 2.3

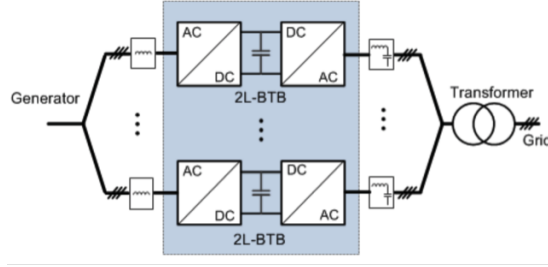


Figure 2.3: Two level back-to-back converters in parallel [15]

In the case that a converter module fails, it is switched off and the rest of the modules continue operation. Even though the output power capability is reduced, the effect may not be particularly harmful for wind turbine systems, as they rarely operate at rated power level.

Control systems have been developed based on the topology of Figure 2.3. The ability of the control to achieve the necessary dynamics for system level events such as Low-Voltage Ride Through (LVRT) has been verified in [14]. On top of higher reliability, other control-derived benefits are obtained: These include higher efficiency by means of a variable number of active converter modules and elimination of grid harmonics by imposing a phase shift between the PWM-pulses [8].

2.4.1.2 Cascaded converter topology

In fault-tolerant topologies which employ a higher number of phases, another attractive variation of parallel redundancy, known as the cascaded inverter topology, has been proposed [9]: According to this, each of the stator phases is excited by an independent single-phase inverter. Even though failure of an individual phase-drive unit will degrade the output power, the need for immediate repair is reduced since no system shutdown is required. A phase redundant system for a n -phase machine is depicted in Figure 2.4: Compared to that of Figure 2.3, this topology is capable of tolerating faults not only in the converter, but in a machine phase as well. In addition, with a careful choice of the number of phases and proper modification of the control scheme, the post-fault performance degradation can be minimized [9].

Finally, a major benefit for high power systems, is that the machine and

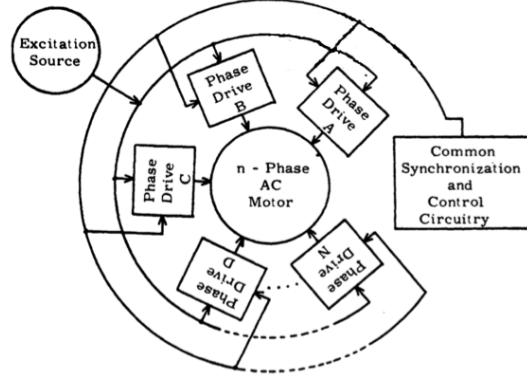


Figure 2.4: The cascaded converter topology [9]

power electronics can easily be reconfigured into a multi-phase generator system without significant cost. Alternatively, parallel 3-phase machine circuits can be combined with modular converters, instead of increasing the number of phases. In this case, if one set of three phases fails, the other sets of three phases continue: The post-fault power is then reduced, but the machine currents remain symmetric after fault.

2.4.1.3 Multilevel Converters

The high power rating of the proposed modular systems, implies that the DC link voltage will increase to levels of about 3-5kV. As two level converters cannot easily handle these voltage levels and series connection of switches has proven to be difficult, multi-level converters are needed [5].

A multi-level topology known as the Modular Multi-level Converter (MMC), has been proposed recently to further increase the dc operating voltage. In the MMC, several elementary switching sub-modules (H-Bridges) are stacked together. A schematic diagram of this topology can be seen in Figure 2.5:

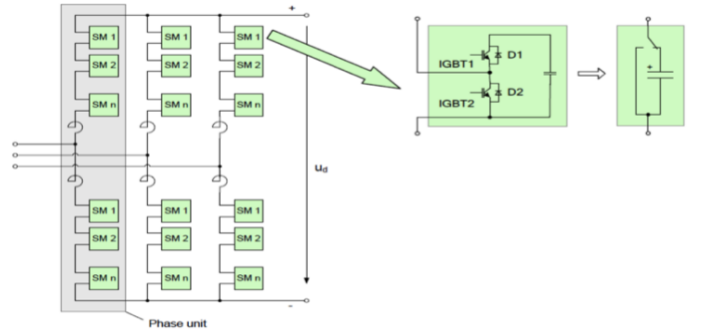


Figure 2.5: Modular multi level converter topology

Using up to a hundred or more levels, the MMC can produce an essentially sinusoidal ac waveform, and eliminate the need for additional filtering [16]. The modular built of the MMC, as well as its suitability for HVDC connected offshore farms, make this topology an attractive choice for high rating modular wind generator systems.

The MMC avoids the challenge of connecting semiconductor switches in series. Instead, the voltage rating is adjusted by adding sub-modules to the stack. In this way, converters with very high power and voltage ratings can be realized. This ability for generation of AC or DC voltage in the 10-100kV range, stimulated the investigation of transformer-less designs [17]. Such a MMC arrangement with eliminated dc link is depicted in Figure 2.6:

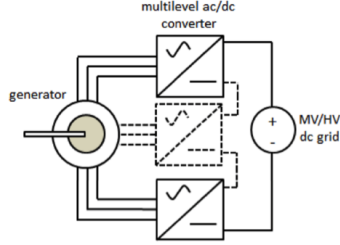


Figure 2.6: Modular multi-level generator-converter topology [5]

The stator in the machine of Figure 2.6 is segmented: Each segment operates independently as a generator and its winding is connected to a converter module. Segmented machines will be reviewed in more detail in the following sections. The design of Figure 2.7 allows connection to a MVDC or HVDC connection point. This is of particular interest in the case of offshore wind farms, where HVDC connection is commonly employed.

Another MMC arrangement, this time without elimination of the dc link is depicted in Figure 2.7 :

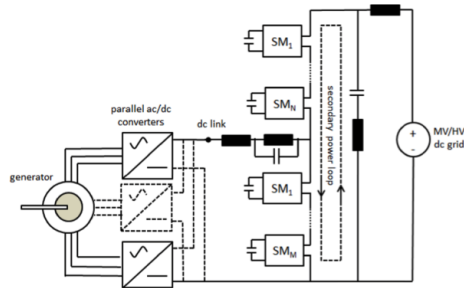


Figure 2.7: Transformerless Modular multi level generator-converter topology with DC link [5]

With this arrangement, the dc link voltages are kept at a level which does

not affect the stator windings insulation. A dc step-up transformer is then used to reach the required MV/HV DC level.

2.4.2 Modularity in the Machine windings

2.4.2.1 Modular Windings

Fractional-slot concentrated (FSC) windings, also known as 'modular' windings, comprise of coils wound around a single stator tooth. Figure 2.8 depicts single-layer (SL) and double-layer (DL) arrangements for FSC windings: We observe



Figure 2.8: SL (a) and DL (b) arrangements for FSC winding

that in the SL arrangement every alternate tooth carries a coil, where as in the DL case every tooth has its own coil wound.

Though not a recent discovery, FSC windings had only been considered for low power machines with reduced number of phases, until the late nineties. In recent years however, FSC winding has gained attention, also for high performance applications including wind generators.

Some well known advantages include [18]:

- Simple assembly and mounting.
- Short, non-overlapping end-windings. Lower copper loss.
- More standardized windings. Reduced cost.
- Low torque ripple for certain pole-slot combinations.
- Suitability for modular, fault-tolerant designs.

A main challenge for application of FSC windings regards the combination of the pole and slot number. If this is not selected carefully, the resulting machine may have very poor winding factor, high rotor losses, and high torque ripple.

It has been shown that the higher pole number, the easier it is to find appropriate pole-slot combinations, which result in high winding factors and low ripple. Therefore, multi-pole PM machines for low-speed direct drives have been confirmed as a very suitable topology to combine with FSC windings. The main remaining problem in this case, is the high rotor eddy-current loss caused by the winding's higher order space harmonics, which may reduce overall efficiency.

2.4.2.2 Modular Windings with dedicated Converters

So far, systems with modularity only in the converter have been considered. Some topologies, such as the cascaded converter topology of Figure 2.4, employ separate converter modules for each phase. In case of a stator coil fault, the respective phase is switched off, allowing the sound phases to continue operation.

A further step in this direction, would be to employ separate converter modules for each stator coil. This concept can be applied only if the stator is equipped with FSC windings, instead of conventional distributed windings. Such an arrangement is depicted in Figure 2.9: This time in case of a stator

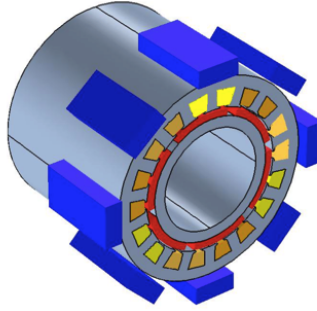


Figure 2.9: Modular winding machine with dedicated converters per coil [6]

coil fault, only the faulted coil is switched off, instead of the whole phase. This means greater ability to process power in post-fault operation. It remains a question though, if the power cost savings outweigh the additional expenses that come with the higher number of power electronic components used.

2.4.3 Segmented Machine Topologies

2.4.3.1 Introducing segmentation in the machine

An interesting concept for achieving greater modularity in the machine itself, is to segment its iron core. A segmented machine example, in which every segment comprises of a concentrated coil and its converter, is presented in Figure 2.10: In case of fault in a stator coil, the whole core module can now be removed and replaced. Apart from easier and quicker maintenance, the segment topology also facilitates the construction process: Fewer drawings and tools are required, as the need for detailed design is reduced. In addition, standardization is enhanced, as the same construction pattern can be used for different machines.

Transportation of the machine is also facilitated, as the individual segments can be transported separately and assembled on site. This is particularly advantageous for the large, heavy machines used in DD wind generator systems. The main drawback of the topology seems to be a higher cogging torque caused by the flux gaps, which are introduced by segmentation.

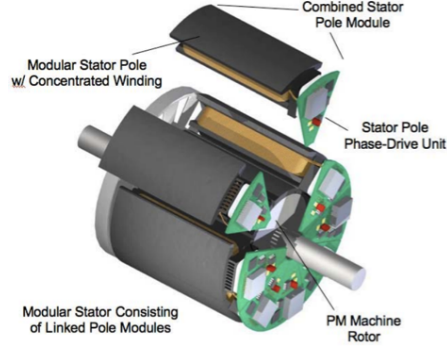


Figure 2.10: Machine with segmented stator and modular winding/converter [19]

Various ideas on how to segment the machine, have been proposed in literature. Most of the proposed designs suggest different ways of segmenting the stator, and these will be presented subsequently. However, segmentation of the rotor has also been proposed [13] in order to make the PMs (or rotor winding) replaceable.

2.4.3.2 E-core segments

The first proposed segment topology is to build E-shaped segments, where every segment comprises of a single concentrated coil. The machine is equipped with a SL winding. Figure 2.11 depicts such an arrangement: The stator teeth in

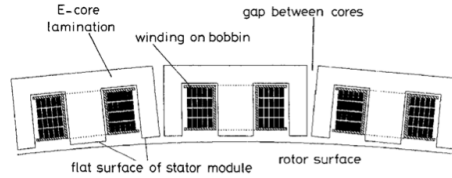


Figure 2.11: Stator with SL winding and E-core segments [13]

Figure 2.11 are straight. For a stator with alternate teeth having tooth tips, investigation on which teeth are more suitable for carrying the winding, has been carried out for different pole-slot combinations [20]. Another variation of the E-core design, suggests larger E-cores with three phase winding instead of single coil ones. This design offers a lower number of flux-gaps, at the expense of reduced manufacturability.

E-core segments combined with DL winding, is another possible variation. However, such a design would require additional spacer teeth, and eventually reduce core utilization. Moreover, some fault-tolerant characteristics of SL modular windings would be lost.

2.4.3.3 T-core segments

In E-core designs, the flux gaps are placed in the center of the stator teeth. Another possible choice is to place the flux gaps in the stator yoke instead. In this way, 'T-core' segments are formed. Such a design is depicted in Figure 2.12: Placing the flux gaps in the yoke can downgrade performance, as it increases the

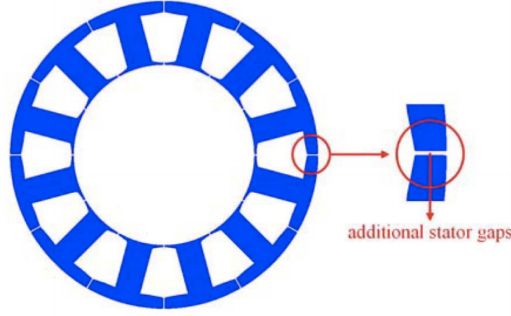


Figure 2.12: Stator with T-core segments [21]

overall reluctance of the magnetic circuit. However, the possibility of achieving very high slot fill-factors has been reported [22]. Again, possible design variations include the choice of winding layers, as well as the selection between single or three-phase segments.

2.4.3.4 Other types of segmentation

A design in which certain sections of the stator core were completely removed, was proposed in [23]. Figure 2.13 depicts this arrangement:

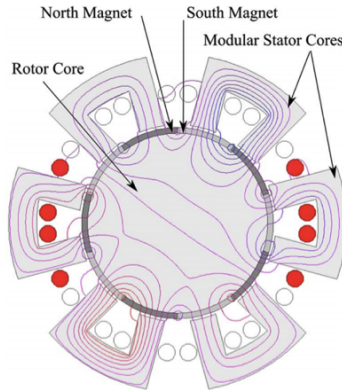


Figure 2.13: Segmented stator with part of the iron core removed [23]

The topology facilitates manufacture and uses less iron material. The performance however was not severely downgraded, only when the number of slots and poles is similar.

Finally, Figure 2.14 depicts a design with segmental teeth and solid back iron, proposed in [24]:

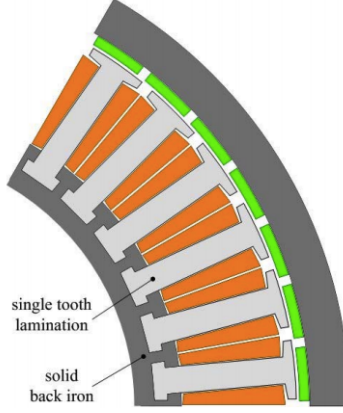


Figure 2.14: Segmental teeth with solid back iron

In this case a faulted coil can be replaced without the need of replacing the back iron. Higher fill factors are possible once again.

2.5 Generator Topologies

2.5.1 The type of electrical machine

So far, the generator topologies that have been proposed for modular systems, are almost exclusively synchronous machines.

The high space harmonics produced by modular windings, make them unsuitable to combine with induction machines (IM). As the IM rotor adapts to any frequency imposed by the stator, each winding space harmonic will be trying to rotate the rotor at a different speed. It will be then hardly possible to produce any useful torque.

The flux-switching PM (FSPM) machine is another recently proposed topology, for fault tolerant wind generator systems [25]. The stator of the machine is equipped with PMs as well as modular windings. Figure 2.15 depicts a segment of a multi-pole FSPM machine proposed for low speed Direct-Drive applications.

The localized flux paths in the resulting magnetic circuit, grant it with an inherent fault-tolerance. However, the machine has been mainly studied for smaller generators so far, due to structural reasons. In addition, the double saliency in the rotor and stator tend to increase the cogging [25].

The well known Radial-flux synchronous machines are eventually the main candidate for Direct-Drive wind generator systems. While electrically excited

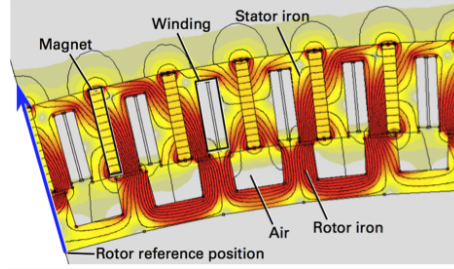


Figure 2.15: FSPM machine segment for low speed Direct-Drive applications

machines are also used, the main trend is towards PM machines with variations regarding the magnets design.

2.5.2 Modular generators with enhanced fault tolerance

Fault tolerance is usually defined [FT] as the ability of a system to remain in operation after a fault.

The design of a machine for fault tolerance typically includes requirements for electrical, magnetic and thermal isolation between the machine windings [5]. Single-Layer (SL) modular windings offer perfect electrical isolation, and better thermal isolation compared to DW. The main improvement though, is enhancing magnetic isolation: Certain slot and pole combinations have been found to result in negligible coupling between phases. This way, a fault in one phase will not induce high currents in neighbouring phases.

Figure 2.16 shows how a machine with 12 slots can be adjusted to have low mutual coupling by shifting the pole number from 8 to 10:

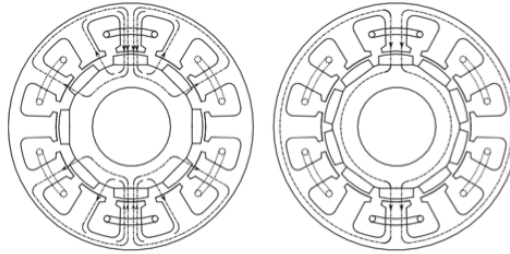


Figure 2.16: Minimizing mutual coupling, by appropriate slot-pole choice

The condition for obtaining machines with negligible coupling, is the following :

$$2p = Q(1 \pm \frac{n}{2m}) \quad (2.1)$$

Where :

- p = pole pairs
- Q = slots
- m = phases
- n = any odd integer less than m , such that $\text{GCD}(n,m) = 1$.

At the same time, to maintain a high coupling between the PM's and the rotor, the ration of n/m needs to remain above 0.6.

Regarding post fault operation, fault-tolerance aims to reduce the harmonics, torque ripple and unbalanced pull that will result from switching off the faulty module. These effects could be minimized in a modular machine, in which the control switches off appropriate healthy modules to reduce the asymmetry [26].

Finally, modular generators with higher numbers of phases have also been broadly proposed for wind generator systems. Machines with parallel coils and converters, can be easily reconfigured in multiphase arrangements. This way, increased fault tolerance is achieved at a low cost, without many added components [5].

2.6 Main challenges for Modular Machine topologies

2.6.1 Challenges in Control

The main challenge for control in modular machine and converter systems, is that each individual module must now be controlled independently.

This requires a more elaborate control system, which can be realized with either a centralized or a distributed scheme: In the first case, the fault-tolerance of the system highly depends on the reliability of the central control unit [6]. As the failure rate of the latter is relatively high, centralized control is not such a good choice for fault tolerant applications.

A layered control system including both a central controller and local controllers in each converter unit, has been proposed in [27]. A schematic diagram for such an architecture can be seen in Figure 2.17:

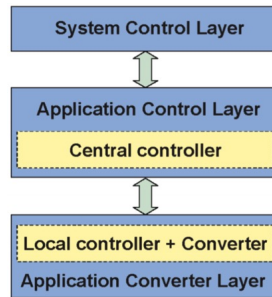


Figure 2.17: Layers of a distributed control scheme for modular machines

The central controller runs the main control loop and delivers commands to the local controllers, which generate the switching commands for the IGBT's. Fast communication between units is achieved with fiber optics. Finally, complete distribution of control has also been in [26]. In the absence of a central controller, the system can become fault-tolerant to failures of control units as well.

2.6.2 Challenges in machine structure

Large, heavy machines used in DD systems already pose mechanical challenges: A constant air-gap is difficult to maintain. Also a bulky supporting structure is needed, much of which is not active [26].

In a segmented stator, the stresses applied locally on each module are not absorbed by the iron core as in monolithic designs, but instead they are transferred directly to the structure. Thus, each segment needs to be supported by a dedicated support unit. Such a design, proposed by Spooner, is shown in Figure 2.18: Similar challenges arise from segmentation of the rotor iron.

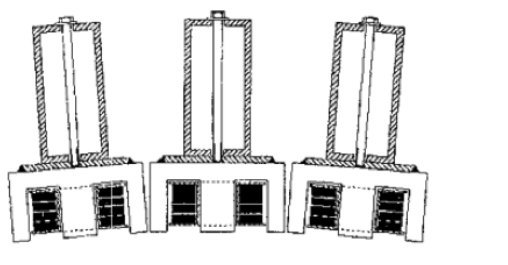


Figure 2.18: Dedicated support structure for each stator segment

2.7 Summary

This chapter presented an overview of proposed ideas for modular wind generator systems. In the following chapters our attention will be focused only on the machine side of modularity.

All opportunities and challenges presented in the above sections, must be reviewed also in terms of cost. Additional costs of modular systems typically include increased number of switches (but reduced rating) for the power electronics. The cost of additional control hardware should be outweighed by resulting savings due to enhanced system reliability/availability. The same is true for additional costs associated with support structures and design for thermal management.

As the cost of power electronics drops and the need for reliable wind generator systems increases, the benefits of modularity will outweigh the costs. However, an individual assessment for each modular feature must be carried

out on a case-by-case basis, and a lot of research and progress is still further ahead.

3 Analytical Modeling of a Radial Flux PM machine

3.1 Analytical models

The present chapter presents an analytical model for calculating the instantaneous magnetic field distribution in the airgap of a surface permanent magnet, radial field synchronous motor. Firstly, the magnetic field due to the PM's alone is calculated in section 3.2. Then, the armature reaction field is calculated and the effect of slotting is included. Subsequently, the induced EMF, power, torque, cogging torque and stator losses are computed.

Analytical models are a useful modeling alternative to FEM: Although they lack in precision, they require significantly lower computational time. As a result, a common practice is to employ analytical models for comparisons of a high number of topologies and use FEM subsequently, in order to test the most promising designs [28]. A similar approach is followed in the present thesis: The presented analytical model, is validated with FEM results, before it is used to draw a first-level comparison between different modular machines.

3.1.1 Choice of analytical model

Several analytical models for RFPM machines have been proposed in literature. The most accurate but also most complex ones, are the recently proposed 'subdomain' models [29]. In subdomain models the actual machine geometry, including the slots, is addressed in the mathematical formulation. The different subdomains or regions of such a model are depicted in Figure 3.1 :

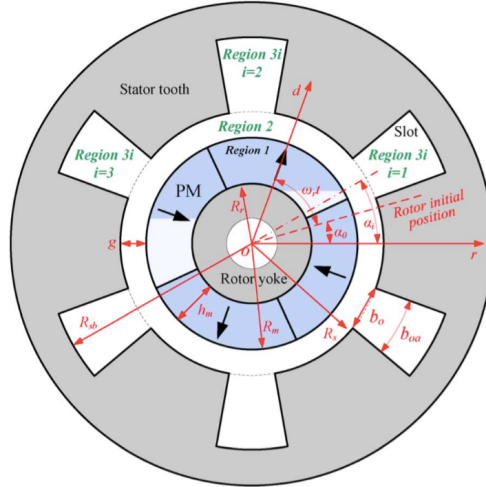


Figure 3.1: Regions of an exact subdomain analytical model[29]

An easier approach is to initially address a simpler, slotless machine geometry and then account for the slotting effect by means of a Relative Permeance (RP) function. The permeance can be modeled either as a real or as a complex function, and in both cases it is derived using conformal mapping [30]. A complex function can account for influence between radial and tangential flux density components due to slotting. However, this comes at the expense of much higher computational time: The conformal mapping is needed to calculate every single point in the permeance waveform, whereas a real permeance requires a conformal mapping calculation only in one point [31].

In the present work, the real RP model developed by Zhu and Howe [32, 33, 34, 35] was employed, due to its recognition and relative simplicity. Starting from Maxwell's equations, the model computes the scalar magnetic potential by solving a Laplace equation in the airgap and a quasi-poissonian equation in the PM.

2-D modeling is chosen instead of 1-D, in order to account for the large effective air-gap of surface PM machines. Variation in the radial direction can then also be considered and the model can account for flux focusing and leakage effects. The equations are expressed in polar coordinates: In this way, the model accounts for the direction of magnetization (radial or tangential), in the useful magnetic flux per pole [32].

3.1.2 Analytical model assumptions

The following assumptions are made to simplify the mathematical formulation and solution of the problem [28]:

1. The machine is slotless, with a constant airgap size.
2. An equivalent time-varying current sheet along the stator bore is assumed instead of the actual windings.
3. The relative permeability of the iron is infinite.
4. The relative permeability of the PM's is constant (but not necessarily unity).
5. All materials are linear and isotropic.
6. Everything remains constant in the z-direction (3-D effects are neglected). This is valid when the airgap is much smaller than the axial length of the machine.
7. End effects are neglected.

The geometry parameters of a 3 MW PM machine found in [36] are used in this chapter, to validate the analytical model.

3.2 Field produced by magnets

The geometry addressed in the present model is depicted in Figure 3.2.

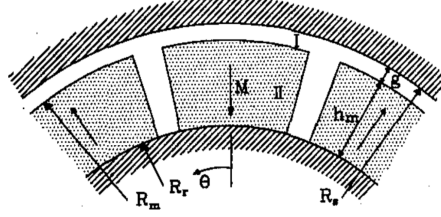


Figure 3.2: Motor geometry for the present analytical model [32]

Unlike Figure 3.1, only two regions are considered here, the airgap (region I) and the PM's (region II).

3.2.1 Magnetization in polar coordinates

The solutions for the governing field equations are subsequently derived, assuming that the magnets have:

- Uniform magnetization only in the radial direction.
- Constant relative recoil permeability (linear 2nd quadrant demagnetization curve).

In polar co-ordinates the magnetization vector $\vec{M}$ is given by [32]:

$$\vec{M} = M_r \vec{r} + M_\theta \vec{\theta} \quad (3.1)$$

Where

$$M_r = \sum_{n=1,3,5}^{\infty} M_n \cos(np\theta) \quad (3.2)$$

$$M_\theta = 0 \quad (3.3)$$

since we assumed only radial magnetization. Thus, the magnetization will have the shape depicted in Figure 3.3 :

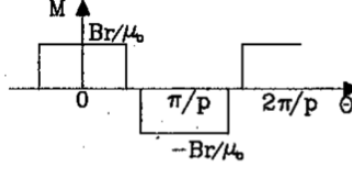


Figure 3.3: PM Magnetization [32]

The Fourier coefficients of the radial magnetization are then given by:

$$M_n = 2(B_r/\mu_o)\alpha_p \frac{\sin\left(\frac{n\pi\alpha_p}{2}\right)}{\frac{n\pi\alpha_p}{2}} \quad (3.4)$$

Where α_p is the magnet pole-arc to pole-pitch ratio.

3.2.2 The Field equations

The Laplace and quasi-Poisson equations for the scalar magnetic potential φ in the two regions, is given in polar coordinates by:

$$\frac{\partial^2 \varphi_I}{\partial r^2} + \frac{1}{r} \frac{\partial \varphi_I}{\partial r} + \frac{1}{r^2} \frac{\partial^2 \varphi_I}{\partial \theta^2} = 0, \text{Airgap}(I) \quad (3.5)$$

$$\frac{\partial^2 \varphi_{II}}{\partial r^2} + \frac{1}{r} \frac{\partial \varphi_{II}}{\partial r} + \frac{1}{r^2} \frac{\partial^2 \varphi_{II}}{\partial \theta^2} = \frac{M_r}{r\mu_r}, PM's(II) \quad (3.6)$$

where φ_I, φ_{II} are related to the radial and tangential components of the magnetic field intensity vector $\vec{H}$ according to:

$$H_r = -\frac{\partial \varphi}{\partial r} \quad (3.7)$$

$$H_\theta = -\frac{1}{r} \frac{\partial \varphi}{\partial \theta} \quad (3.8)$$

The field intensity $\vec{H}$ is related to $\vec{B}$ according to :

$$\vec{B}_I = \mu_0 \vec{H}_I \quad (3.9)$$

$$\vec{B}_{II} = \mu_m \vec{H}_{II} + \mu_0 \vec{M} \quad (3.10)$$

Once we obtain a solution for the scalar potential φ , we can derive an expression for the magnetic flux density using equations (3.7-3.10). The general solutions of equations (3.5) and (3.6) are then obtained as:

$$\varphi_I(r, \theta) = \sum_{n=1,3,5}^{\infty} (A_n I r^{np} + B_n I r^{-np}) \cos np\theta, \text{Airgap}(I) \quad (3.11)$$

$$\begin{aligned}\varphi_{II}(r, \theta) = & \sum_{n=1,3,5}^{\infty} (A_{nII}r^{np} + B_{nII}r^{-np})\cos np\theta \\ & + \sum_{n=1,3,5}^{\infty} \frac{M_n}{\mu_r(1-(np)^2)} r \cos np\theta, PM's(II)\end{aligned}\quad (3.12)$$

To compute the constants A,B in the above equations, the following boundary conditions are applied:

$$H_{\theta I}(r, \theta)|_{r=R_s} = 0 \quad (3.13)$$

$$H_{\theta II}(r, \theta)|_{r=R_r} = 0 \quad (3.14)$$

$$B_{rI}(r, \theta)|_{r=R_m} = B_{rII}(r, \theta)|_{r=R_m} \quad (3.15)$$

$$H_{\theta I}(r, \theta)|_{r=R_m} = H_{\theta II}(r, \theta)|_{r=R_m} \quad (3.16)$$

Where $R_m = R_s - g$, with notation according to Figure 3.2. Boundary conditions (3.13,3.14) denote that the magnetic field intensity is zero in the iron parts of the machine, which is the case for an infinitely permeable iron core.

Conditions (3.15,3.16) denote the continuity of the radial component of the $\vec{B}$ field, and the tangential component of the $\vec{H}$ field in the boundary between the magnet surface and the airgap.

Applying the boundary conditions, we calculate the scalar potential φ , and after substituting in equations (3.7,3.8), the $\vec{B}$ field components are computed from equations (3.9,3.10). Focusing on the radial component of the $\vec{B}$ field in the airgap region, the following expression is finally obtained:

$$\begin{aligned}B_{rI}(r, \theta) = & \sum_{n=1,3,5}^{\infty} \frac{\mu_0 M_n}{\mu_r} \frac{np}{(np)^2 - 1} \\ & \left\{ \frac{(np-1) + 2\left(\frac{R_r}{R_m}\right)^{np+1} - (np+1)\left(\frac{R_r}{R_m}\right)^{2np}}{\frac{\mu_r+1}{\mu_r} \left[1 - \left(\frac{R_r}{R_s}\right)^{2np}\right] - \frac{\mu_r-1}{\mu_r} \left[\left(\frac{R_m}{R_s}\right)^{2np} - \left(\frac{R_r}{R_m}\right)^{2np}\right]} \right\} \\ & \left[\left(\frac{r}{R_s}\right)^{np-1} \left(\frac{R_m}{R_s}\right)^{np+1} \left(\frac{R_m}{r}\right)^{np+1} \right] \cos np\theta\end{aligned}\quad (3.17)$$

Figure 3.4 presents the slotless machine PM field computed with equation 3.17, in comparison to the PM field obtained with FEM for an actual slotted design, with q=1 slot per pole per phase.

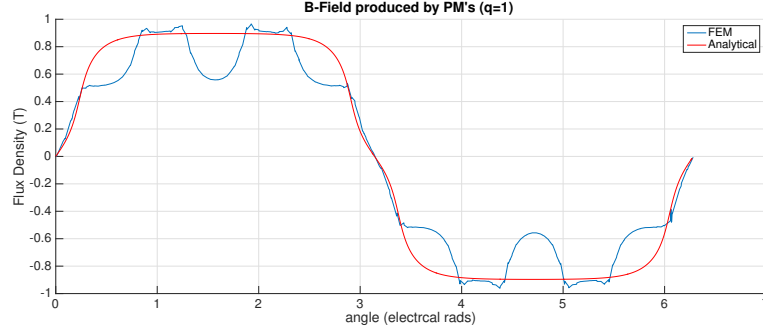


Figure 3.4: Analytical PM (no slotting) VS FEM calculated field

The predicted flat top magnitude is in good agreement between the two waveforms. In the FEM waveform, the effect of slotting is also evident: For $q=1$, we observe 3 dips per pole in the FEM waveform, as expected. The effect of flux focusing in the stator teeth edges can also be observed.

3.3 Magnetic Field due to stator currents

Following a procedure similar to that of section 3.2, the 2-D armature reaction field in polar coordinates is obtained in this section. In addition, the model has been adjusted appropriately to account for both overlapping and non-overlapping winding configurations.

3.3.1 Magnetic Field produced due to a stator coil

To formulate the description of the armature field, we first need to assume a specific current density distribution. A distributed current sheet representation will be used here. According to this, one conductor corresponds to a current density value which is constant along the opening of the slot b_o , as can be seen in Figure 3.5 :

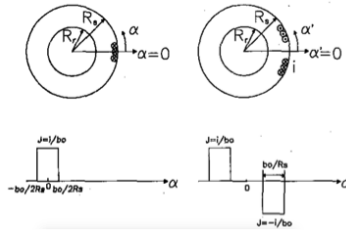


Figure 3.5: Current density distribution [32]

Therefore, the current density can be expressed as a two-branch function:

$$J = \begin{cases} \frac{i}{b_o}, & \text{for } -\frac{b_o}{2R_s} \leq a \leq \frac{b_o}{2R_s} \\ 0, & \text{elsewhere} \end{cases} \quad (3.18)$$

where i is the current through the conductor and a is the angle co-ordinate. The solution of the Laplace equation (3.5) in the airgap region, must be obtained once again, but this time, the associated boundary conditions are the following:

$$H_a(r, \theta)|_{r=R_s} = -J \quad (3.19)$$

$$H_a(r, \theta)|_{r=R_r} = 0 \quad (3.20)$$

Applying the boundary conditions and substituting in equations (3.7-3.10), we obtain the radial component of the $\vec{B}$ field, produced by a single coil in the stator [33]:

$$B(a, r) = \frac{2\mu_0 i}{\pi \delta} \sum_n \frac{1}{n} K_{son} K_{pn} F_n(r) \cos na \quad (3.21)$$

Where :

- $\delta = g + h_m/\mu_{rm}$, is the effective airgap length, when the magnet width is h_m .
- $K_{son} = \frac{\sin(n \frac{b_o}{2R_s})}{n \frac{b_o}{2R_s}}$, is the slot-opening factor, which approaches unity, as the slot-opening width approaches zero.
- $K_{pn} = \sin(\frac{na_y}{2})$, is the winding pitch factor of a winding with coil span a_y .
- $F_n(r) = \delta \frac{n}{r} \left(\frac{r}{R_s} \right)^n \frac{1 + \left(\frac{R_r}{r} \right)^{2n}}{1 - \left(\frac{R_r}{R_s} \right)^{2n}}$, is a function accounting for the effective airgap. If the effective airgap is assumed to be very small, then $r \simeq R_r \simeq R_s$, and $F_n(r)$ approaches unity.

3.3.2 Magnetic Field produced from a 3-phase distributed winding

In the previous section the radial component of the $\vec{B}$ field distribution due to a single coil with one turn, was derived. Therefore, we can derive a similar expression for a distributed phase winding, if we multiply the expression of (3.21), by the number of series turns per phase W , and account for the field attenuation due to the distribution of the phase winding in neighbouring slots:

$$B(a, r, t) = \frac{2W\mu_0 i}{\pi \delta} \sum_n \frac{1}{n} K_{son} K_{dpn} F_n(r) \cos na \quad (3.22)$$

Where:

- $a = 0$ corresponds to the magnetic axis of the phase winding.
- $K_{dpn} = K_{dn}K_{pn}$, is the winding factor, defined as the product of the pitch and the distribution factor. For a machine with Q slots and q slots per pole per phase, the distribution factor is calculated as:

$$K_{dn} = \frac{\sin\left(q\frac{n\pi}{Q}\right)}{q\sin\frac{n\pi}{Q}} \quad (3.23)$$

In equation (3.22) the field in is a function of time, since the current i is also a function of time.

The magnetic field due to a 3-phase winding is now easy to determine, by summing up the $\vec{B}$ of the individual 3-phases:

$$\begin{aligned} B_{winding}(a, r, t) &= B_a(a, r, t) + B_b(a, r, t) + B_c(a, r, t) \\ &= \frac{W\mu_0}{\pi} \frac{2}{\delta} \sum_n \frac{1}{n} K_{son} K_{dpn} F_n(r) \\ &\quad \left[i_a \cos na + i_b \cos n\left(a - \frac{4\pi}{3}\right) + i_c \cos n\left(a - \frac{4\pi}{3}\right) \right] \end{aligned} \quad (3.24)$$

Figure 3.4 presents the slotless machine stator field computed with equation 3.24, in comparison to the one obtained with FEM for an actual slotted design, with $q=1$ slot per pole per phase.

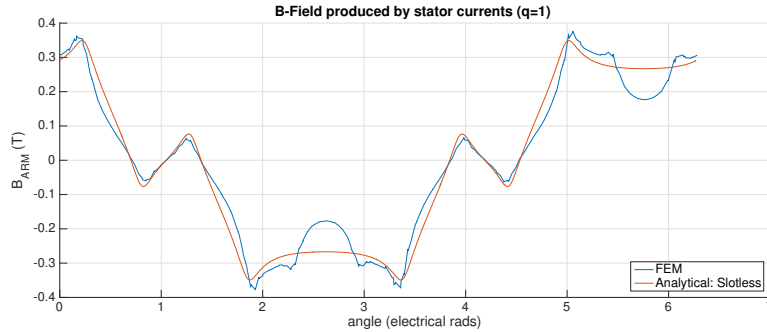


Figure 3.6: Armature field for $q = 1$

The analytically predicted flat top flux density is somewhat lower compared to the FEM value. Nevertheless, the discrepancy is always within 5%.

3.3.3 Magnetic Field produced from a 3-phase FSC winding

In the case of FSC windings, transition from a field produced by single-coil to a full 3-phase winding, is not so straightforward and eq. (3.24) is no longer valid.

To overcome this difficulty, we used a function which applies the star of slots method [star], to find the winding layout of the machine. Using this, we know the phase (a,b or c), direction (positive or negative) and winding angle of the conductors in each slot of the machine. A slightly modified version of equation (3.21) is then used for each coil:

$$B(a, r) = \frac{2\mu_0}{\pi} \frac{i}{\delta} K_{dir} \sum_n \frac{1}{n} K_{son} K_{pn} F_n(r) \cos n(a - \theta) \quad (3.25)$$

Where:

- $K_{dir} = \pm 1$, denotes the winding direction (positive or negative).
- θ , corresponds to the winding angle of each slot.

Summing up for all the coils of each phase, we finally compute the field produced by a phase winding. Figure 3.4 presents the slotless machine stator field computed using equation 3.25, in comparison to the one obtained with FEM for an actual slotted design, with $q=0.4$ slots per pole per phase.

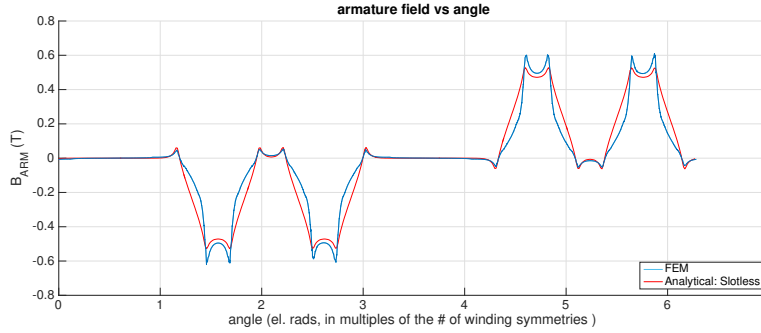


Figure 3.7: Armature field for $q = 0.4$

The working harmonic in FSC windings is not the fundamental. Instead, the wavelength of the B-field in Figure 3.7 depends on the number of winding symmetries, i.e. how many times the main pattern is repeated to form the full machine winding.

3.4 Effect of stator slotting

3.4.1 The relative permeance function

In order to account for the effect of alternation between tooth and slot in the stator inner surface, a relative permeance function is introduced in this section. Following the approach of Zhu and Howe [34], a real permeance function is calculated based on conformal transformation method. The conformal mapping assumes infinitely deep rectilinear stator slots, as shown in Figure 3.8

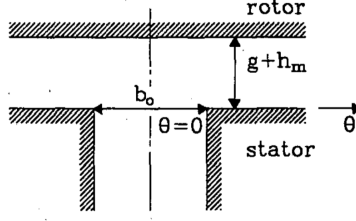


Figure 3.8: Armature field for $q = 0.4$

The permeance function can be expressed in the form of a Fourier series according to:

$$\tilde{\lambda}(r, a) = \sum_{\mu=0}^{\infty} \tilde{\Lambda}_{\mu}(r) \cos \mu Q(a + a_{sa}) \quad (3.26)$$

The angle a_{sa} is needed so that the reference of the permeance function always coincides with the axis of phase A coils. The value of a_{sa} depends on whether the winding pitch is an odd or even multiple of the slot pitch [35]:

$$a_{sa} = \begin{cases} \frac{\pi}{Q}, & \text{odd} \\ 0, & \text{even} \end{cases} \quad (3.27)$$

The first coefficient $\tilde{\Lambda}_0$ in the sum of eq.(3.26), corresponds to the average value of the permeance function. In a 1-D model this is assumed to be unity. In 2-D models, a common choice is $\tilde{\Lambda}_0 = 1/K_c$ [37], where the Carter factor K_c is given by:

$$K_c = \frac{\tau_s}{\tau_s - \gamma g} \quad (3.28)$$

Where:

- τ_s is the machine slot pitch.
- $\gamma = \frac{4}{\pi} \left[\frac{b_s}{2g} \arctan \left(\frac{b_s}{2g} \right) - \ln \sqrt{1 + \left(\frac{b_s}{2g} \right)^2} \right]$ is the γ factor as found in [34].

The Carter factor expresses the reduction in the total flux per pole, as a result of slotting. The total flux per pole in a slotted machine the same to that of the equivalent slotless machine, with an air-gap enlarged by K_c . To include the fringing effect more accurately, a compensation coefficient K_f was also considered [38], resulting in a slight improvement in the waveforms match.

For higher μ values, the Fourier coefficients $\tilde{\Lambda}_{\mu}$ of the permeance function are derived as following:

$$\begin{aligned}
\tilde{\Lambda}_\mu(r) &= \frac{2}{\alpha_t - \frac{\alpha_t}{2}} \int^{\frac{\alpha_t}{2}} \tilde{\lambda}(r, a) \cos \mu Q a da \\
&= -\beta(r) \frac{4}{\pi \mu} \left[0.5 + \frac{\left(\mu \frac{b_0}{\tau_s} \right)^2}{0.78125 - 2 \left(\mu \frac{b_0}{\tau_s} \right)^2} \right] \\
&\quad \sin \left(1.6 \pi \mu \frac{b_0}{\tau_s} \right), \mu = 1, 2, 3 \dots
\end{aligned} \tag{3.29}$$

The function $\beta(r)$ of equation (3.28) stems from the conformal transformation, and is given by :

$$\beta(r) = \frac{1}{2} \left[1 - \frac{1}{\sqrt{1 + \left(\frac{b_0}{2\delta} \right)^2 (1 + v^2)}} \right] \tag{3.30}$$

Where v is obtained by solving the following equation:

$$\frac{y\pi}{b_0} = \frac{1}{2} \ln \left[\frac{\sqrt{a^2 + v^2} + v}{\sqrt{a^2 + v^2} - v} \right] + \frac{2\delta}{b_0} \arctan \frac{2\delta}{b_0} \frac{v}{\sqrt{a^2 + v^2}} \tag{3.31}$$

Where:

- $a^2 = 1 + \left(\frac{2\delta}{b_0} \right)^2$
- $y = r - R_s + \delta$, for an internal rotor machine.

3.4.2 Slotted machine results

Eventually, the new PM and stator fields for the slotted machine, are the product of the fields computed for the slotless machine and $\tilde{\lambda}(r, a)$:

$$B_{PM}(a, r, a_{ma}) = B_{PM-slotless}(a, r, a_{ma}) \tilde{\lambda}(r, a) \tag{3.32}$$

$$B_{arm}(a, r, a_{ma}) = B_{arm-slotless}(a, r, a_{ma}) \tilde{\lambda}(r, a) \tag{3.33}$$

Where the time dependence is now expressed with the rotor position variable a_{ma} . Figure 3.9 presents both slotless and slotted machine analytical PM field in comparison to FEM, for $q=1$.

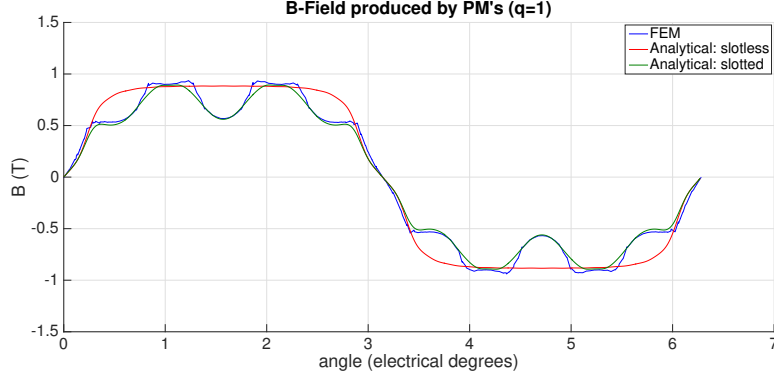


Figure 3.9: Analytical PM (slotless and slotted) VS FEM calculated field

We observe that the RP function is indeed capturing the effect of slotting. A small undershoot can be observed for the slotted model, in the regions across the stator teeth. Using the RP function, the fringing flux near tooth tips is considered, but underestimated. A more accurate but lengthier computation could be made assuming a complex permeance.

Figure 3.10 presents both slotless and slotted machine analytical stator field in comparison to FEM, for $q=1$.

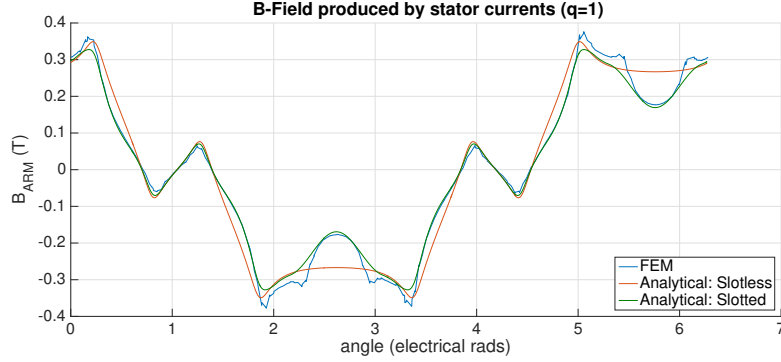


Figure 3.10: Analytical stator field (slotless and slotted) VS FEM - $q=1$

Again, a slightly lower flat-top amplitude is observed for the analytical model. Regarding the slotting effect, conformal mapping is used to find the point in the middle of the slot opening. In this respect, we observe a good agreement with FEM.

For the FSC winding case, Figure 3.11 presents the slotless and slotted analytical stator field in comparison to FEM, with $q=0.4$.

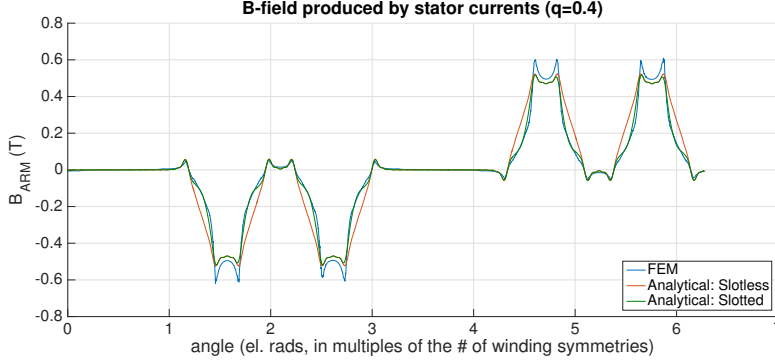


Figure 3.11: Analytical stator field (slotless and slotted) VS FEM - $q=0.4$

Similar remarks to those made for the PM field, apply also for the armature field waveforms. The discrepancies due to neglecting flux concentration on the teeth edges can be observed once again. Besides these, there is a good agreement between the FEM and analytical waveforms.

3.5 Adjusting the radius and slot opening size

In the previous sections, the B-fields of the analytical model were validated with FEM. So far, all fields were computed at the middle of the airgap. The effect of varying the radial direction is depicted in Figure 3.12.

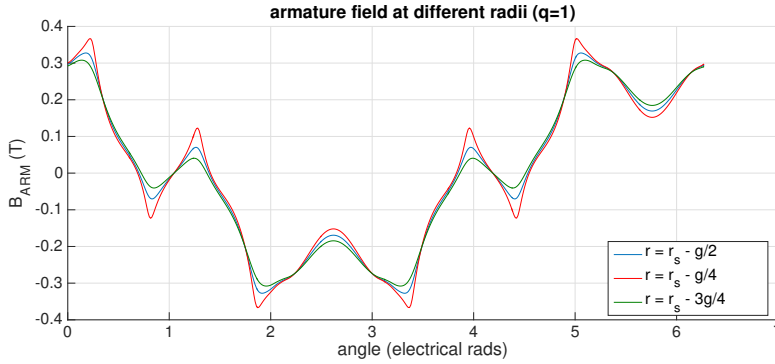


Figure 3.12: Effect of varying the radial direction

We observe that near the stator radius, the spikes in the waveforms grow larger. As we approach the rotor, fewer space harmonics cross the airgap and the waveform becomes more smooth.

The effect of varying the slot opening size is depicted in Figure 3.13

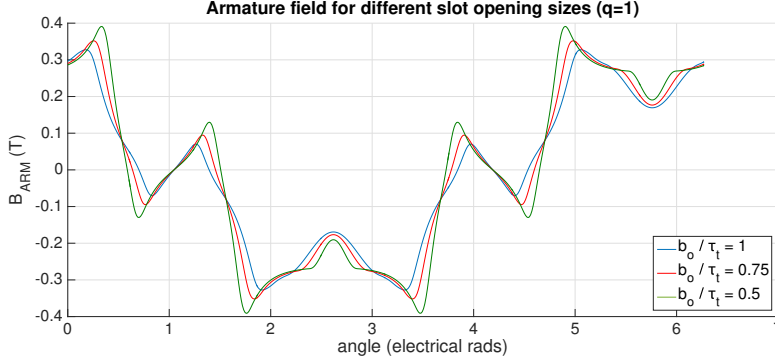


Figure 3.13: Effect of varying the slot opening

Starting with straight slots without tooth tips, the slot opening size b_0 reduces down to half the slot width.

We observe that while the dip width is decreasing as expected, the height of the dip grows larger for lower b_0 values. This could be explained by the fact that the conformal mapping assumes rectilinear slots. Only this simplified slot shape is then also considered for the analytical model.

3.6 Back-EMF and Output Power

To calculate the induced EMF in one phase of the stator coil, we need to express the flux ψ linking the PM field with the stator coil as a function of the rotor position:

$$\psi = \int_{-\frac{\alpha_y}{2}}^{\frac{\alpha_y}{2}} B_{open-circuit}(a, r, a_{ma})|_{r=R_s} R_s l_{ef} da \quad (3.34)$$

Where l_{ef} is the effective axial length of a stator coil. Elaborating on the above expression yields:

$$\psi = \widetilde{\Lambda}_0 \sum_n \frac{\Phi_n}{np} K_{pn} \cos np a_{ma} \quad (3.35)$$

Where:

$$\Phi_n = 2B_n R_s l_{ef} \quad (3.36)$$

Where B_n are the Fourier coefficients of the PM field computed in equation (3.9). Therefore, the induced EMF in one turn of a stator coil, can be derived from Faraday's law:

$$e = -\frac{d\psi}{dt} = \sum_n \omega_r \Phi_n K_{pn} \sin np a_{ma} \quad (3.37)$$

The EMF of a phase winding with W turns, is then expressed as:

$$e = \sum_n W \omega_r \Phi_n K_{dpn} \sin np a_{ma} a \quad (3.38)$$

In the case of FSC windings the winding factor is calculated according to [39]. Once the induced EMF of each phase is calculated, the instantaneous airgap power is derived according to [40]:

$$P = e_a i_a + e_b i_b + e_c i_c \quad (3.39)$$

Keeping the stator field constant, (at $t=0$), we rotate the rotor for one electrical period, to find the angle at which the maximum power occurs. Figure 3.14 presents the analytically computed power as a function of rotor position, in comparison to FEM, for $q=0.4$.

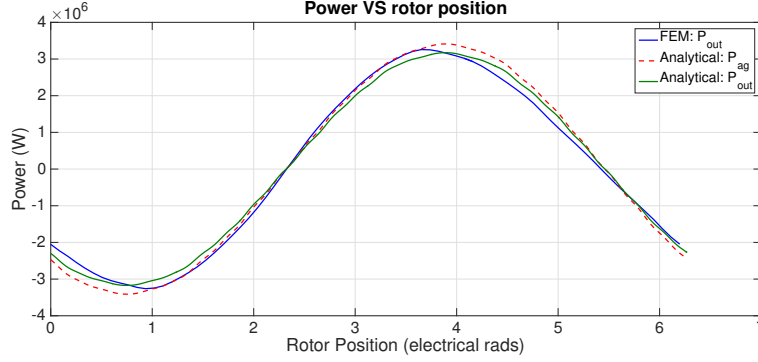


Figure 3.14: Power as function of rotor position - q 0.4

To obtain P_{out} , the losses calculated in section 3.8 were subtracted from the airgap power P_{ag} . We observe that the maximum P_{out} is in good agreement with the FEM value. The small undershoot is a result of the differences in the computed PM fields, mainly due to the limited precision of the RP function.

Setting the initial rotor angle to the maximum power value, we calculate the instantaneous output power for a synchronous rotation under nominal load. Figure 3.15 presents the power components of each phase, as well as the output power in comparison to FEM, for a machine with $q=0.4$:

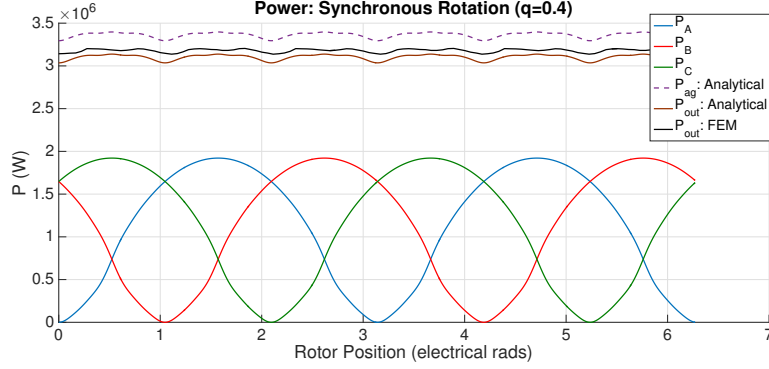


Figure 3.15: Power as function of rotor position: Synchronous rotation

The output power is calculated as the mean values of the P_{out} .

3.7 Torque and Cogging Torque

3.7.1 Torque

When both radial and tangential components of the magnetic flux density are accurately predicted (FEM, subdomain models), the electromagnetic torque is computed using the Maxwell Stress Tensor (MST) [31]. Here, the torque is expressed using the power computed in eq.(3.39), according to [40] :

$$T = \frac{e_a i_a + e_b i_b + e_c i_c}{\omega_r} \quad (3.40)$$

Again, keeping the stator field constant ($t=0$), the torque versus rotor position is presented in Figure 3.16, for $q=0.4$:

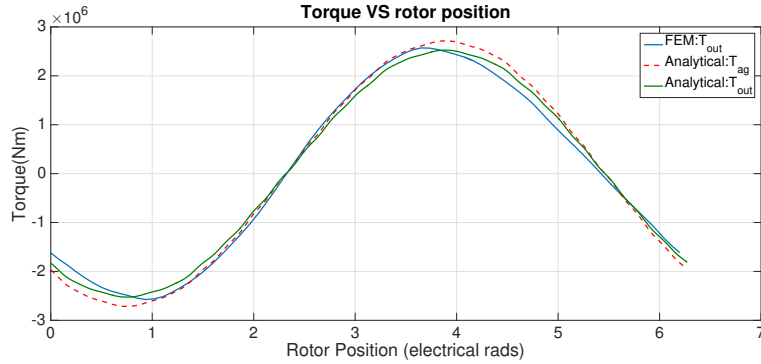


Figure 3.16: Torque as function of rotor position - $q = 0.4$

Similar remarks can be made regarding the peak values of the waveforms. We can observe that the peak value for the torque appears at a load angle of about

90 degrees, as is the case with synchronous, non-salient machines. Following the same procedure, Figure 3.17 depicts the torque components of each phase, as well as the output torque in comparison to FEM, for a machine with $q=0.4$:

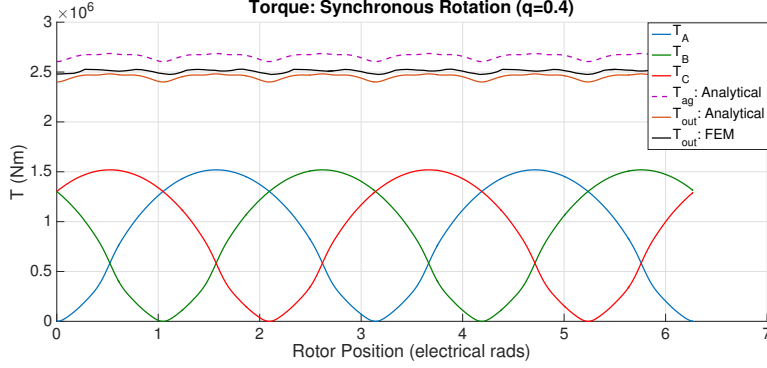


Figure 3.17: Torque as a function of rotor position: Synchronous rotation

The remarks made for 3.15 are also valid here. The machine's output torque will be the mean value of the T_{out} waveform. The torque ripple is also in good agreement.

3.7.2 Cogging Torque

Cogging torque is an important quantity for wind generators. Machines with low cogging can start operation at lower wind speeds.

Various different methods have been proposed for cogging torque calculation. The most accurate ones use the precise radial and tangential field components predicted by subdomain models. In our case, cogging is computed from the energy variation with the rotation angle, according to [31]:

$$T_{cog}(\alpha) = \frac{\partial W_{airgap}(\alpha)}{\partial \alpha} = \frac{\partial}{\partial \alpha} \left[\frac{1}{2 \mu_0} \int_V G^2(\theta) B^2(\theta, \alpha) dV \right] \quad (3.41)$$

Where:

- α is the rotor position,
- θ is the angle along the circumference,
- W_{airgap} is the airgap magnetic energy,
- $G(\theta)$ is the airgap relative permeance,
- $B(\theta, \alpha)$ is the airgap flux density in the slotless machine.

Then, the squares of $G(\theta), B(\theta, \alpha)$ need to be computed. The squared permeance function is expressed in the interval $[-\pi/Q + \alpha, \pi/Q + \alpha]$ corresponding to a slot pitch:

$$G^2(\theta) = G_{a0} + \sum_{k=1}^{\infty} G_{a_{kN_s}} \cos(kQ\theta) \quad (3.42)$$

Similarly the squared flux density is expressed in the interval $[-\pi/P + \alpha, \pi/P + \alpha]$ corresponding to a pole pitch:

$$B^2(\theta, \alpha) = B_{a0} + \sum_{m=1}^{\infty} [B_{a_{mP}} \cos(mP(\theta - \alpha))] \quad (3.43)$$

Substituting the above expressions in (3.41), yields:

$$T_{cog}(\alpha) = -\frac{\pi L_{ef}}{4 \mu_0} (R_m^2 - R_s^2) \cdot \sum_{n=1}^{\infty} n N_L [G_{a_{nN_L}} B_{a_{nN_L}} \sin(n N_L \alpha)] \quad (3.44)$$

Where:

- $R_m = R_s - g$, is the PM outer radius.
- N_L is the least common multiple of P,Q.

Both magnitude and periodicity of the cogging torque are then determined by the choice of P,Q. An analytical expression for the Fourier coefficients based on eq. (3.17), can be found in [31].

Figure 3.18 presents the analytical results in comparison to FEM, for a machine with $q=0.5$.

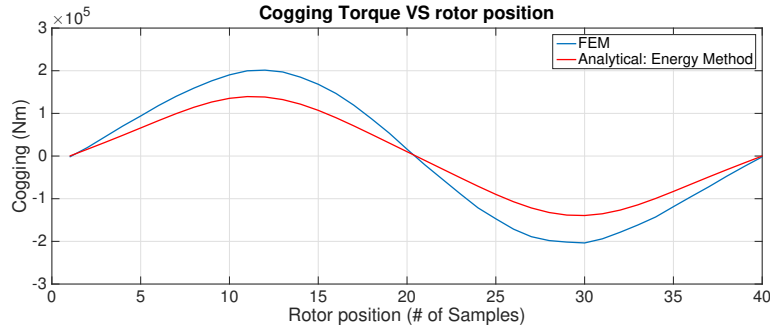


Figure 3.18: Cogging Torque as function of rotor position - $q = 0.5$

The waveforms show a good match in terms of shape, but a significant difference can be observed regarding the peak values. This could be explained by the fact that the analytical model is not considering the energy variation in

the iron and PM regions. The energy associated with the circumferential field component is also neglected.

Finally, Figure 3.19 tests the cogging torque model, for $q=0.4$.

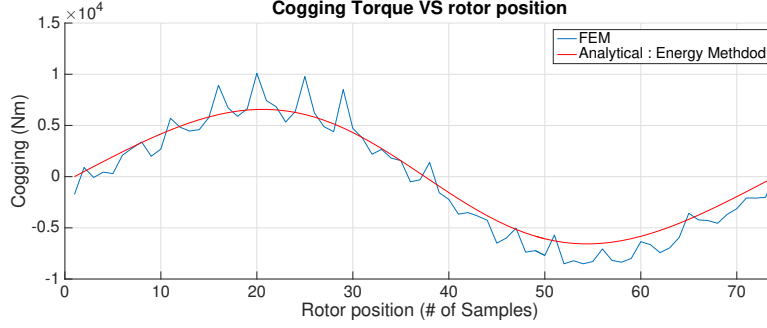


Figure 3.19: Cogging Torque as function of rotor position - $q = 0.4$

This time, there is a significant difference in the shape of the two models, as the FEM waveform shows large fluctuation. A higher number of samples compared to Figure 3.18 was used this time to capture this effect. The peak value however, is in good agreement with the mean value of the FEM waveform.

Overall, the energy-based approach for cogging is less accurate than other techniques (subdomain/lateral force models). Nevertheless it is fast and powerful for optimizing the machine design to minimize cogging .

3.8 Losses

Analytical tools for predicting machine efficiency are presented in the present section. At first, the eddy current and hysteresis losses in the stator iron are computed. Then, the copper loss in the stator windings is predicted. Eddy current losses in the rotor iron and PM's are ignored. As the currents are assumed purely sinusoidal, the respective harmonic loss is also ignored.

3.8.1 Iron Losses

An expression for the for eddy current and hysteresis loss density (loss per unit volume) is commonly employed, in order to calculate iron losses:

$$p_{\text{iron}} = p_h + p_e = k_h B^\beta \omega_s + k_e B^2 \omega_s^2 \text{ W/m}^3 \quad (3.45)$$

Where:

- ω_s , is the angular frequency
- k_h, k_e are hysteresis and eddy current constants, depending on the lamination material.

- β , is a constant typically ranging between 1.8–2.2, depending on lamination.
- B , is the flux density amplitude.

The above expression assumes a sinusoidal flux density distribution, which is not the case in PM machines. The hysteresis loss computation is still valid as it depends only on the peak value of the flux density, if we neglect minor hysteresis loops.

The eddy current loss however, may give a large error especially with FSC windings, due to their high harmonic content. One solution is to consider the average eddy current loss density, which is expressed as [41]:

$$p_e = \frac{2k_e}{T} \int_0^T \left(\frac{d\mathbf{B}}{dt} \right)^2 dt \text{ W/m}^3 \quad (3.46)$$

The eddy-current loss is here assumed to vary with the square of the rate of change of the flux density vector. A tooth component and a yoke component are distinguished and the following assumptions are made [42]:

1. The tooth eddy-current width is independent of the the PM angular width.
2. The tooth flux density is assumed to be approximately uniform.
3. The tooth flux density is assumed to rise linearly, remain constant while the PM passes and then drop linearly to zero.
4. The tooth eddy-current loss is proportional to q and analogous to the square of P .
5. The yoke eddy-current loss is considering only the tangential component of yoke flux density.

Based on the above assumptions, the application of (3.46) is simplified. The tooth eddy-current density is then expressed as:

$$p_{et} = \frac{4m}{\pi^2} q k_e (\omega_s B_{th})^2 \text{ W/m}^3 \quad (3.47)$$

Where $B_{th} = \frac{w_t + w_s}{w_t} B_g$, is the peak flux density in the teeth region. It is expressed as a fraction of the airgap peak flux density B_g (computed by the analytical model), according to the tooth and slot widths w_t, w_s respectively. Similarly, the yoke eddy-current density is expressed as:

$$p_{ey} = \frac{1}{\alpha} \frac{8}{\pi^2} k_e \omega_s^2 B_y^2 \text{ W/m}^3 \quad (3.48)$$

Where $B_y = \frac{w_m}{d_y} B_g$, is the peak flux density in the yoke region. It is expressed as a fraction of B_g , depending to the PM pitch w_m and stator yoke length d_y .

For the hysteresis losses, we make again a distinction between teeth and yoke regions. For the teeth region the hysteresis loss density is given by:

$$p_{ht} = k_h \omega_s B_{th}^\beta \text{ W/m}^3. \quad (3.49)$$

Similarly, the yoke hysteresis loss density is calculated according to:

$$p_{hy} = k_h \omega_s B_{yk}^\beta \text{ W/m}^3. \quad (3.50)$$

Finally, the total iron losses are obtained by summing the eddy-current and hysteresis loss in the teeth and yoke, and multiplying by the respective volumes V_t, V_y , to obtain the actual power loss in Watts:

$$P_{loss,iron} = (p_{et} + p_{ht})V_t + (p_{ey} + p_{hy})V_y \text{ W} \quad (3.51)$$

Figure 3.20 presents the analytically computed losses in comparison to FEM, for a machine with $q=1$.

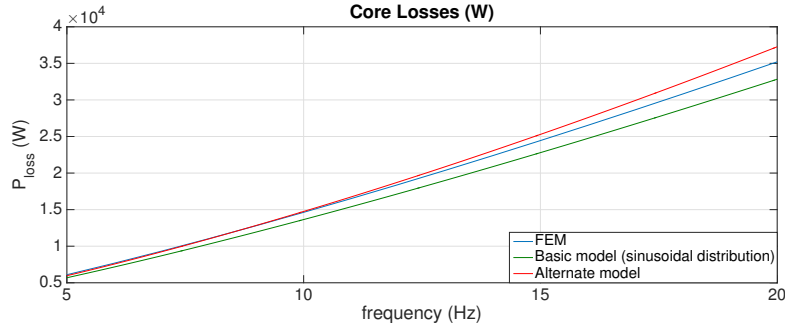


Figure 3.20: Iron losses as a function of frequency - $q = 1$

The waveforms of two analytical modes are depicted: The first, basic model makes use of eq.(3.45), and assumes a purely sinusoidal flux density distribution. The respective losses are consequently quite lower compared to FEM, as the higher harmonics of the flux density field are not taken into account. The FEM calculation of iron loss is also based on eq. (3.45). However, the B-field is calculated much more accurately, at every node of the mesh.

The second, alternate model stems from eq. (3.51) and achieves a better agreement with FEM results. This time, the losses are slightly overcalculated. This is due to the limited accuracy of the assumed flux waveform in the tooth and yoke, as well as the error introduced by only using the magnitude of $\vec{B}$. In addition, geometry effects such as varying the slot pitch, magnet thickness and air-gap length, also influence the flux waveforms.

A third, modified alternate model which accounts for these effects by means of experimentally-derived correction factors, is presented in [42]. Figure 3.21 presents the results of Figure 3.20, with the addition of the modified alternate model:

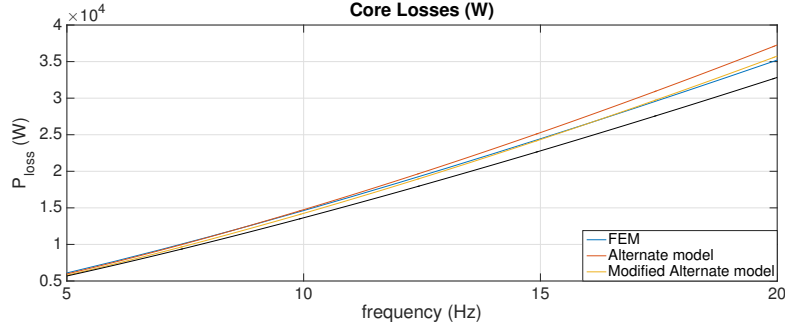


Figure 3.21: Modified alternate model for Iron Losses - $q = 1$

We observe that the modified model is indeed achieving a better match with FEM results. However, the correction factors are determined from experimental curves which vary with geometry parameters. Thus, the second model is finally chosen for the iron loss, as its application is fast and not restricted by varying the geometry.

3.8.2 Copper Losses

Copper losses refer to the ohmic losses due to the current in the copper windings of the machine. As PM machines have windings only in the stator, the copper loss is significantly lower than that of EE machines. At low operating frequencies, as is the case machines Direct-Drive PM machines, copper loss tends to be dominant compared to iron loss. Copper losses are calculated according to:

$$P_{loss,cu} = N_{ph} I_{ph}^2 R_{ph} \quad (3.52)$$

Where:

- N_{ph} , is the number of phases
- I_{ph} , is the phase current
- R_{ph} , is the phase resistance

The phase resistance is assumed to have only a DC component, i.e. skin and proximity effects are neglected. This assumption is valid when the skin depth, expressed in the following equation, is larger than the conductor's diameter.

$$\delta_{skin} = \sqrt{\frac{2\rho_{cu}(T)}{\omega_e \mu_0}} \quad (3.53)$$

Where:

- $\rho_{cu}(T)$, is the resistivity of copper at temperature T.

- ω_e , is the angular electrical frequency.

The low operating electrical frequency results in a large skin depth, which allows us to consider only the DC resistance:

$$R_{ph} = \frac{W \rho_{cu} l_{turn}}{A_{cu}} \quad (3.54)$$

Where:

- $l_{turn} = 2(l_{ef} + \tau_{cm} k_c)$, is the average length of a turn, where τ_{cm} is the coil pitch measured in the middle of the slot and k_c is a factor depending on the type of winding [43]:

$$k_c = \begin{cases} \frac{\pi}{2}, & \text{for distributed windings} \\ 1.36, & \text{for single-layer FSC windings} \end{cases} \quad (3.55)$$

- A_{cu} is the cross-sectional area of the conductor. Since we don't know the conductor radius, it is calculated by dividing the slot copper area over N_c .

Substituting eq.(3.54) into eq.(3.52), we finally obtain the copper loss of the machine winding. The results obtained with the analytical model are compared to FEM, for $q=1$, in the following table:

Table 1: Copper Loss - $q=1$

Model	Copper Losses (kW)
Analytical	239.40
FEM	236.99

We observe that the copper losses calculated by the two models are in good agreement, with an error of about 1%. This result should be viewed as a cross-check: The FEM calculations are identical to those presented above, with the exception of different modeling of the conductor length in eq.(3.55). We also note that the copper loss is considerably higher compared to the iron losses. This is explained by the low operating frequency of the machine, which keeps the eddy current and hysteresis losses limited.

3.9 Summary

This chapter presented a 2-D analytical model for calculating for a RFPM synchronous machine. The flux density fields were initially predicted for a slotless machine, and a relative permeance function was then used to account for slotting. Subsequently, important quantities for evaluating the performance (Power, Torque, Torque ripple) and efficiency (Iron loss, Copper loss) of the machine were computed.

The model was validated with FEM, for both distributed and FSC winding topologies. The accuracy of the analytical model was satisfactory at most occasions. A slight undershoot was observed in the output power, resulting from the

inaccuracies of the relative permeance function. In the cogging torque model, a significant error can occur in the peak value, for certain pole-slot combinations. More sophisticated models (subdomain/lateral force) are needed to get accurate results.

Overall, the presented model lacks in accuracy compared to the recently developed subdomain models. These however, are far more time-consuming, as they must solve large systems of linear equations at each rotor position [38]. Thus, the speed and simplicity of the presented model make it advantageous for problems which require a large number of iterations. Such a problem is the investigation of several machine topologies, which will be presented in the next chapter.

4 Comparison of modular winding machines by means of analytical modeling

The present chapter presents a first level comparison between different modular machine topologies. Different slot/pole combinations of modular winding designs are compared. A varying axial length is assumed to maintain the same speed and airgap power in all machines. The base case of distributed winding with $q = 1$ is also considered as a reference. The validated analytical model developed in Chapter 3, is used here to promote the topologies which perform the best, based on energy yield, efficiency and cost .

4.1 Design choices for comparison

A 10MW machine described in [44] was used as a starting point for the comparison. The main dimensions of the original, distributed winding machine are given in Table 2:

stator radius r_s (m)	5
stack length l_s (m)	1.6
number of pole pairs p	160
number of slots per pole per phase q	1
air gap g (mm)	10
stator slot width b_{ss} (mm)	16.4
stator tooth width b_{st} (mm)	16.4
stator slot height h_{ss} (mm)	80
stator yoke height h_{sy} (mm)	40
rotor yoke height h_{ry} (mm)	40
magnet height l_m (mm)	20
rotor pole width b_p (mm)	80

Table 2: Geometry parameters for the original, distributed winding machine

To obtain comparable designs, all machines were chosen to have:

- the same rated airgap power P_{ag} .
- the same rated mechanical speed (10 RPM).
- the same stator radius.
- varying axial length. This is left as an adjustable degree of freedom in order to meet the airgap power requirement.

Regarding other design choices, fixed geometric ratios (such as the magnet-to-pole pitch ratio) are used to maintain a constant iron and PM area. In addition, the fill factor was constant for all machines, although a higher value may be achieved in practice for FSC windings [45].

Finally, the back iron thickness is fixed in both stator and rotor. The thickness of the stator yoke could be reduced with increasing pole numbers, since the flux per pole becomes lower [46]. However, for reasons of simplicity and mechanical robustness, it is kept constant.

4.2 Comparison Criteria

The quantities computed in Chapter 3, are now used to form the comparison criteria, as described subsequently:

1. The **annual energy yield**. This is computed by calculating the machine's output power at various wind speeds. The output power increases with the cube of the wind speed until the rated value (12m/s), as depicted in Figure 4.1 :

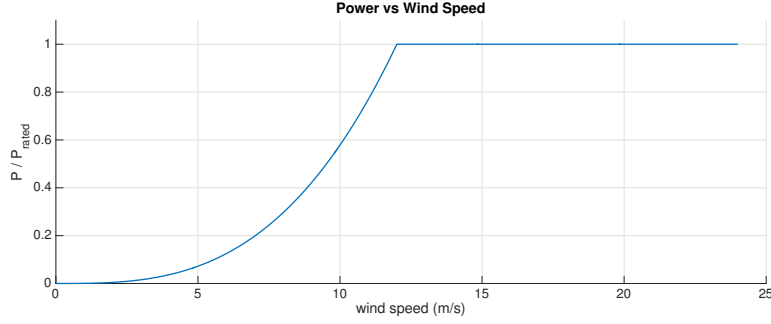


Figure 4.1: Power as a function of wind speed

Above the rated wind speed, the output power and mechanical speed remain constant by means of pitch control. To obtain the energy yield, a wind speed pattern must be selected. A Weibull distribution with an average speed value of 10m/s is chosen here, as depicted in Figure 4.2:

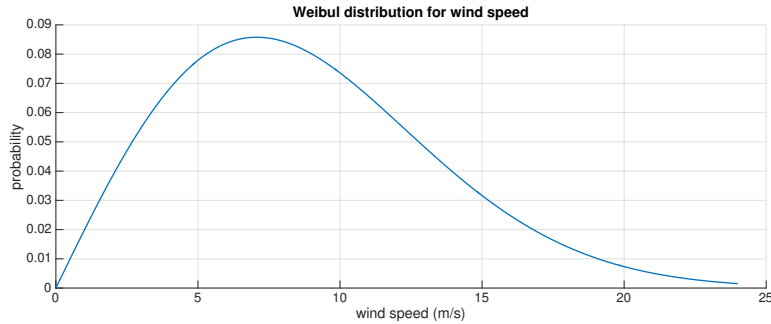


Figure 4.2: Weibull probability density function for wind speed

The probability for each wind speed is subsequently expressed in hours/year and multiplied with the respective output power. Summing up for all wind speeds we obtain the annual energy yield.

2. The **efficiency** of the machine. This is computed at rated operation, based on the iron and copper losses computed in Chapter 3.
3. The **cost** of the machine. This includes the cost of the iron, copper and magnets. It is computed by calculating the respective volumes and multiplying by the material's density and price per kg.

4.3 Choice of slot pole combinations

The initial machine has $p = 160$ pole pairs with $q = 1$. The pole number is then chosen to vary between 150 – 170 pole pairs. The aim is to find all possible slot-pole combinations resulting in FSC winding machines, within the specified pole range. Our investigation is limited to single-layer windings, as they are more suitable for modular configurations and exhibit better fault-tolerant characteristics.

Three-phase machines with FSC winding have a number of q slots per pole per phase less than or equal to 0.5 [47]. A common lower boundary for FSC windings is $q = 0.25$, corresponding to 3 slots for 2 pole-pairs [48]. In order to obtain all possible slot-pole combinations, the following procedure was followed:

1. Begin with the lower limit pole number $P = 300$ (150 pole pairs).
2. Starting from $q = 0.25$, increase q with a step of $2/P$, until we reach $q = 0.5$. This corresponds to adding 6 slots every time, which is the minimum number of slots required in a single-layer winding. This way, we determine all possible q values for the given pole number.
3. Repeat step 2 for each pole number.

Finally, the following limitations must be also taken into account: The number of slots needs to be an even integer. For a balanced FSC winding, it must also satisfy the following condition:

$$\frac{Q}{[\text{GCD}(Q, 2p)]} = 3k \quad (4.1)$$

Where k is integer.

Following the described procedure, we obtain a map of possible slot-pole combinations, depicted in Figure 4.3:

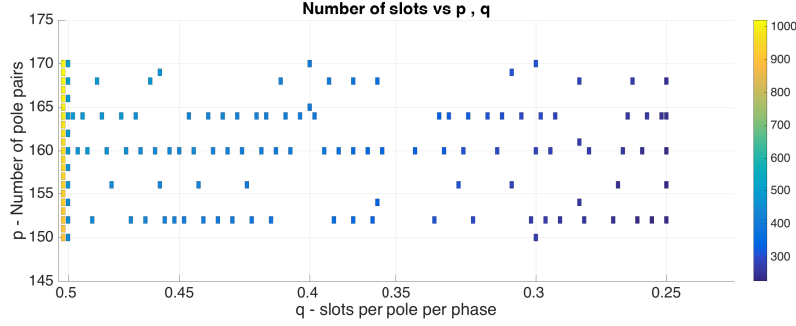


Figure 4.3: Map of slot-pole combinations

A total of 138 different slot-pole combinations are depicted in the map. Blue colored bars correspond to low slot numbers Q , whereas yellow bars correspond to high Q . The q axis ranges from $0.25 - 0.5$, as explained previously. The base case of distributed winding with $q=1$ is also included for reference and comparison (yellow bars).

The results of the topology investigation for each specified criterion, are presented in the following sections.

4.3.1 Annual Energy Yield

Computing the annual energy yield is based on the calculation of the machine's output power at various wind speeds. To maintain the same airgap power, the axial length of the machines was adjusted, as depicted in Figure 4.4:

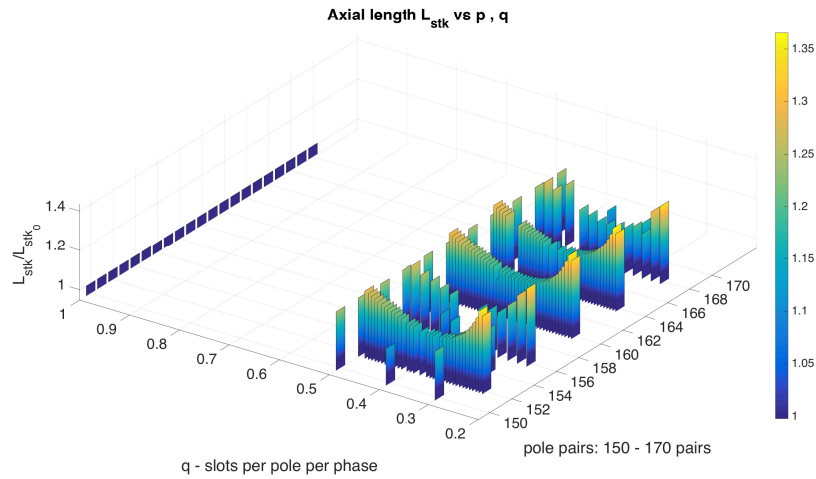


Figure 4.4: Adjusting the axial length - L_{stk} as a ratio of the initial length

We observe that FSC winding machines may need to be up to 35% lengthier

compared to the initial machine ($q=1, p=160$), in order to produce the same airgap power. This is a result of the lower winding factors obtained with FSC windings.

Subtracting the losses from the airgap power, the output power is derived and used for the energy yield calculations. The results for the energy yield are presented in Figure 4.5:

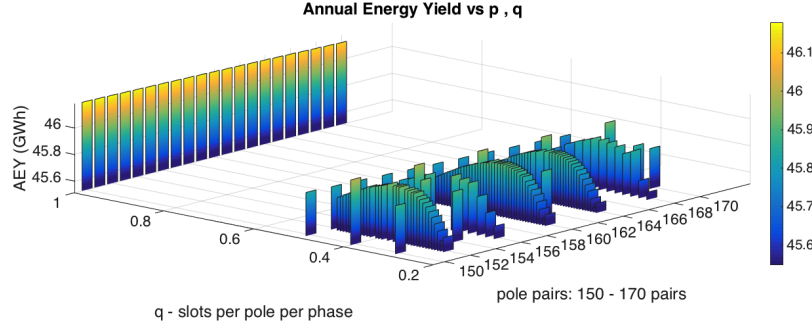


Figure 4.5: Annual energy yield (GWh) vs p, q

Machines with q values between $1/2$ and $1/3$, result in the highest energy yield. In particular, $q = 0.4$ results in the highest yield among FSC winding designs, followed by $q = 0.3355$ and $q = 0.3618$.

Keeping q constant, a total drop in the order of 0.5% is observed in the energy yield, as we move from 150 to 170 pole pairs. Since the airgap power is the same for all machines, this result implies that stator losses increase with the pole number. This will be confirmed in the next subsection.

4.3.2 Efficiency

Figure 4.6 depicts the results for the core losses at rated wind speed:

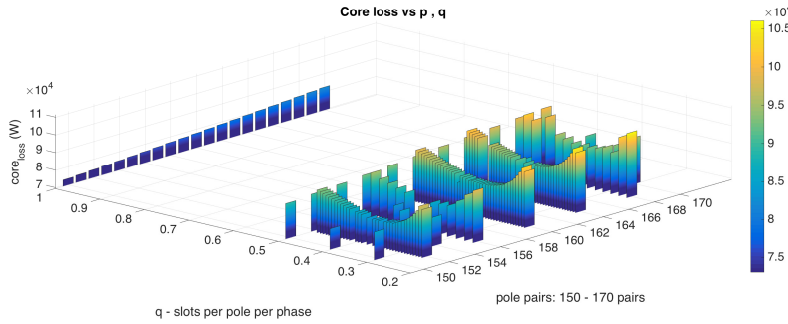


Figure 4.6: Core Loss vs p, q

A clear trend can be observed here: As the pole numbers increase, the

electrical frequency increases and therefore the core losses are rising. Figure 4.7 presents the results for the copper losses at rated wind speed:

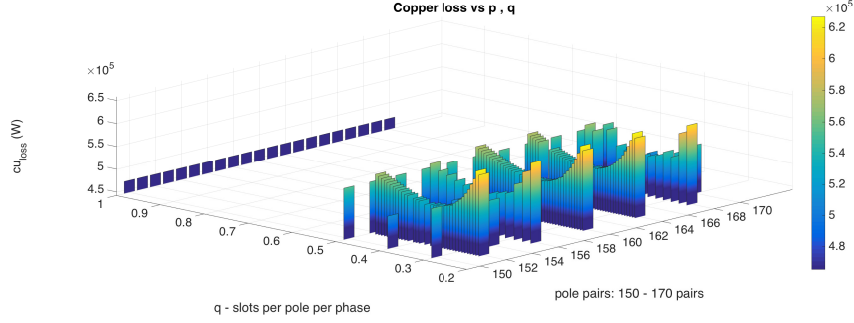


Figure 4.7: Copper Loss vs p,q

We observe that the computed copper losses are significantly higher than the iron losses. This is expected for the low-speed machines of our study. At higher frequencies the balance is reversed and the iron losses tend to become dominant.

We also note that the copper losses remain virtually unaffected, when varying the pole number for the same q . In fact, a very small drop is observed, probably due to the limited precision in the calculation of the end windings length from $eq.(3.55)$.

As copper loss is directly proportional to the conductor length, the graph follows a similar trend to that of the machine axial length. Finally, the overall efficiency results accounting for both iron and copper loss, are depicted in Figure 4.8:

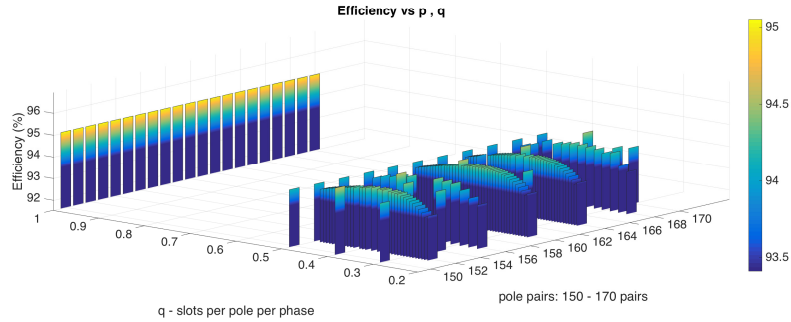


Figure 4.8: Efficiency vs p,q

As expected, the distributed winding machine achieves the highest efficiency, in the order of 95%. Regarding FSC designs, the highest efficiency is achieved for $q = 0.4$.

We observe that the overall efficiency slightly reduces for larger pole numbers, as a result of higher core loss. However, the overall reduction hardly reaches 0.5%. A small change was indeed expected, as the main loss contribution comes from the copper loss which remains virtually unaffected by varying p .

4.3.3 Cost

Figure 4.9 presents the comparison results regarding the overall cost of each topology:

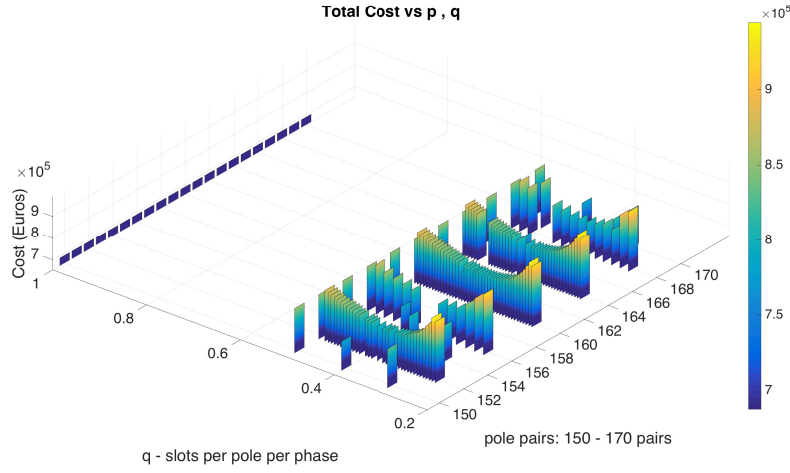


Figure 4.9: Cost vs p,q

Keeping q constant, the total cost effectively the same for different pole numbers. The axial length variation is reflected in the total cost of the machine.

The individual results for the Copper, Iron and PM costs of each topology are presented in Figures 4.10,4.11,4.12:

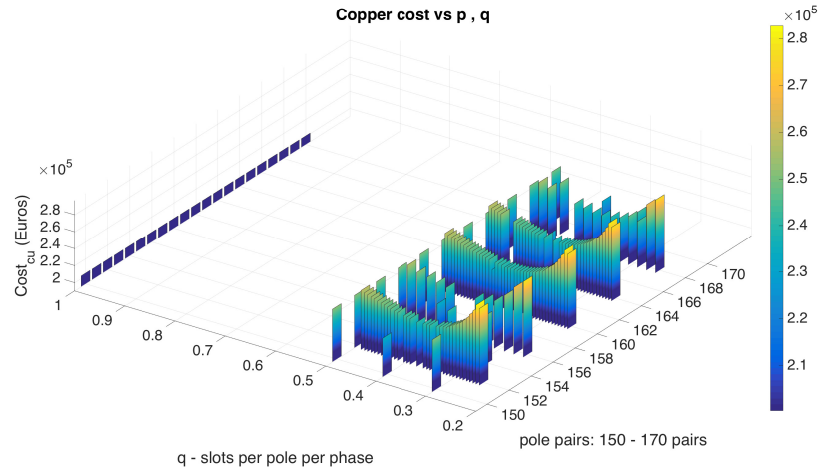


Figure 4.10: Copper cost as function of p,q

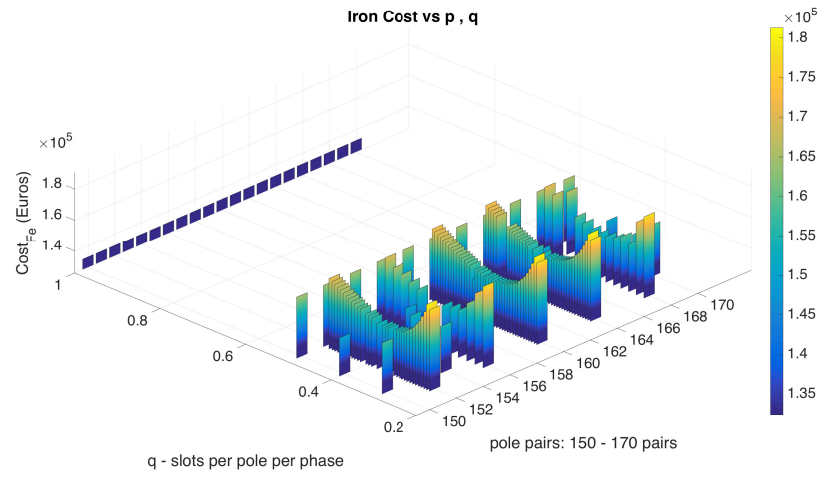


Figure 4.11: Iron cost as function of p,q

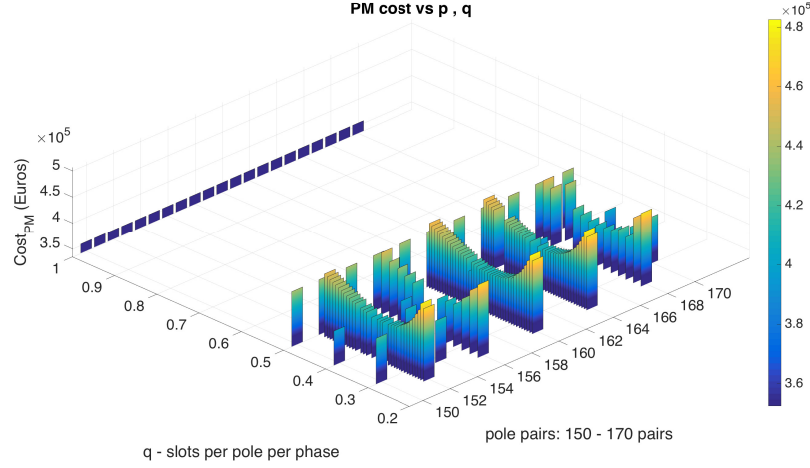


Figure 4.12: PM cost as function of p, q

Again, the axial length variation is reflected in each individual cost of the machine. We observe that keeping q constant, the iron and PM costs remain effectively constant while varying p . This is expected as the total iron and PM areas are kept constant. The PM cost makes up for about 50% of the total cost, due to the high price of PM rare earth materials.

4.3.4 Conclusions - Final choice of topologies

The main conclusions drawn from the comparison results are summarized below:

- Distributed winding topologies perform better compared to FSC winding, mainly due to higher winding factor.
- For the examined range, decreasing the pole number seems an attractive choice, as it leads to a slight increase in efficiency and energy yield. The lower core loss for lower electrical frequencies is the main reason behind this trend.
- The topology with $q = 0.4$ slots per pole per phase achieves the highest yield among FSC winding designs. On top of that, it can be combined with several pole number values, including the lowest pole pair value $p = 150$.
- Other promising q values are $q = 0.3355$, $q = 0.3618$ and $q = 0.3882$. Their performance is very close to that of $q = 0.4$, and they can be combined with different pole number choices.
- For $q < 0.3$, the resulting designs exhibit only a slight drop in energy yield and efficiency. However, the total cost can grow 30% higher compared to the initial machine. Therefore, this range of low q values corresponds to the least attractive topologies.

Based on the above, the final choice of machines to be promoted for segmentation, is presented in Table 3:

Pole pairs (p)	Slots per pole per phase (q)	Slots (Q)
150	0.4	360
152	0.3355	306
152	0.3618	330
160	0.4	384

Table 3: Topologies promoted for segmentation

4.4 Summary

The present chapter presented a topology comparison over a large number of machines with different slot-pole combinations. A 10 MW machine with distributed winding was used as a starting point. The axial length of each machine was left to vary so as to meet the requirement for equal airgap power. In this way, differences in machine performance are reflected to the machine costs and efficiency.

The present comparison focused more on finding every possible q value in a limited range of pole numbers. Alternatively, a higher range of pole numbers could have been considered in the comparison, while addressing only a few typical q values.

The resulting FSC winding topologies are not optimized. Optimization of a large number of machines would require a lot of time and computational effort. A possible alternative approach would be to address fewer but optimized topologies (e.g. with a fixed pole number). This would result in a less broad but more accurate comparison.

Although cogging torque was calculated in Chapter 3, it was not considered in the comparison. This is because in practice, cogging is minimized by means of skewing.

Finally, four different FSC winding topologies were selected. In the following chapter, these will be subjected to different types of segmentation, to determine the optimal segmentation technique and examine the effect of the introduced flux gaps in the stator.

5 Comparison of segmented machine designs

5.1 Design choices and Comparison criteria

In Chapter 4, a first-level comparison of 10MW machines with modular windings was carried out. The present chapter aims to test each promoted machine with different types of segmentation, by means of FEM modelling. The final choice of machines promoted for segmentation, is repeated in Table 4:

Pole pairs (p)	Slots per pole per phase (q)	Slots (Q)
150	0.4	360
152	0.3355	306
152	0.3618	330
160	0.4	384
160	1	960

Table 4: Topologies promoted for segmentation

In particular, a segmented stator with either E-core or segmental teeth modules is tested for each FSC winding machine. Segmentation with distributed winding arrangement is also included in the comparison. In this case however, only 3-phase E-core segments can be used as building blocks. The optimal segmentation technique is determined based on machine output power, efficiency and cost.

Again, the fill factor of the machines is kept constant for simplicity, although higher values may be achieved in practice for segmental teeth designs [6]. A uniform flux gap size of $4mm$ is initially assumed for all machines. Subsequently, the effect of the flux gap size on output power and torque ripple, is also addressed.

Following the same reasoning as in Chapter 4, a varying axial length is considered, in order for FSC winding designs to meet the power requirement set by the initial $q=1$ machine. This way, all machines have the same airgap power, and differences in the output power stem only from the loss computations.

5.2 Comparison and Results

Figures 5.1, 5.2 present the geometry and mesh of two E-core and segmental teeth FEM models respectively. The machine airgap as well as the additional flux gaps introduced by segmentation, are highlighted with a blue colour for each segment design:

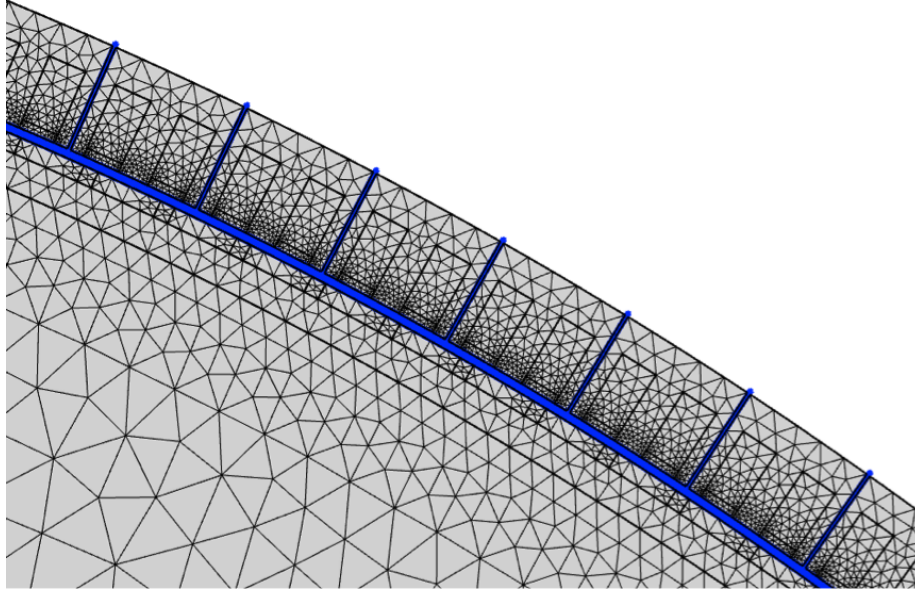


Figure 5.1: E-core segmented stator geometry: $p = 150, q = 0.4$

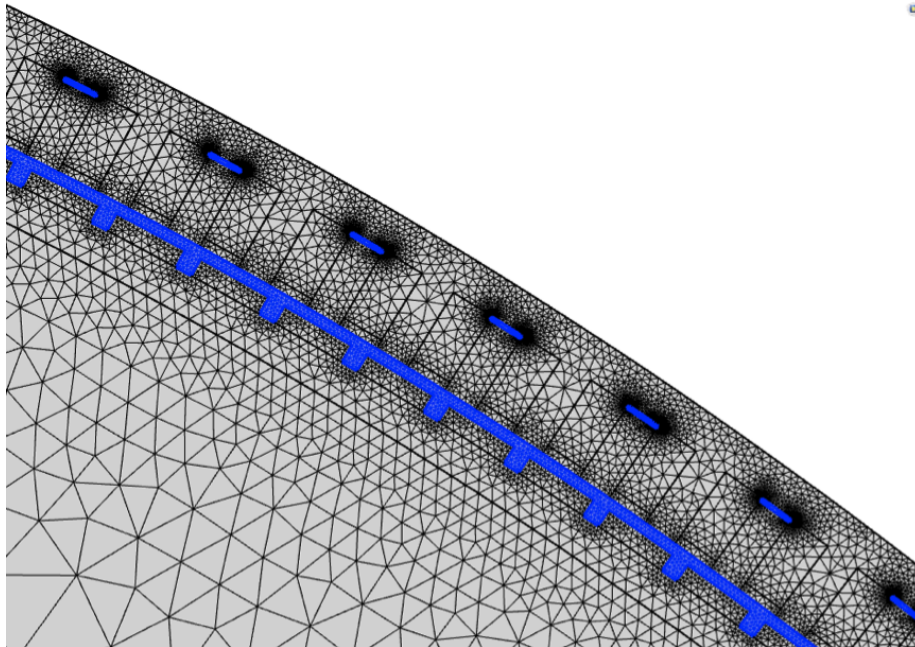


Figure 5.2: Segmental teeth segmented stator geometry: $p = 150, q = 0.4$

In the FSC winding machine of Figure 5.1, each E-core segment corresponds to one stator coil. For distributed winding machines, larger three phase E-cores are used instead. This design offers a lower number of flux-gaps, at the expense of reduced manufacturability.

Similarly, Figure 5.2, corresponds to the segmental teeth design. In this case a faulted coil can be replaced without the need of replacing the back iron. It can be seen that only half of the stator teeth are segmented. This is because in a single-layer FSC winding, only half of the teeth carry a tooth-wound coil.

Consequently, both segment designs can be used to obtain fault tolerance in case of a single coil fault.

In both figures, we observe that the mesh becomes very fine in the blue coloured areas, corresponding to the airgap and flux gap regions. A particularly fine mesh is used in the edges of the segmented teeth in Figure 5.2. High saturation is expected in these regions due to the flux gaps.

5.2.1 Output Power

Table 5 and Figure 5.3 present the output power results for the segmented machines, with a $4mm$ flux gap. The results for the respective non segmented models are also included for completeness:

Topology	p - q	E-core	Segmental teeth	No segmentation
1	150 – 0.4	11.12	10.07	11.61
2	152 – 0.3355	11.09	9.96	11.56
3	152 – 0.3618	11.03	9.70	11.44
4	160 – 0.4	10.96	9.68	11.41
5	160 – 1	11.19	—	11.72

Table 5: Output Power (MW) for the segmented machines

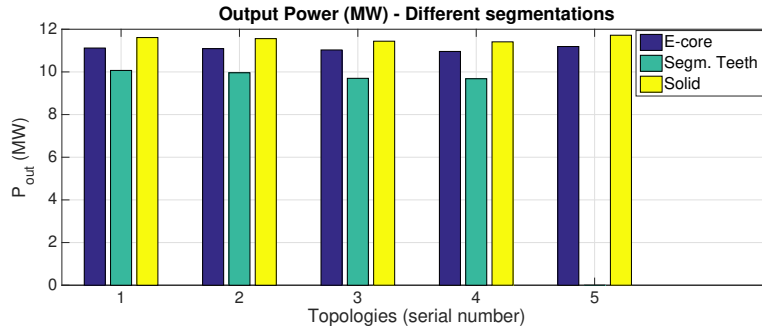


Figure 5.3: Output Power (MW) for the segmented machines

As explained earlier, only (3-phase) E-core segments are applied for the distributed winding machine. For the FSC winding machines, we observe that

in all cases the output power is about 15% lower in the case of segmental teeth design. A power reduction of about 5% is observed for E-core designs, compared to the non segmented machines.

5.2.2 Efficiency

Similarly, Table 6 and Figure 5.4 present the efficiency results for the segmented machines, with a 4mm flux gap.

Topology	p - q	E-core	Segmental teeth	No segmentation
1	150 – 0.4	94.61	93.11	94.67
2	152 – 0.3355	94.36	92.84	94.39
3	152 – 0.3618	94.31	92.66	94.36
4	160 – 0.4	94.30	92.65	94.35
5	160 – 1	94.80	—	94.89

Table 6: Efficiency (%) for the segmented machines

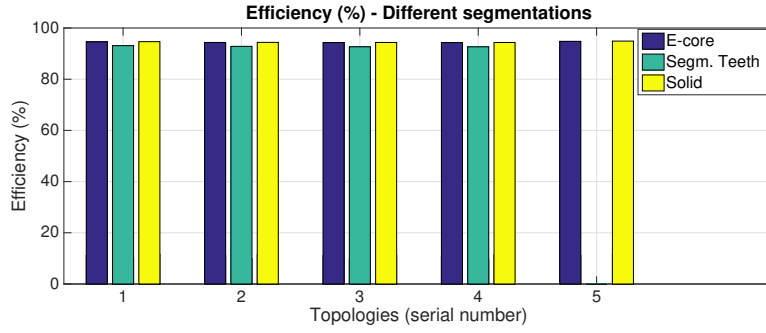


Figure 5.4: Efficiency (%) for the segmented machines

The efficiency is in the range of 93-95%, similar to that of the analytical model in Chapter 4. The efficiency of E-core models is very close to that of non-segmented machines.

Segmental teeth designs result in about 1.5% lower efficiency compared to E-cores. This is a direct consequence of the lower power obtained with segmental teeth designs. In addition, higher core losses are observed in segmental teeth designs, due to saturation in the tooth edges in the vicinity of the flux gaps.

At lower pole numbers, lower electrical frequencies result in slightly better efficiencies, due to reduced core losses.

5.2.3 Cost

Subsequently, Table 7 and Figure 5.5 present the computed cost for the segmented machines, with a 4mm flux gap.

Topology	p - q	E-core	segmental teeth	No segmentation
1	150 – 0.4	822.71	825.34	818.61
2	152 – 0.375	851.84	854.41	847.61
3	152 – 0.3618	866.14	868.13	861.84
4	160 – 0.4	872.02	874.64	867.68
5	160 – 1	807.78	–	803.74

Table 7: Cost results (kEuros) for the segmented machines

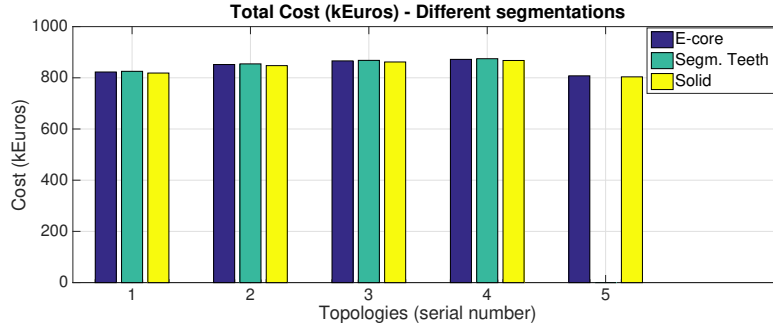


Figure 5.5: Cost results (kEuros) for the segmented machines

We observe that the cost of FSC winding machines is higher compared to the distributed winding machine. Again, the difference comes from the increased axial length required to meet the power requirement. This increases all individual costs (Copper, Iron, PM) in the machine.

Moreover, segmental teeth designs result in slightly higher cost compared to E-cores. This discrepancy implies that a slightly lower amount of iron is removed in segmental teeth designs. The iron removed from segmented models results in reduced cost compared to the solid, non segmented machines. This result could be misleading, as the additional manufacturing costs of segmentation have not been considered.

5.2.4 Conclusions

The main conclusions drawn from the comparison results are summarized below:

- For FSC winding machines, a longer machine is needed to produce the same airgap power, due to reduced winding factor. This is reflected in the cost calculations: The FSC winding machines have about 7-8% higher cost due to their increased axial length.
- Segmental teeth machines lag behind E-core designs with about 15% lower output power. Placing the flux gaps in this position downgrades performance, as it increases the total reluctance of the magnetic circuit.

- Lower efficiencies of about 1.5% are obtained for the segmental teeth models compared to E-cores. Higher core losses are observed in segmental teeth designs, due to saturation in the tooth edges in the vicinity of the flux gaps.
- Among E-core designs, the highest efficiency is achieved with the $p = 150, q = 0.4$ model. Having the lowest pole number, the model achieves a slightly reduced core loss.
- The winding pattern shows symmetries only in the case of $q = 0.4$ and $q = 1$. In the rest of the designs, sectioning cannot be employed in FEM and the whole machine must be simulated.
- The topologies with $q = 0.3618$ and $q = 0.3355$ are almost equally attractive to $q = 0.4$ in terms of output power and efficiency. However, the well documented fault-tolerant and low cogging characteristics of the latter, make it the most attractive choice overall.
- A slightly higher cost was observed for segmental teeth designs. The difference is very small and is mainly due to less iron being removed during segmentation.

5.3 Effect of the flux gap size on machine performance

So far, the size of the stator flux gaps was kept constant at $4mm$. The effect of the flux gap width on the average output power as well as power ripple, will be addressed in the following sections.

The E-cored topology $p = 150, q = 0.4$, which achieved the highest efficiency among the compared designs, is subsequently tested with a varying flux gap width. The resulting segmented machines are compared to each other as well as to the original non segmented design.

5.3.1 Effect on output power

Table 8 presents the effect of flux gap size on the output power:

Flux gap width (mm)	Output Power (MW)	Power reduction (%)
0	11.61	—
2	11.28	2.84%
4	11.12	4.22%
5	10.88	6.30%

Table 8: Output power for various flux gap widths

We observe that increasing the flux gap results in a reduction in the mean output power of the machine. The highest degradation of about 6.5% was noticed for the largest gap size of $5mm$. The reason for this degradation, is

the defocusing effect of the magnetic flux, introduced by the flux gaps. This results in reduced flux linkage, back-EMF and eventually output power for the machine.

As reported in [20] however, this relation does not form a general rule but depends on the slot-pole combination: When $Q > P$, the defocusing effect is observed and performance is degraded. The tested E-cored topology ($p = 150, q = 0.4$) does indeed fall into this category.

Instead, when $Q < P$, an opposite flux focusing effect is observed and machine performance may eventually benefit from segmentation. This condition however corresponds to slot-pole combinations with $q < 1/3$. This range of q values was found to lag behind other candidates in the first-level comparison of Chapter 4.

5.3.2 Effect on power ripple

Finally, Table 9 presents the effect of flux gap size on the output power:

Flux gap width (mm)	Power Ripple (%)
0	0.16
2	0.22
4	0.99
5	1.63

Table 9: Power ripple for various flux gap widths

We observe that increasing the flux gap results in a higher power ripple for the machine. As expected, the highest degradation in performance was noticed for the largest gap size of $5mm$. It has been reported in [20] that certain (small) gap sizes may even slightly reduce the ripple. In our case, a minimum size of $2mm$ was assumed for a uniform flux gap, which already corresponds to one fifth of the machine airgap ($10mm$). At any case, very low ripples below 2% are achieved for the $q = 0.4$ topology, regardless of the gap size.

5.4 Summary

The present chapter carried a comparison of machines with different types of segmentation. The machines promoted from the first-level comparison were now tested with E-core and alternate teeth segmentation, and compared with respect to efficiency and cost. The results certified that performance of segmental teeth designs lags behind due the higher reluctance of the resulting magnetic circuit.

The design which performed the best among candidates, was the E-cored machine with $p = 150, q = 0.4$. This was subsequently simulated for different flux gap widths. The results revealed a degradation in the output power due to the flux gaps, as was expected for the particular choice of p, q . The question that remains, is whether the energy lost due to segmentation of the machine, is worth more than the savings

The power ripple increased with the flux gap for the case tested. It has been reported that certain very low flux gap values may actually reduce the ripple. It remains a question however, how small a uniform flux-gap can be achieved in such big machines. Finally, another option in the design of segmented machines is to optimize the flux gap size, with the objective of minimizing the mutual coupling between phases.

6 Conclusion

6.1 Conclusions

This thesis work has looked at modelling and comparison of modular machine topologies for large wind generators. The main conclusions drawn are presented subsequently:

6.1.1 Theoretical work

As a first step, a literature review on the various aspects of modularity was conducted. Focusing on the machine-side of modularity, machines with modular windings as well as machines with segmented cores were considered. The radial-flux PM machine was confirmed as the most suitable generator topology for modular low-speed wind generators. Finally, specific design choices that result in fault-tolerant topologies with excellent magnetic isolation were highlighted.

6.1.2 Analytical modelling

Subsequently, the thesis focused on the development of a 2-D analytical model for the radial-flux PM machine. The model is able to predict the PM and armature magnetic field distribution for both distributed and FSC windings. The effect of slotting is taken into account by means of a relative permeance function. A very good agreement between analytical and FEM predicted B-fields was achieved. It was observed that the main source of discrepancy stems from the relative permeance function, which cannot fully account for the effect of flux concentration on the edges of the stator teeth.

In a next step, the model computed important quantities to evaluate electromechanical performance (Back-EMF, Power, Torque, cogging torque) as well as efficiency (iron loss, copper loss) of the machine. Again, a good match with FEM results was obtained, with the exception of cogging torque: It was found that it is much more difficult to predict the cogging torque accurately by using relative permeance models. When precise prediction is necessary, the recently developed subdomain models are the most suitable.

Overall, the presented relative permeance model lacks in accuracy compared to subdomain models. These however, are far more time-consuming, as they must solve large systems of linear equations at each rotor position. Thus, the speed and simplicity of the presented model make it advantageous for problems which require a large number of iterations, such as the comparison of a large number of different machines.

6.1.3 First-level comparison

Subsequently, the thesis drew a first level comparison between different modular winding machines. Different slot/pole combinations of modular winding designs are compared. The case of distributed winding with 1 slot per pole per phase was also considered for reference and comparison. The comparison found that

FSC winding machines lag behind the distributed winding design, mainly due to reduced winding factor. Their axial length was then let free to vary, in order to meet the power requirement of the base case and maintain common radius and mechanical speed.

The best topologies were finally selected, based on energy yield, efficiency and cost. Regarding FSC winding machines, the $q=0.4$ topology was confirmed to be the most attractive choice, as it combines very high winding factor with fault-tolerant characteristics. Secondly, topologies with $q = 0.3355$ and $q = 0.3618$ were also found to be attractive. Finally, regarding the choice of pole number, it was found that reducing the pole number was a good practice within the range examined. This was because of the resulting lower electrical frequency and corresponding iron losses.

6.1.4 Second-level comparison

The second-level comparison of segmented models, revealed that E-core segments are a more attractive approach for segmentation compared to segmental teeth, in terms of performance and efficiency. The distributed winding machine is still performing better than the FSC winding candidates, but only 3-phase segments can be employed.

The effect of varying the flux gap size was considered with respect to the average output power and the power ripple. A performance degradation was reported for higher gap widths in the $p = 150, Q = 360$ model, as has been indeed reported for machines with fewer poles than slots.

6.2 Future Work

Regarding future work in the analytical model, one direction in which the present model could be further extended regards its permeance function. A more detailed conformal transformation could be applied, in order to consider the exact tooth shape, instead of assuming straight teeth. Another potential improvement regards the model's losses: The computed armature fields have not been used in the rest of the analytical model. One way in which armature field distribution could be useful is the analytical determination of eddy current loss in the PM's. This type of loss is especially interesting for FSC windings due to their high harmonic content. Including segmentation in the analytical model could also be addressed. Finally, a comparison with a more precise analytical model (e.g. subdomain model) could also be insightful.

Regarding the topology investigation, comparing only optimized machines could be considered. Double-Layer windings could also be included, especially for machines without a critical requirement for fault-tolerance. The costs corresponding to ease of manufacture or replaceability of the modules have only been addressed qualitatively in the present work. A more detailed, quantified investigation of the costs related to modularity could be conducted. Regarding segment topologies, more segment designs (e.g 3-phase segments with FSC winding, T-core segments) could be tried.

Regarding the design choices, an interesting approach would be to formulate the comparison in a way that maintains the same axial length for all machines. This could give better indication of the actual differences particularly in the machines' costs. Also, parameters that were kept constant for simplicity (e.g. fill factor), could be addressed more realistically. Finally, optimizing the flux gap size to minimize the torque ripple or the mutual inductance between phases (magnetic isolation) could be considered in the design.

References

- [1] EWEA. Wind in Power: 2014, European Statistics. Technical Report February, European Wind Energy Association, 2015.
- [2] GWEC. Global Wind Energy Statistics. Technical report, Global Wind Energy Council, 2012.
- [3] EWEA. The European offshore wind industry key trends and statistics 1st half 2015. Technical report, European Wind Energy Association, 2015.
- [4] EWEA. Wind in power: 2011 European Statistics. Technical report, European Wind Energy Association.
- [5] Henk Polinder, H. Lendenmann, R. Chin, and W.M. Arshad. Fault tolerant generator systems for wind turbines. *2009 IEEE International Electric Machines and Drives Conference*, pages 675–681, 2009.
- [6] Udai Shipurkar, Henk Polinder, and Jan A Ferreira. Modularity in Wind Turbine Generator Systems - Opportunities and Challenges. 2016.
- [7] Jens Birk and Björn Andresen. Parallel-connected converters for optimizing efficiency, reliability and grid harmonics in a wind turbine. *2007 European Conference on Power Electronics and Applications, EPE*, pages 4–10, 2007.
- [8] Björn Andresen and Jens Birk. A high power density converter system for the Gamesa G10x 4 , 5 MW Wind turbine Keywords. *Power Electronics and Applications, 2007 European Conference on*, pages 1–8, 2007.
- [9] Thomas M. Jahns. Improved Reliability in Solid-State AC Drives by Means of Multiple Independent Phase Drive Units. *IEEE Transactions on Industry Applications*, IA-16(3[1] T. M. Jahns, Improved Reliability in Solid-State AC Drives by Means of Multiple Independent Phase Drive Units, IEEE Trans. Ind. Appl., vol. IA-16, no. 3, pp. 321331, 1980.):321–331, 1980.
- [10] J. J. Wolmarans, M. B. Gerber, H. Polinder, S. W H De Haan, J. A. Ferreira, and D. Clarenbach. A 50kW integrated fault tolerant permanent magnet machine and motor drive. *PESC Record - IEEE Annual Power Electronics Specialists Conference*, pages 345–351, 2008.
- [11] Time Magazine. <http://time.com/10115/google-project-ara-modular-smartphone/#ixzz2uxhipHxf>.
- [12] Wenping Cao, Barrie C. Mecrow, Glynn J. Atkinson, John W. Bennett, and David J. Atkinson. Overview of electric motor technologies used for more electric aircraft (MEA). *IEEE Transactions on Industrial Electronics*, 59(9):3523–3531, 2012.
- [13] A C Williamson. Odular Design of Wind. *Power*, 143(5), 1996.

- [14] Rod Jones and Philip Waite. Optimised power converter for multi-MW direct drive permanent magnet wind turbines. *Proceedings of the 2011 14th European Conference on Power Electronics and Applications*, pages 1–10, 2011.
- [15] Frede Blaabjerg. Future on Power Electronics for Wind Turbine Systems. *IEEE Journal of Emerging and Selected Topics in Power Electronics*, 1(3):139–152, 2013.
- [16] S. Kouro, M. Malinowski, K. Gopakumar, J. Pou, L.G. Franquelo, Bin Wu Bin Wu, J. Rodriguez, M.a. Perez, and J.I. Leon. Recent Advances and Industrial Applications of Multilevel Converters. *IEEE Transactions on Industrial Electronics*, 57(8):2553–2580, 2010.
- [17] Sverre Skalleberg Gjerde and Tore M. Undeland. Fault tolerance of a 10 MW, 100 kV transformerless offshore wind turbine concept with a modular converter system. *15th International Power Electronics and Motion Control Conference and Exposition, EPE-PEMC 2012 ECCE Europe*, pages 1–8, 2012.
- [18] Florence Libert. *Design, optimization and comparison of permanent magnet motors for a low-speed direct-driven mixer*. 2004.
- [19] N R Brown, T M Jahns, and R D Lorenz. Power Converter Design for an Integrated Modular Motor Drive. *Industry Applications Conference, 2007. 42nd IAS Annual Meeting. Conference Record of the 2007 IEEE*, pages 1322–1328, 2007.
- [20] Guang Jin Li, Zi Qiang Zhu, Martin P. Foster, Dave A. Stone, and Han Lin Zhan. Modular Permanent-Magnet Machines with Alternate Teeth Having Tooth Tips. *IEEE Transactions on Industrial Electronics*, 62(10):6120–6130, 2015.
- [21] Z. Q. Zhu, Z. Azar, and G. Ombach. Influence of additional air gaps between stator segments on cogging torque of permanent-magnet machines having modular stators. *IEEE Transactions on Magnetics*, 48(6):2049–2055, 2012.
- [22] Alan G. Jack, Barrie C. Mecrow, Phillip G. Dickinson, Dawn Stephenson, James S. Burdess, Neville Fawcett, and J. T. Evans. Permanent-magnet machines with powdered iron cores and prepressed windings. *IEEE Transactions on Industry Applications*, 36(4):1077–1084, 2000.
- [23] Greg Heins, Dan M. Ionel, and Mark Thiele. Winding Factors and Magnetic Fields in Permanent-Magnet Brushless Machines With Concentrated Windings and Modular Stator Cores. *IEEE Transactions on Industry Applications*, 51(4):2924–2932, 2015.

- [24] Jianxin Shen, Canfei Wang, Dongmin Miao, Mengjia Jin, Dan Shi, and Yunchong Wang. Analysis and optimization of a modular stator core with segmental teeth and solid back iron for pm electric machines. *2011 IEEE International Electric Machines and Drives Conference, IEMDC 2011*, pages 1270–1275, 2011.
- [25] Chester Ditmanson, Peter Hein, Stefan Kolb, Joaquin Molck, and Steffen Bernet. A new modular flux-switching permanent magnet drive for large wind turbines. *2013 IEEE Energy Conversion Congress and Exposition, ECCE 2013*, pages 896–903, 2013.
- [26] Max a Parker, Chong Ng, Li Ran, and Senior Member. Fault-Tolerant Control for a Modular Generator Converter Scheme for Direct-Drive Wind Turbines. *Industrial Electronics IEEE Transactions on*, 58(1):305–315, 2011.
- [27] Dan Kinzer and Ritu Sodhi. Converter Applications. pages 307–310, 1999.
- [28] Anoop Jassal. *Eddy Current Loss Modeling for Design of PM Generators for Wind Turbines*. 2014.
- [29] L. J. Wu, Z. Q. Zhu, D. Staton, M. Popescu, and D. Hawkins. Subdomain model for predicting armature reaction field of surface-mounted permanent-magnet machines accounting for tooth-tips. *IEEE Transactions on Magnetics*, 47(4):812–822, 2011.
- [30] L. J. Wu, Z. Q. Zhu, D. Staton, M. Popescu, and D. Hawkins. Comparison of analytical models for predicting electromagnetic performance in surface-mounted permanent magnet machines. *2010 IEEE Vehicle Power and Propulsion Conference*, pages 1–6, 2010.
- [31] Li Zhu, S. Z. Jiang, Z. Q. Zhu, and C. C. Chan. Comparison of alternate analytical models for predicting cogging torque in surface-mounted permanent magnet machines. *2008 IEEE Vehicle Power and Propulsion Conference, VPPC 2008*, pages 1–6, 2008.
- [32] Z.Q. Zhu and D. Howe. Instantaneous magnetic field distribution in permanent magnet brushless DC motors. IV. Magnetic field on load. *IEEE Transactions on Magnetics*, 29(1):152–158, 1993.
- [33] Z Q Zhu and David Howe. Instantaneous Magnetic Field Distribution in Brushless Permanent Magnet dc Motors Part 1 j. 29(1), 1993.
- [34] Z Q Zhu and D Howe. Instantaneous Magnetic Field Distribution in Brushless Permanent Magnet dc Motors , : Effect of Stator Slotting. 29(1), 1993.
- [35] Q Zhu. Instantaneous Magnetic Field Distribution in Permanent Magnet Brushless dc Motors, Part IV: Magnetic Field on Load. 29(1):152–158, 1993.

- [36] Henk Polinder, Frank F A Van Der Pijl, Gert Jan De Vilder, and Peter J. Tavner. Comparison of direct-drive and geared generator concepts for wind turbines. *IEEE Transactions on Energy Conversion*, 21(3):725–733, 2006.
- [37] Xinghua Wang, Qingfu Li, Shuhong Wang, and Qunfeng Li. Analytical calculation of air-gap magnetic field distribution and instantaneous characteristics of brushless DC motors. *IEEE Transactions on Energy Conversion*, 18(3):424–432, 2003.
- [38] L J Wu, Z Q Zhu, D Staton, M Popescu, and D Hawkins. Comparison of Analytical Models for Predicting Electromagnetic Performance in Surface-Mounted Permanent Magnet Machines.
- [39] Maarten J. Kamper, Arnold J. Rix, Dominic A. Wills, and Rong Jie Wang. Formulation, finite-element modeling and winding factors of non-overlap winding permanent magnet machines. *Proceedings of the 2008 International Conference on Electrical Machines, ICEM'08*, (3):1–5, 2008.
- [40] Ayman M. El-Refaie, Thomas M. Jahns, and Donald W. Novotny. Analysis of surface permanent magnet machines with fractional-slot concentrated windings. *IEEE Transactions on Energy Conversion*, 21(1):34–43, 2006.
- [41] Gordon R. Slemon and Xian Liu. Core losses in permanent magnet motors. *IEEE Transactions on Magnetics*, 26(5):1653–1655, 1990.
- [42] Chunting Mi, Gordon R. Slemon, and Richard Bonert. Modeling of iron losses of permanent-magnet synchronous motors. *IEEE Transactions on Industry Applications*, 39(3):734–742, 2003.
- [43] F. Magnussen and C. Sadarangani. Winding factors and Joule losses of permanent magnet machines with concentrated windings. *IEMDC 2003 - IEEE International Electric Machines and Drives Conference*, 1:333–339, 2003.
- [44] H. Polinder, D. Bang, R.P.J.O.M. van Rooij, A.S. McDonald, and M.A. Mueller. 10 MW Wind Turbine Direct-Drive Generator Design with Pitch or Active Speed Stall Control. *2007 IEEE International Electric Machines & Drives Conference*, 2:1390–1395, 2007.
- [45] Ayman M. EL-Refaie. Fractional-slot concentrated-windings synchronous permanent magnet machines: Opportunities and challenges. *IEEE Transactions on Industrial Electronics*, 57(1):107–121, 2010.
- [46] Xiaolong Zhang and Ronghai Qu. Pole number selection strategy of low-speed multiple-pole permanent magnet synchronous machines. *Proceedings of the 2013 IEEE International Electric Machines and Drives Conference, IEMDC 2013*, pages 1267–1274, 2013.

- [47] Z. Q. Zhu, Z. P. Xia, L. J. Wu, and G. W. Jewell. Influence of slot and pole number combination on radial force and vibration modes in fractional slot PM brushless machines having single- and double-layer windings. *2009 IEEE Energy Conversion Congress and Exposition, ECCE 2009*, pages 3443–3450, 2009.
- [48] J. Cros and P. Viarouge. Synthesis of high performance PM motors with concentrated windings. *IEEE International Electric Machines and Drives Conference, IEMDC 1999 - Proceedings*, 17(2):725–727, 1999.