Low Cost Electronically Controlled Motors for Vehicle Actuation Systems

Thesis submitted for the degree of
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By

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This thesis is dedicated to the memory of my Grandad, Percy Griffiths, who died at the age of 88 years on 19th October 2007, without ever growing old.

I only wish you could have read it.

Thankyou for all the lovely things that you did for me.

You were a truly wonderful man and I will never forget you.

Until we meet again, sleep peacefully.
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Abstract

An investigation into improving a design of the flux switching motor for use in an automotive application is presented. This concentrates on developing computer simulation tools to model the motor performance effectively and hence enable the development of an alternative design.

The flux switching motor is introduced as a relatively new type of brushless drive with a high level of robustness and low cost power electronics. These properties in particular make the flux switching motor well suited to the automotive industry, where reliability and cost are paramount.

This thesis details research carried out to develop a new circuit simulation of the motor and its power electronics using the program PSpice. The model developed here is then used as a tool to investigate the performance of the motor. This is used in conjunction with static finite element analysis performed in the program Opera and later on with time stepping finite element analysis also in Opera. These simulation tools are tailored to model the flux switching motor accurately and aid with the improvement, in terms of torque production, of the design.

The models are used to produce a new motor design making use of 12 rotor teeth instead of the 4 teeth originally presented. This design shows promise in early static simulations with the potential for higher power due to a higher operating frequency, although high iron loss indicates this may be outweighed. However, time stepping simulation shows that torque production is actually lower than the original 4 tooth design due to lower flux in the smaller rotor teeth.
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This thesis details the research carried out to develop a new type of brushless motor drive for use in automotive applications.

The area of research is becoming more important as automotive manufacturers seek cheap yet reliable drives for the increasing number of electronic vehicle systems. The general robustness and reliability of brushless motors lend themselves well to the automotive sector, but the relatively expensive power electronics needed to control them can rule out their use. The flux switching motor, introduced here as a new type of brushless drive, makes use of inexpensive power circuits and hence could be potentially ideal for this market.

The research conducted here endeavours to develop the flux switching motor and create an optimal design using finite element and circuit simulation tools. These development tools will also be tailored to model the motor effectively, making use of both static and time stepping finite element analysis (FEA) as well as PSpice circuit simulation.

1.1 Objectives

The objectives of this research are listed below:

- To develop a reasonably accurate model of the flux switching motor within a PSpice circuit simulation, which is relatively quick and easy to use as a development tool.
- To use static finite element analysis to manipulate the geometry of the lamination profile of an existing motor design to improve the torque output.
- To use Rotating Machines software to perform time stepping finite element analysis on existing and potential designs.
- To combine the use of the simulation development tools detailed above to produce a more optimised design of flux switching motor with higher torque output.
1.2 Format of Thesis

This thesis is composed of 7 chapters including the introductory chapter.

Chapter 2 details the background information highlighting the purpose of this work and looks at the growing importance of electronic drives in the automotive sector. It also introduces the flux switching motor, describing its operation and indicating why it may be well suited for this application. An existing motor design is presented with accompanying simulation and experimental data.

Chapter 3 documents the process of developing a PSpice circuit simulation of the motor and its power electronics. Here, the PSpice model is used to replicate both simulation and experimental data already available for the existing design. This model is shown to be a reasonably accurate representation of the motor and therefore a useful development tool.

Chapter 4 presents the work carried out using static finite element analysis to manipulate the geometry of the motor laminations. This investigation looks to improve the torque output of the machine by changing rotor and stator parameters and hence produce a new improved design. At the end of this chapter a design is chosen to be taken forward for further investigation.

Chapter 5 seeks a more radical approach to improving the motor design. Here, the number of rotor teeth are increased from 4 to 12 and the laminations are changed significantly. Still making use of static finite element analysis, a number of 12 tooth designs are compared with the original 4 tooth design and the new design chosen in the previous chapter. These designs are investigated more closely taking into account copper and iron losses. The chapter concludes by choosing a 12 tooth design with the most potential for improvement on the original. This is then investigated further in the next chapter.

Chapter 6 makes use of the Opera Rotating Machines (RM) program to perform time stepping finite element analysis. This chapter brings together results from the PSpice circuit simulation and compares them to the RM results. The PSpice model is found to
be a good match within reason to the more complex RM model. Hence, it can be used as a preliminary tool to predict the performance of new motor designs with various winding configurations at different speeds. Both models are then used to simulate the new 12 tooth design chosen in chapter 5. The RM results for the original 4 tooth and new 12 tooth motors are then compared.

Chapter 7 brings together the work and highlights conclusions found in each chapter. An overall conclusion is then formed and some suggestions for further possible work are made.
Chapter 2
Background

2.1 Introduction

This chapter gives some background information regarding the area of research and indicates why the work detailed in this thesis was necessary. It details specific applications where the technology in question may be beneficial and then goes on to introduce the Flux Switching Motor, how it operates and how it is investigated throughout this thesis.

The importance of electrical motors in the automotive industry is growing. Traditionally these motors have mainly been brushed DC (series wound DC motors) or permanent magnet brushed DC motors, which are ideal for the automotive sector due to their ease of use from battery/DC supply and low cost. Brushed motors are a proven and extensively developed technology in many everyday appliances. The brushed AC motor (series universal), which can be run off DC or AC supplies making it very flexible, is commonly found in the home in many applications from vacuum cleaners to power showers. However, the development in recent years of brushless motor technology is starting to challenge the established brushed machine. Brushless machines have usually been regarded as expensive due to the power electronic circuits used to drive and control them. This would obviously make them unsuitable for the automotive market where cost is an overriding factor. But, as the technology has progressed, the cost has decreased considerably making them a more viable alternative. In some cases, as discussed later on in this chapter, the brushed motor is no longer sufficient for many of the tasks required of motors in modern automobiles and it is here where brushless motors can take advantage. Manufacturers of brushless motors are pushing to develop smaller motors yet maintain power and speed. The driving force behind market growth is motorists' insatiable demand for enhanced safety, comfort, fuel economy and improved emission performance. [1]
2.1.1 Anti-lock Braking Systems

There are many automotive applications involving the control of the vehicle and driver aids that would be well suited to the use of brushless motors. Some of these will be discussed later in this chapter. A good starting point though, as it is the basis for such systems as traction control, is anti-lock braking.

An anti-lock braking system (ABS) is a system on motor vehicles which prevents the wheels from locking while braking. The purpose of this is twofold: to allow the driver to maintain steering control under heavy braking and, in most situations, to shorten braking distances (by allowing the driver to hit the brake fully without the fear of skidding or loss of control). Disadvantages of the system include increased braking distances under certain conditions and the creation of a "false sense of security" among drivers who do not understand the operation and limitations of ABS. [2]

There are four main components to an ABS system:

- Speed sensors
- Pump
- Valves
- Controller [3]

The speed sensors constantly monitor the rotation speed of each wheel. When they sense that any number of wheels are rotating considerably slower than the others (a condition that will bring it to lock) the valves are moved using solenoids to decrease the pressure on the braking circuit, effectively reducing the braking force on that wheel. Wheel(s) then turn faster and when they turn too fast, the force is reapplied. This process is repeated continuously, and this causes the characteristic pulsing feel through the brake pedal. [2]
Figure 2.1 below shows the location of the anti-lock brake components and figure 2.2 shows the anti-lock brake pump and valves.

Figure 2.1 Location of anti-lock brake components [3]

Figure 2.2 Anti-lock brake pump and valves [3]

This thesis is focused on the pumping aspect of the system and developing a new type of brushless motor for this and similar automotive applications such as traction control. The diagram in figure 2.3 below shows the exact role that the motor plays within the system and indicates the kind of demands it is placed under:
2.1.2 How ABS Works

When the brake pedal is pushed, the brake calliper is applied to the disc using pressure in the brake fluid supplied from the high pressure master cylinder. This is created by the engine. In anti-lock systems, the calliper is required to be rapidly applied and then removed from the disc automatically while the driver continues to apply constant pressure to the pedal. This is achieved by draining off the brake fluid into an accumulator (to release the calliper) and then replacing this fluid from the master cylinder (to reapply the calliper) as required. If this was the complete system, then as fluid was drained from the master cylinder (ending up in the low pressure accumulator) then the pedal would continue to drop as it was pushed (ΔP) and eventually reach the floor. Hence, in order to avoid this, the brake fluid must be returned from the low pressure accumulator to the high pressure master cylinder. This is done using a brushed DC motor operating with a torque of around 0.8Nm sufficient
enough to pump the brake fluid from an area of low pressure to an area of much higher pressure that can be at as high as 180bar.

The brushed motor is perfectly well suited to this application as it is capable of producing the torque required and is reliable enough to meet the demands. It is unlikely in everyday use that the anti-lock system would be used frequently as it is only really required in emergency situations when very heavy braking is essential. Thus it may only be used a few times a year and so the issue of the brushes wearing out becomes insignificant.

Indeed, the brushed motor is well suited to many applications in the modern automobile. As with ABS, electric windows will only be operated occasionally and so again there is no issue with brush wear. The same goes for electrically adjustable seats and wing mirrors, where the capabilities are sufficient and the motor can be made small enough and most significantly, cheaply enough. In automotive design, cost is the overriding issue and it is essential that components can be manufactured and fitted as cheaply as possible. Therefore, in many automotive applications, the brushed motor is a satisfactory solution.

However, today's automobiles are becoming more hi-tech, with increasing use of motor driven devices as well as gadgetry that has become standard in terms of driver aids. These features combine to make a more comfortable, convenient and often safer environment, be it climate control, satellite navigation or traction control. These devices all place extra demand on a cars' electrical system and many require the use of motors. Even fundamental components have been replaced or will be in the future, such as steering, braking and accelerator systems, by electrically operated, motor driven systems. There is clearly a need here for robustness and reliability based around tried and tested technology. At the same time though, the brushed machine falls short in being capable of these tasks. In many applications such as electro-hydraulic steering, the motor is used frequently and is required to produce high torques and speeds over long time frames. The use of a brushed motor no longer becomes an option. Couple this with the fact that such systems have been investigated where a motor can be used to directly drive the brake calliper onto the disc (no hydraulics at all) and it can be clearly seen why the automotive industry is starting to
look towards brushless technology. A brushless motor here would be able to operate at high speeds with a high level of robustness and reliability of its controlling circuits. This could eventually allow the complete drive by wire vehicle.

Traditionally, brushless motors can be expensive due to the added cost of power electronic circuits needed to control them. The flux switching motor becomes a promising solution here as a brushless drive that requires very simple and cheap power electronics.

2.2 Other Motor Types Suitable for Automotive Applications

This section details the advantages and disadvantages of certain motor types that the flux switching motor must compete against if it is to be successful in the automotive sector.

2.2.1 The DC Motor

In a DC motor, the armature winding is placed on the rotor and the field winding on the stator. The stator has salient poles excited by one or more field windings that can be shunt or parallel wound. A direct current is passed through the field winding to produce an air gap flux distribution that is symmetrical about the pole axis. The voltage induced in the armature winding is alternating. A commutator - brush combination is used as a mechanical rectifier to make the armature terminal voltage unidirectional and the mmf wave due to the armature current fixed in space. The brushes are placed so that as an armature coil passes through the region in the middle of two field poles, the current though it changes direction. This makes all the conductors under one pole carry current in one direction and hence the mmf due to the armature current is along the axis midway between two poles. The armature mmf is in quadrature with the field mmf – maximising torque production.

The advantages of brushed DC motors are:

- Speed can be controlled over a wide range with relative ease by controlling the armature voltage or resistance, or by field control (field weakening).
A wide variety of volt-ampere or torque-speed characteristics can be obtained from various connections of the field winding.

The DC motor is simple and versatile and can be configured in several ways to suit various applications.

The disadvantages of brushed DC motors are:

The use of brushes gives rise to several downsides to the DC motor;

Brushes eventually wear out and require replacing, which can be difficult in some applications and may result in the replacement of the whole motor.

Brushes produce considerable noise.

As brushes wear down they produce dust and grease, leading to the motor becoming dirty and possibly affecting performance. Hence frequent maintenance may be required.

As brushes move over the commutator segments sparks are produced. This can cause significant electromagnetic interference, which can affect other nearby systems.

**2.2.2 Permanent Magnet DC Motors**

In PMDC motors, permanent magnets are used to replace the stator wound poles of a conventional brushed DC motor, resulting in a relatively smooth stator structure as opposed to salient poles. The rotor consists of a DC armature with commutator segments and brushes.

The advantages of permanent magnet motors are:

Increased efficiency due to no field windings and hence no copper loss.

Can be made smaller than wound pole motors due to no space being required for field windings. This reduced size reduces material cost and compensates somewhat for the extra cost of permanent magnets.

The disadvantages of permanent magnet motors are:

Excessive armature current can result in demagnetization, which can also be caused by excessive heating if overloaded for a prolonged period.
The magnetic field is permanent and thus may require shielding in some applications to avoid any damage from foreign matter.

Permanent magnets cannot usually produce as high a flux density as wound pole motors.

Operation flexibility is limited as speed cannot be controlled via the field flux (field weakening) and must therefore be controlled by armature voltage.

2.2.3 Brushless DC Motors

This motor is a type of synchronous machine with radially magnetized permanent magnets mounted on a steel core rotor while the armature is placed on the stator. A position sensor and an inverter are used for commutation and hence no brushes are required. The operating characteristics are similar to a brushed dc machine.

The advantages of brushless DC motors are:
- They are compact due to no field windings, brushes and mechanical commutators, which results in a smaller rotor size and a high power density.
- The absence of brushes also means low maintenance with high speed and torque capability and no sparking.
- A high torque/inertia ratio and a fast dynamic response can be achieved.
- This machine is also highly efficient with better heat dissipation than a brushed motor due to a stationary armature winding.

The disadvantages of brushless DC motors are:
- High system cost due to relatively complex power electronics (inverters) and controllers as well as rotor position sensors.
- Permanent magnets also tend to be relatively expensive.

2.2.4 Switched Reluctance Motors

Switched reluctance motors have saliency in both the rotor and the stator, with the excitation windings wound on the stator and no winding on the rotor. Torque is
produced by the tendency of the rotor pole to align with the stator pole to maximise
the flux linkage when the stator pole winding is excited by a current.
These motors generally have one pair of poles less on the rotor than on the stator with
two or more phases. A coil is wound around each stator pole and is connected, usually
in series, with the coil on the diametrically opposite stator pole to form a phase
winding. The individual phases are excited by current pulses causing the rotor to align
accordingly with each set of stator poles. By switching the phases at the correct time
with regard to rotor position, the rotor is made to rotate. A rotor position sensor is
therefore necessary.

The advantages of switched reluctance motors are:
No rotor winding means that these motors are rugged and are well suited to high speed
applications.
They are reliable due to a very simple power control circuit and require only unipolar
current.
They can achieve a wide constant power speed range and have good power density.

The disadvantages of switched reluctance motors are:
They suffer from excessive torque ripple, which can be a problem in certain
applications where a steady torque output would be desireable. This also contributes
to acoustic noise.
Although the construction of the machine is simple, its control can often be
complicated and hence the power electronics are expensive.
A rotor position sensor is required which adds to the cost.
2.3 The Flux Switching Motor

The flux switching motor is being considered as an alternative to the brushed and permanent magnet machines currently used in many automotive applications mentioned earlier in this chapter. The aim of the research is to improve the model of the flux switching motor [4, 5] for use in any system, from all pumping applications to general automotive use such as in anti-lock brakes or power steering systems.

In previous research C. Pollock and H. Pollock have designed a flux switching motor for automotive applications. [6, 7]

The flux switching motor has been investigated previously for use in many other applications including power tools, vacuum cleaners and hand dryers [5]. The motor is a combination of the switched reluctance motor and inductor alternator [8-10], with a doubly salient structure. There are no brushes or permanent magnets and both field and armature windings are on the stator. These windings are fully pitched, meaning they span the same number of stator poles as there are phases. Excitation current is applied to both field and armature windings constantly throughout operation. The armature winding is supplied with bipolar current, while the field is supplied with a constant unipolar direct current. The armature current must therefore be switched and this can be controlled by very simple power electronics, which are low cost and reliable. The field current produces a magnetic flux in the field, which coupled with the rotation of the rotor produces a varying flux in the armature creating a back emf (so called because it opposes the applied voltage on the motor).

This principle can be described as follows;

Field → Field flux → Rotation → Varying flux linking armature → Back emf

This back emf and the armature current give rise to the torque.
The flux switching motor has several advantages over other motor types (detailed earlier in this chapter) which can be exploited in a range of applications as well as automotive. These include:

- Relatively low cost
- Reduced noise
- Low maintenance and longer service life
- Speed control via simple, cheap and reliable power electronics
- Robust design, making it suitable for high speed applications.

The flux switching motor has been shown to produce torque, power and efficiency outputs that are better than or comparable to a permanent magnet brushed motor of the same size. [6].

2.3.1 Principle of Operation

Due to the doubly salient construction of the motor, torque is developed by the tendency of the rotor to align itself in a position of minimum reluctance with the stator, within a magnetic circuit energised by the phase windings. [4]

Figure 2.4 below illustrates how the motor operates using conventional dot and cross notation. In 2.4(a), the combined flux pattern produced by the currents in the field and armature can be seen to cause the rotor to align in a vertical position. However, when the direction of the current in the armature is reversed in 2.4(b), the resultant flux changes direction and the rotor now aligns horizontally. The stator flux vector does not rotate, but simply switches between horizontal and vertical directions. If the armature current is switched at the correct time with respect to the rotor position, then the rotor will rotate. This is done by using a position sensor on the rotor.
To enable optimum performance it is necessary to align the position sensor so that switching occurs at (as close as possible to) the right time. This is done by observing the back emf produced in the armature when direct current is applied to the field and the rotor is rotated (either by hand or using another machine, eg a drill). The position sensor can then be aligned so that it is leading this back emf by a certain number of degrees. This can be down to trial and error in determining how much the position sensor square wave leads the back emf in order to optimise performance on a particular design of motor. A more efficient way of doing this has now been developed using software to create the switching advance with regard to the speed of operation. This means the position sensor can be completely aligned with the back emf.

A method of sensor-less control has also been developed at the University of Leicester by C. Pollock[11]. This makes use of an algorithm to calculate the switching angle based on measurements of the mutual coupling between the field and armature windings. The turn on and turn off points are controlled to be as close as possible to the appropriate mutual inductance value for the speed and torque.
2.3.2 Power Electronics

In order to switch the current in the armature, a power converter is necessary. There are numerous topologies that can be used for this, including the common H-bridge configuration shown below in figure 2.5. In this circuit, S1 and S4 are switched in phase with each other, but out of phase with S2 and S3. So, when S1 and S4 are on, current flows through the armature winding from left to right. S1 and S4 are switched off and S2 and S3 are then switched on allowing current to flow the opposite way through the armature from right to left.

![Figure 2.5 H-bridge power converter](image)

However, this circuit uses four power switches and a simpler way of providing a bipolar current is possible using only two ground referenced power switches. This eliminates the cost of isolated or floating gate drives. Figures 2.6 and 2.7 show two such circuits with the field winding connected either in a shunt (2.6) or series (2.7) configuration. As with the full bridge circuit shown above, a diode is necessary in the series field circuit to provide a path for the current when the armature switches are off.

A shunt connected field can make use of a field control switch to allow de-energisation of the field winding when the motor is stopped and offers the significant advantages of field control (eg, field weakening) during operation.[5]. Hence, a shunt connection allows easy control over a large torque range.
A series connected field allows the field current to be automatically modulated with the magnitude of the armature current. It also provides a significant amount of EMC filtering to the motor drive, which is an important cost reduction.[12]. When connected in series, torque is approximately proportional to the square of the armature current, which means a high starting torque can be achieved.

Figure 2.6 Power converter for flux-switching motor employing only two power switches by using bifilar windings and shunt connected field.

Figure 2.7 Power converter for flux-switching motor employing only two power switches by using bifilar windings and series connected field.
In the above circuits, bifilar windings are used where the armature consists of 2 coils that are magnetically coupled and have the same number of turns. With the MOSFET S1 turned on, and S2 off, Armature 1 is conducting. As the switching point approaches, S1 is turned off and the stored energy associated with the armature inductance causes the armature current to transfer to the second closely coupled winding, Armature 2. A current also flows in the MOSFET diode back to the battery via the freewheel diode in parallel with the field. When the current in the armature reaches zero, S2 is switched on and carries the armature current in the forward direction. Armature 2 is connected in opposite polarity to Armature 1 (as shown by the dot notation in the above figures) and so excitation is reversed. At the point of switching, there will be some leakage energy, which is not coupled from one winding to the other. This small amount of energy is absorbed in a very brief avalanche breakdown of the appropriate power MOSFET’s immediately after turn off. [6]

The advantages of using only two power switches are that the gate drives for the two power devices are kept simple and the obvious lower cost of half as many devices. The disadvantages are that there will be double the voltage rating in the switches and the presence of leakage inductance. Also, only around 75% of the copper in the machine is being used at any one time.

MOSFET’s are used as opposed to IGBT’s, which cannot withstand the overvoltage. The MOSFET’s provide an integral diode to act as a path when the alternate switch turns off. They also absorb the small amount of leakage energy not transferred to the second closely coupled armature winding. [5, 12]
2.4 Modelling the Motor

In order to improve the performance characteristics of the motor for use in this application, it will be necessary to make use of computer simulation to model the motor accurately.

2.4.1 Finite Element Analysis (FEA)

Finite Element Analysis consists of the following steps:

- The process begins with the creation of a geometric model.
- The model is broken down into smaller elements of simple shapes (finite elements) connected at specific node points, creating a mesh.
- Equations of equilibrium, in conjunction with applicable physical considerations are applied to each element, and a system of simultaneous equations is constructed.
- The system of equations is solved for unknown values using the techniques of linear algebra or nonlinear numerical schemes, as appropriate.

2.4.2 Opera-2d Static Finite Element Analysis

Opera-2d is a suite of software programs for 2-dimensional electromagnetic field analysis created by Vector Fields. These programs make use of finite element analysis to solve partial differential equations describing how fields behave, for example, Poisson's equation, the Helmholtz equation and the Diffusion equation. When designing in the areas of magnetostatics, electrostatics and electromagnetics, the solution of these equations plays an important role.

Opera-2d uses a powerful pre-processor allowing the user to enter data using a graphical interaction to produce a geometric model. The model space is then divided
up into a contiguous set of triangular (as in this software) or rectangular elements to create a mesh. A solution can then be calculated using one of several suitable analysis modules available for various types of electromagnetic excitation conditions eg. static or steady state. The correct solution is determined by an iterative method including non-linear effects if these are modelled.

A postprocessor can then be used to examine the result, again using a graphical interaction to examine such system variables as potentials, currents, fields, forces and temperature. The user can define additional variables to tailor the results to specific applications. The mesh can also be refined for greater accuracy by analysing numerical errors produced by the mesh being too coarse. [13]

2.4.3 Opera Rotating Machines – Time Stepping FEA

The Rotating Machine Program is a Transient Eddy Current Solver, extended to include the effects of rigid body (rotating) motion. The solver also provides for the use of external circuits and coupling to mechanical equations. [14]

The program uses time stepping analysis to dynamically model the motor, taking into account effects such as eddy currents and therefore produces a more accurate representation. This is not possible using static analysis, where the results are only affected by the present rotor position. The time stepping analysis makes calculations for each time interval as the rotor rotates and hence, for each calculation takes into account the previous rotor position and speed.

In the rotating machine program, the windings are connected to external circuits and the values of resistance, inductance, capacitance and number of coil turns can be specified along with the supply voltage. The current in each coil is then calculated from the applied voltage and the circuit properties. This differs from the static model where the coils are excited by specifying the current density that will produce the required mmf.
2.4.4 PSpice

PSpice is a circuit simulation program used to verify circuit designs and to predict circuit behaviour. It makes use of component libraries to build schematics of circuits made up of various electrical components such as resistors, capacitors, inductors, transistors etc. Various types of analysis can then be performed on the circuits, for example, non-linear DC analysis, non-linear transient analysis, linear AC analysis, temperature and noise analysis. The results can be viewed by placing probe markers on various parts of the circuit to display the waveforms produced at those points. Data files can also be produced to display or export data from the simulation.

2.4.5 Limitations in the Modelling Tools

The modelling tools above all have strengths in particular areas useful for design analysis in this field. However, they all have notable limitations meaning that no one tool is capable of modelling a motor design sufficiently on its own. By using the different methods of modelling together in a process, their advantages can be exploited and their limitations overcome. For example:

- PSpice is an excellent tool for simulating circuits, but is not good for simulating magnetic properties. In order to keep the model simple, the components used, such as inductors, are linear and so this model is no use for simulating non-linear characteristics present in the flux switching motor. In order to use non-linear inductors, a vast amount of static data would be needed, defeating the object of the simple model.

- Static FEA is very useful for performing analysis on the behaviour of fields and analysing magnetic properties of a motor design. It is no use for simulating the electric circuits and does not take into account any dynamic effects.

- Time stepping FEA is useful for simulating the dynamic effects present during the operation of the motor. It is also able to simulate electric circuits, but this
aspect is limited and fairly basic. It is also very time consuming and somewhat complicated to use, requiring a lot of computing time.

2.4.6 The Development Process

The flowchart below illustrates the process undertaken using FEA and circuit simulation used to investigate the flux switching motor and improve upon its design in terms of its lamination geometry and winding configuration.
Repeat parts of the design process to bring about further improvements

No

Are static FEA and PSpice results a satisfactory improvement?

Yes

Use winding configuration, obtained in PSpice, in the RM to perform time-stepping FEA

Compare current and torque waveforms produced using time stepping with PSpice results and published data

Use static FEA to investigate geometry of rotor and stator laminations in order to improve torque output (see chapter 4)

Use static FEA to produce inductance and mutual inductance values and waveforms to create a PSpice circuit model

Circuit simulation in PSpice to produce current and torque waveforms for a particular winding configuration

Vary winding configuration in PSpice by changing number of turns, then recalculating resistor and inductor values accordingly until output torque is optimised

Use circuit simulation in PSpice to produce current and torque waveforms for a particular winding configuration

Use static FEA to produce inductance and mutual inductance values and waveforms to create a PSpice circuit model

Vary winding configuration in PSpice by changing number of turns, then recalculating resistor and inductor values accordingly until output torque is optimised

Are static FEA and PSpice results a satisfactory improvement?

Yes

Use winding configuration, obtained in PSpice, in the RM to perform time-stepping FEA

Compare current and torque waveforms produced using time stepping with PSpice results and published data

Repeat parts of the design process to bring about further improvements
2.5 Published Data on Existing Motor Design

A motor design has been developed by C. Pollock in cooperation with TRW for an automotive application [6, 7]. This design has been simulated using FEA and a prototype has been built. Some experimental data has thus been produced and compared to the simulated results. This section details this work with simulated FEA and experimental waveforms produced by C. Pollock.

The design of this FSM is of an 8/4 construction (i.e., 8 stator teeth and 4 rotor teeth). A diagram of the prototype motor's lamination profile can be seen below in figure 2.8. The field winding slots are labelled 4 and 5, and the armature winding slots are labelled 7 and 8.

![Prototype motor lamination design](image)

**Figure 2.8 Prototype motor lamination design [7]**

The following figure 2.9 shows the back emf's produced by the prototype motor in the armature winding when current is fed into the field winding and the rotor is rotated mechanically. The experimental back emf is shown with the position sensor square wave for three different values of field current. A comparison to simulated results is also displayed for each field current, which was simulated at 3030rpm. The shape of the back emf waveforms induced in the armature windings is dependant on the geometric lamination profile of the motor.
a.) 20 A field current

b.) 40 A field current

c.) 60 A field current

Figure 2.9 Induced armature voltage (back emf) – comparison of measurement (left) and simulation (right) [7]
Dynamic (time stepping) finite element methods have then been used by David Moule at TRW to simulate the motor under running conditions to produce waveforms of voltage, field and armature current, and torque. These can be seen below in the following figures for two different speeds along with some experimental results obtained while running the prototype for comparison.

These results indicate the advance and cut-off times used, which are necessary for the power switches to operate correctly. The advance means the time between the corresponding MOSFET’s turning on and the back emf waveform passing through zero. The cut-off is the time between one corresponding (diagonally opposite) set of MOSFET’s turning off and the other set turning on. During this time all the devices are off and this ensures that no short circuit or “shoot-through” can occur by current flowing straight through two MOSFET’s from the positive side to ground.

Figure 2.10 Simulated winding and supply voltages, 2175rpm warm conditions

(Note: winding voltages include voltage drop in leads and end inductances.)
The voltage waveforms above show the variation in supply voltage along with the bipolar armature winding voltage. The field winding voltage is mainly positive except immediately after switching of the power devices, when the voltage at the point where the field and armature windings meet is higher than the supply voltage.

Figure 2.11 Simulated diode and armature current waveforms, 2175 rpm warm conditions [7]
Figure 2.12 Simulated waveforms at 0.8 Nm torque, 2175 rpm, 600 µs advance (8 degrees), 168 µs cut-off (2 degrees), warm conditions [7]

Figure 2.13 Practical waveforms at 0.8 Nm torque, 2175 rpm, 600 µs advance (8 degrees), 168 µs cut-off (2 degrees) [7]
Figure 2.14 Simulated waveforms at 1.0 Nm torque, 1090 rpm, 800 μs advance (5 degrees), 168 μs cut-off (1 degree), warm conditions [7]

Figure 2.15 Practical waveforms at 1.0 Nm torque, 1090 rpm, 800 μs advance (5 degrees), 168 μs cut-off (1 degree) [7]
The field and armature current waveforms shown in the above figures show the relatively constant field current and the bipolar armature current. Figure 2.11 shows the pulses of current through the freewheel diodes within the H-bridge configuration used. It also shows the larger, longer pulses of current through the freewheel diode in anti-parallel with the field winding. Figures 2.12 and 2.14 show that the field and armature currents are equal in magnitude for a large part of the cycle and it can also be seen how the current returns to the supply when switching occurs.

The simulated time stepping FEA analysis waveforms compare very well to the experimental waveforms shown in figures 2.13 and 2.15, showing that the time stepping finite element model is an accurate representation. Both the simulation and experimental waveforms show how the symmetry of the armature current waveform is affected by the asymmetry of the rotor and stator.

Figure 2.16 Simulated torque and average torque waveforms at 1090 rpm [7]
Figures 2.16 and 2.17 above show simulated torque and average torque waveforms at the two different speeds investigated. It is clear how the instantaneous torque varies greatly, whereas the average torque tends towards a steady value with some ripple. This high torque ripple is due to the fact that the armature current has to reverse direction twice in each electrical cycle. A small component of torque is produced by the current in the field winding and so the resultant torque is not quite zero when the armature current passes through zero. [6]
2.6 Initial Development Work

The work carried out by David Moule at TRW was taken as a starting point for investigation, with the results being used for comparison throughout the development process of the motor undertaken here. The initial steps of this development are detailed below in the following sections where simulation and experimental work was carried out to provide a basis for the design process.

2.6.1 Modelling of Flux Waveforms Using Static Finite Element Analysis

Waveforms of field and armature flux were produced using Opera simulation. The prototype motor (shown in figure 2.9) has 4 parallel field coils with 21 turns in each coil, and these are in series with 4 parallel armature coils with 19 turns in each coil. The stack length is 30mm. A current of 20A was fed into the field winding (5A into each 21 turn coil) at 20 different rotor positions between 0 and 90 degrees. The resultant waveforms are for one field and one armature turn per unit stack length, and the flux pattern produced repeats every 90 degrees.

The field and armature flux waveforms can be seen in figure 2.18.

Figure 2.18 Field and armature flux per metre per turn plotted against rotor position.
2.6.2 Experimental Back EMF Tests

A back emf test was performed on the motor in order to determine the accuracy of the Opera model. The rotor was rotated at 3030rpm using an electric drill, while 20A current was fed into the field winding (so, 5A in each 21 turn coil). The back emf produced in the armature winding was then measured and can be seen in figure 2.19.

Figure 2.19 Armature back emf waveform produced at 3030rpm with 20A field current

This back emf waveform was then integrated using Matlab, and the resulting DC offset was removed, to produce the armature flux linkage waveform for one armature coil of 19 turns. This waveform could then be compared to the simulation, which was scaled accordingly by multiplying by the number of armature turns (19) and the stack length (0.03m). The experimental and simulated waveforms can be seen in figure 2.20.
The simulated waveform for the armature flux linkage is a reasonably good representation when compared to the experimental waveform. However, there is some difference between the waveforms with the experimental results producing a higher flux linkage. This error was found to be around 5.7% at the peak of the waveforms. The error may be down to a number of factors including dynamic and 3D effects not taken into account by the static analysis, copper and iron losses, not taking into account the end windings and inaccuracy in the mesh and material data. In order to improve the simulation and achieve a greater accuracy, it would be necessary to take into account some of the above parameters that are not currently modelled.

2.6.3 Development of a PSpice Circuit Simulation Model

A PSpice circuit simulation model was then developed building on the data presented above. This model sought to accurately simulate the motor using basic PSpice components and account for the effects of back emf and mutual inductance using feedback circuits incorporated into the windings and their associated switches. In this
way, predictions could be made on how the motor would perform at various speeds
with regard to the back emf's produced and the currents present in the windings. The
resulting model would provide a fairly simple way of simulating the motor
performance and could be used to investigate various winding and speed combinations
quickly and with relative ease. This process is detailed in chapter 3.
Chapter 3
Developing a Circuit Simulation of the Flux Switching Motor in PSpice

3.1 Introduction

This chapter details the steps taken to model the flux switching motor and its power electronics using PSpice circuit simulation. Circuit simulation tools have been used previously to model switched reluctance motors [15, 16] and brushless ac motors [17], but until now the flux switching motor has not been modelled in this way.

3.2 The Motor Circuit

The circuit used to control the motor has been described in the previous chapter and it was explained how there is more than one way to switch the current in the armature using power devices. The cheapest of these methods is to use bifilar windings and hence only two power switches are required. However, in order to keep the model simple, the circuit considered here is the classic H-bridge making use of 4 MOSFET's to switch the armature current. This will produce exactly the same results as using the bifilar windings, but will be easier to model in the circuit simulation program, PSpice. The circuit can be seen below in figure 3.1.

![Flux Switching Motor circuit diagram](image)

Figure 3.1 Flux Switching Motor circuit diagram
The work carried out previously by Prof. Pollock to model the prototype motor in conjunction with TRW was presented in chapter 2. [7] The speeds and switching patterns used to obtain the simulated and experimental results will be used in the development of the PSpice model to allow direct comparison.

3.3 Initial PSpice Circuit Model

Initially, a simple model was developed in PSpice, which basically replicated the circuit shown in figure 3.1. This model did not take into account any of the mutual effects produced by the windings when the motor is running. These effects include the mutual inductances between the windings and the back emf's induced in each winding by the current in the other. Hence, this model would not be adequate enough to use as a predictive modelling tool, but could be used as a starting point to produce an accurate representation of the flux switching motor.

In the initial model, the MOSFET's used were IRF1404's which were downloaded from the International Rectifier website and are the same as those used for the experimental work carried out on the motor already constructed. The MOSFET's were switched by using voltage sources linked to .csv files which contain a time base and on/off voltages of either 10 or zero volts. The gate voltages were synchronised with a cut-off time of 168μs given by work carried out previously. [7] This cut-off time is when all the MOSFET's are off to ensure there is no overlap of 'on times' which would cause a short circuit. The inbuilt diode in the MOSFET package was also used as the freewheel diode in parallel with the field resistance and inductance. This allows current to return to the supply without passing back through the field winding.

The field and armature windings were modelled as a resistor in series with an inductor. The values of these components were derived from available TRW data. [7] The inductance values were found from finite element analysis which was used to produce an average inductance value from each winding. This was done by dividing the coil flux per metre by the ampere turns over a 90 degree rotation of the rotor in Opera and then finding a mean value for the inductance of each winding in per turn per metre. This could then be multiplied by the stack length (0.03m) and the number of turns squared in each coil (21 in the field and 19 in the armature). The end winding
inductance was also accounted for as being about 10% of the overall inductance. The resistance values were measured using a multimeter and also include the resistance of the leads used (3mΩ per winding). The model was based on 4 sets of coils (21 field or 19 armature) in parallel.

In order to make the model accurate enough to produce meaningful predictions of new motor designs (that have not yet been built), it is necessary to account for the mutual effects between the windings. This involves using mutual inductance equations and employing feedback between the windings within the model.
3.4 Modelling Mutual Effects

The mutual effects between the field and armature give rise to a mutual inductance between the windings and result in back emf’s being induced in each winding due to current in the other. The switch on point of the power switches is determined by observing the back emf in the armature winding. This is then aligned so that it leads the armature back emf by a number of degrees to enable optimum performance through switching at as close as possible to the right time.

PSpice circuit simulation has been used previously to model mutual effects between coupled coils for a variety of applications. [18, 19]. One of the main objectives of this flux switching motor PSpice model is for it to remain as simple as possible in order to achieve quick yet reasonably accurate results. Hence the modelling of the mutual effects must be implemented in a way that allows parameters such as speed to be varied quickly and easily.

In order to model these mutual effects accurately and hence improve upon the initial model, it is necessary to consider the laws and equations governing an electrical machine.

In any electrical machine, the instantaneous voltage (v) and current (i) are related by a combination of Ohm’s and Faraday’s laws, giving rise to the following equation:

\[ v = Ri + L \frac{di}{dt} + i \frac{dL}{dt} \]  \hspace{1cm} (3.1) [20, 21]

In this equation, R is the resistance in Ohms and L is the self inductance of the coil in Henrys. In operation, L will vary due to the magnitude of the current (i.dL/dt). However, in this model, this saturation effect will be ignored as the variation is only relatively small and so a constant value of L will still be used as in the initial model. This variation in L for both field and armature windings can be seen in the inductance waveforms produced from finite element analysis in figure 3.2 below (on p39).
In a two winding machine, as is being modelled here, there are two equations which obviously become a little more complex due to the mutual effects. These equations are shown below:

\[ V_f = i_f R_f + \frac{d}{dt}(L_f i_f) + \frac{d}{dt}(M_{fa} i_a) \]  \hspace{1cm} (3.2) [20]

\[ V_a = i_a R_a + \frac{d}{dt}(L_a i_a) + \frac{d}{dt}(M_{af} i_f) \]  \hspace{1cm} (3.3) [20]

In these equations, the subscript f refers to the field winding and the subscript a refers to the armature winding, ie \( V_f \) is the voltage across the field and \( i_a \) is the current through the armature. The two voltages across the windings are now given not only by the components of that winding, but also by a component of the other winding (\( M_{ia} \) or \( M_{if} \)). So now, \( L_f \) and \( L_a \) are the coefficients of self inductance and \( M_{fa} \) and \( M_{af} \) are the coefficients of mutual inductance between the two windings.

These coefficients of mutual inductance were found using finite element analysis in Opera in the same way that the self inductances were found. This time though, to find \( M_{fa} \) the field coil flux per metre was divided by the armature ampere turns and vice versa to find \( M_{af} \). These values could then be plotted against the angle of rotation over 90 degrees (after which they would of course repeat). These waveforms of both self and mutual inductance can be seen below in figure 3.2
The inductance waveforms produced by finite element analysis show that the mutual inductance waveforms vary much more greatly than the self inductances and are almost (but not quite) sinusoidal. $M_{fa}$ and $M_{af}$ are also virtually identical. To make the model simpler without losing too much accuracy, it can be assumed that the two mutual waveforms are identical and so $M_{fa}$ and $M_{af}$ can be replaced simply with $M$.

It can be seen that there is a clear variation in self inductance. These variations are however relatively small (compared to the mutual inductance) and so, as stated earlier, the self inductances will be modelled as constant to keep the model as simple as possible. This should not create too much of a limitation in its predictive capability, but this is looked at more closely later on in this chapter when predicting the torque output in PSpice.

Although the mutual waveforms look very much sinusoidal, they are not and have straighter rising and falling sides than a sine wave. It is these straighter sides that give rise to the flat topped shape of the back emf. $M$ is a function of $\theta$ (the angle of rotation) and so the back emf is given by $i.dM/dt$. Hence, if the mutual inductance wave was a perfect sine wave, then the back emf wave would be a perfect cosine wave. Once more, in order for simplicity, the mutual inductance wave, $M$ will be...
taken to be a perfect sine wave. This will produce a limitation in the model and will reduce its accuracy by producing a sinusoidal back emf instead of the actual flat topped shape, which is characteristic. This will be investigated further later on in this chapter to see if a more accurate representation of $M$ would have a significant impact on the results produced by the model.

The above equations can now be expanded to give the following:

\[ V_f = i_f R_f + L_f \frac{di_f}{dt} + i_f \frac{dL_f}{dt} + M \frac{di_a}{dt} + i_a \frac{dM}{dt} \]  \hspace{1cm} (3.4) [20]

\[ V_a = i_a R_a + L_a \frac{di_a}{dt} + i_a \frac{dL_a}{dt} + M \frac{di_f}{dt} + i_f \frac{dM}{dt} \]  \hspace{1cm} (3.5) [20]

where the back emf is the $i.dM/dt$ part of the equations.

These are the equations that must be incorporated in the PSpice model to produce an accurate representation of the motor.

### 3.4.1 Applying the Equations in PSpice

In order to model the motor correctly, it was necessary to use some more complex PSpice components that were added to the simple initial model. The equations were broken down into their individual terms so that each term could be represented by a set of components, which could then be added together. This effectively produced two feedback circuits where the field fed back into the armature and vice versa. What follows shows how the equations were broken down and what components were used to represent the terms and create the necessary effects within the circuit.

The initial $iR$ terms are already present in the model, as are the $L.di/dt$ terms (as the self resistance and inductance) and so require no additional components. The $i.dL/dt$
component will be ignored as it has already been decided that the self inductance will be modelled as constant and hence does not change with time.

The M.d\(i/dt\) term can be modelled by using a sinusoidal voltage source to represent M and then multiply this by the voltage across the inductor (self inductance) effectively giving di/dt (ie. V/L = di/dt). The PSpice components used to create this term can be seen below in figure 3.3

![Figure 3.3 PSpice components used to give M.d\(i/dt\)](image)

The i.dM/dt term (the back emf) can be split up and simplified to make it easier to model as follows:

\[
\frac{dM}{dt} = \frac{dM}{d\theta} \frac{d\theta}{dt}
\]  

(3.6)

Leaving:

\[
i \left( \frac{dM}{d\theta} \frac{d\theta}{dt} \right)
\]  

(3.7)

which is the speed dependent back emf.
This means that $M$ can now be differentiated with respect to $\theta$ (the angle of rotation) and then multiplied by the angular velocity. Because $M$ is being modelled as a sine wave, when differentiated it will become a cosine wave, which can easily be represented by a sinusoidal voltage source with a phase difference of 90 degrees in PSpice. This can then be simply multiplied by a dc voltage source with the voltage set at the value of the angular velocity being simulated. In turn, the result from this can then be multiplied by the armature current to give the entire term. The PSpice components used to create this term can be seen below in figure 3.4.

![Figure 3.4 PSpice components used to give $i.dM/dt$]

When these two sets of components are replicated for the other winding and then added to the initial model, they produce the completed circuit shown below in figure 3.5.
Figure 3.5 Complete PSpice simulation circuit
3.5 Preparing the Simulation

Now that the circuit was complete and ready to simulate the motor with the mutual effects, it was necessary to set up the switching pattern correctly. This meant replicating the switching patterns used to obtain the results already available at the two different speeds. At the speed of 1090rpm, there was a 800μs advance (corresponding to 5 degrees) and 168μs cut-off (1 degree) and at 2175rpm there was a 600μs advance (8 degrees) and 168μs cut-off (2 degrees). The necessity for the advance and cut-off have been described in Chapter 2.

So, the back emf produced by the PSpice model would have to be obtained and then the switching square waves could be aligned to give the appropriate advance where the back emf crosses zero of either 800μs (5 degrees) or 600μs (8 degrees). In order to do this another circuit was created to allow current to be fed into the field and the voltage observed in the armature. This circuit was effectively the same as the completed circuit shown in figure 3.5, but without the power switches. It can be seen below in figure 3.6.
Figure 3.6 Back emf test circuit
As the above circuit diagram shows, a current source was used to feed 100A into the field winding and the voltage across the armature winding was obtained. This voltage was the back emf and as predicted it was sinusoidal as opposed to flat topped. Even though this was not exactly the correct shape, it could still be used to determine the advance needed on the switching square waves from where it crossed zero. It was hoped that the shape would have minimal impact on the overall results.

By running the above two circuits simultaneously in the same PSpice project, it was possible to observe the back emf waveform and the square waves fed into the MOSFET gates at the same time and hence align the switching pattern accordingly.

The square wave gate voltages and the back emf waveforms for the two speeds and corresponding switching advances can be seen below in figures 3.7 and 3.8.

Figure 3.7 Back emf and square waves at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree)
The simulation could now be run and results obtained for comparison with previous data made available from TRW. These results can be seen later on in this chapter.

### 3.6 Predicting the Torque Output in PSpice

The torque produced by a machine with two mutually coupled windings is given by the following equation:

\[
T = \frac{1}{2} i_f^2 \frac{dL_f}{d\theta} + \frac{1}{2} i_a^2 \frac{dL_a}{d\theta} + i_f i_a \frac{dM}{d\theta}
\]  \hspace{1cm} (3.8) [20]

Where \(f\) refers to the field winding and \(a\) refers to the armature winding.

However, the self inductances, \(L_f\) and \(L_a\), have been modelled as being constant as discussed earlier on in this chapter. So, this assumption therefore means that the above equation simplifies to the following:
This equation can now be implemented in PSpice to produce real time torque waveforms. Figure 3.9 shows the components used to obtain the torque.

\[ T = i_r i_a \frac{dM}{d\theta} \]  

(3.9) [20]

At this point, it is important to consider just how much of an impact the assumption of constant self inductance has on torque output. This will give some idea of the amount of torque, which would be produced in the actual motor due to the variation in self inductance, that isn’t produced by the PSpice model.

To determine this, the self inductance waveforms shown in figure 3.2 can be differentiated. As the value of \( L_a \) clearly has the greatest variation, this waveform can be differentiated and compared to \( dM/d\theta \) in order to determine the largest possible inaccuracy due to this assumption. The differentiated \( L_a \) and \( M \) waveforms are shown in figure 3.10 below.
From the above plots, it can be determined that at the points of maximum variation of self and mutual inductance, the self inductance is around 15% of the mutual inductance.

The differentiated armature self inductance can then be used in equation 3.8 along with an armature current value to determine how much torque would be produced. The TRW data displayed in chapter 2 showed that the armature current reached as high as 115A at 1090rpm, so this figure can be used to show the maximum torque that could be produced. Figure 3.11 shows the result from the following part of equation 3.8 using the differentiated armature inductance (in radians) and 115A.

\[
T = \frac{1}{2} i_a^2 \frac{dL_u}{d\theta}
\]  

(3.10)
Figure 3.11 Torque produced by armature self inductance for armature current of 115A (equivalent to 1090rpm)

The above plot shows that the torque produced by the armature self inductance is both positive and negative. At its maximum (ie. worst case scenario) it is producing a contribution of no more than 0.016Nm. This can be compared to the torque figures produced by TRW in chapter 2. These show that at 1090rpm the peak torque reaches a maximum of around 2.25Nm. Hence, in the worst case, the armature self inductance produces a torque contribution of 0.7% of the total peak torque. This figure is very small and so there will be very little impact on the PSpice torque prediction. However, if this peak occurs when the torque output is very low it may be more significant. Hence it is important to take into account the average contribution of torque produced by the self inductance. This too is very small as it is has both positive and negative components and its value from the above plot is $8.8 \times 10^{-5}$Nm. This value is negligible when compared to the average overall torque, which is around 1Nm at 1090rpm.
3.7 Improving the Model

As stated previously in this chapter, the assumption that the mutual inductance waveform (M) is a perfect sine wave may impact on the results produced by the simulation. Due to this assumption, the back emf present in the model is sinusoidal rather than the characteristic flat topped waveform observed practically in the actual motor. It is unlikely that this will impact significantly on the simulated waveforms produced or the predictive capability of the model. Nevertheless, it is necessary to determine just how much impact this assumption has. This can be done using curve fitting to determine the higher order harmonics of the mutual inductance wave by Fourier transform. These can then be differentiated (to obtain dM/dθ) and placed in series in the PSpice model to produce a much more accurate representation of the mutual inductance than a simple sine wave.

The mutual inductance wave is made up of component harmonics in the following form:

\[ e_a + e_b \sin(\omega t) + e_c \sin(2\omega t) + e_d \sin(3\omega t) + e_e \sin(4\omega t) + f_b \cos(\omega t) + f_c \cos(2\omega t) + f_d \cos(3\omega t) + f_e \cos(4\omega t) \]

Using a Fourier curve fitting approach, the values of the above terms can be found so that the resultant wave almost exactly matches the mutual inductance wave, where \( e_a \) is the DC term followed by the sine and cosine component terms at increasing harmonic frequencies.

The terms were found to be as follows:

\[
\begin{align*}
e_a & = 2.89E-08 \\
e_b & = -6.35E-07 \\
e_c & = 8.08E-08 \\
e_d & = 3.25E-08 \\
e_e & = -3.17E-08 \\
e_f & = 3.36E-06 \\
e_c & = -1.39E-08 \\
e_d & = -7.93E-08 \\
e_e & = -1.66E-08
\end{align*}
\]
The actual mutual inductance waveform and the curve fitted waveform of the above composition, which uses 90 data points, can be seen in figure 3.12.

![Figure 3.12 Actual mutual inductance waveform and Fourier fitted waveform](image)

**Figure 3.12 Actual mutual inductance waveform and Fourier fitted waveform**

The waveforms shown in figure 3.12 are very similar and so this Fourier fitted wave is a good approximation of the actual mutual inductance waveform $M$.

The component frequencies could then be split up into individual terms and differentiated. Then it was necessary to multiply by the number of armature turns (19) and field turns (21) and the stack length (0.03) as the above waveforms are in inductance per turn per metre. Each harmonic component could then be replicated in PSpice by simply using a sinusoidal voltage source with the corresponding amplitude and frequency. Hence, this meant using 8 voltage sources in series to represent the mutual inductance instead of just one source used for the simple sine wave representation as before. The phase relationship of the voltage sources was changed accordingly to represent the sine and cosine terms. The components used to represent $M$ and $\frac{dM}{d\theta}$ can be seen in figure 3.13 below.
3.8 Comparing PSpice Results

The more complex PSpice model using harmonic frequencies to represent mutual effects can now be compared with the simple model to see how much the results are affected by this more accurate modelling method.

This can be seen in the figures below which compare current, voltage and mutual waveforms for the two PSpice models along with data already made available for TRW. Two different speeds, 1090 and 2175rpm were used to allow direct comparison with the TRW data.
Figure 3.14 Back emf and switching square waves at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.15 Back emf and switching square waves at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
The above figures clearly show the difference in shape of the back emf produced by the two PSpice models. This back emf appears as a sinusoidal wave in the simpler model, which assumes the mutual inductance waveforms are sinusoidal. The model using harmonic components to represent the mutual inductance waveforms produces a back emf with more of a flat top. This is closer to the back emf waveforms observed experimentally for the prototype motor. Obviously this means that the more complex PSpice model is a better representation of the actual motor. However, it is likely that this increased complexity is unnecessary and that the shape of the back emf waveform here will not impact too greatly on the torque prediction.

Figures 3.16 and 3.17 show the mutual inductance waveforms at 1090rpm for the two PSpice models.
Figure 3.16 $M$ and $\frac{dM}{d\theta}$ at 1090rpm, 800$\mu$s advance (5 degrees), 168$\mu$s cut-off (1 degree) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.17 $M$ and $\frac{dM}{d\theta}$ at 1090rpm, 800$\mu$s advance (5 degrees), 168$\mu$s cut-off (1 degree) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
It can be seen from the above figures that there is very little noticeable difference between the mutual inductance waveforms, M. In both the simple sinusoidal and more complex (using harmonics) PSpice models, the M waveform appears to be sinusoidal to the naked eye. However, when differentiated with respect to rotor position, the difference between the two models can be seen. The simple model obviously produces a perfect cosine wave while the more complex model produces a distorted waveform due to the M waveform not being a perfect sinusoid. Again, it is unlikely that this subtle difference, although clearly visible here, will impact greatly upon the torque results produced by the two models.

Figures 3.18 and 3.19 show the voltages produced across the field and armature windings at 1090rpm for both the PSpice models.
Figure 3.18 Field and armature voltages at 1090rpm, 800\(\mu\)s advance (5 degrees), 168\(\mu\)s cut-off (1 degree) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.19 Field and armature voltages at 1090rpm, 800\(\mu\)s advance (5 degrees), 168\(\mu\)s cut-off (1 degree) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
The voltage waveforms above show very little difference between the two PSpice models. Both exhibit field and armature voltages that are roughly the same magnitude at just under 5V in the armature and 4V in the field, except for spikes when switching occurs. The armature has a bipolar voltage, while the field remains mainly positive except just after switching. Both models produce waveforms that are also very similar in shape, with only a slight phase difference which is due to arbitrary starting points for the two simulations. The difference in the modelling of the mutual effects clearly does not have a great influence in the voltages observed across the windings.

Figures 3.20 and 3.21 show the field and armature currents produced at 1090rpm by the two PSpice models. Figures 3.22 and 3.23 show simulated and experimental field and armature currents at 1090rpm, produced by TRW for comparison.
Figure 3.20 Field and armature currents at 1090rpm, 800\(\mu\)s advance (5 degrees), 168\(\mu\)s cut-off (1 degree) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.21 Field and armature currents at 1090rpm, 800\(\mu\)s advance (5 degrees), 168\(\mu\)s cut-off (1 Degree) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
Figure 3.22 Previously available TRW simulated waveforms at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree), warm conditions [7]

Figure 3.23 Previously available TRW practical waveforms at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) [7]

CH1 – Position
CH2 – DC ECU voltage (5V/div)
CH3 – Armature current (50A/div)
CH4 – Battery current (50A/div)
The above current waveforms show that the two PSpice models produce very similar results with the same magnitude and only a slight difference in shape. Again, the more complex modelling of the mutual effects using harmonic components has little impact when compared to the simple sinusoidally modelled simulation.

When the PSpice results are compared to the TRW simulation results, the waveforms are of similar shape and magnitude. Both simulations peak at around 115A. Although the shapes are not exactly the same, with the TRW simulation having a higher peak just before switching, they are a good representation. This difference could be down to subtle differences in the switching times. The original TRW data is not available for direct comparison and so there may well be slight errors here as well as in the PSpice model.

When compared to the experimental armature current in figure 3.23, the PSpice waveforms are again very similar in terms of magnitude and shape with a peak just over 100A.

Figures 3.24 and 3.25 show the torque waveforms produced by the two PSpice models at 1090rpm along with the average torque. Figure 3.26 shows simulated torque and average torque waveforms at 1090rpm for comparison.
Figure 3.24 Torque at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) obtained from the simple PSpice model representing mutual effects as sinusoidal waves

Average torque = 0.94Nm

Figure 3.25 Torque at 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) obtained from the more accurate PSpice model using harmonic components to represent mutual effects

Average torque = 0.84Nm
The torque waveforms produced by the two PSpice models are very similar in terms of magnitude, peaking around 1.75N, but differ somewhat in shape. The simple PSpice model produces a more symmetric waveform compared to the model using harmonics to represent the mutual effects. Clearly the sinusoidal assumption has an effect on these results, but not to a great extent. Both sets of torque waveforms go slightly negative. This is most likely to be down to the simulation not being switched at exactly the suitable position required for optimised performance. Again, there may be unknown errors in the TRW data as well as the PSpice simulation which result in a less than optimised switching pattern.

When compared to the TRW simulated torque, the PSpice torque is obviously quite different. The TRW simulation has alternating peaks of higher and lower torque unlike the PSpice waveforms. If the torque produced in “warm” conditions (the green line) for the TRW simulation is compared to the PSpice torque, then the magnitudes are similar, peaking at 1.7-1.9N and 1.75N respectively. In terms of average torque, the TRW simulation produces a steady state average of around 1N in warm
conditions. This is in comparison to 0.94 and 0.84 for the simple and harmonic PSpice models respectively. In this case, the simple PSpice model is within 10% of the TRW simulation.

Although the PSpice torque waveforms are not a very close match to the TRW simulated waveforms, the simple model produced results within 10% in terms of magnitude. This would be good enough to give a useful impression of a motor's performance and make the PSpice simulation a viable initial model when developing new designs.

The following figures show results similar to those displayed on previous pages, but this time for a speed of 2175rpm.
Figure 3.27 Back emf and switching square waves at 2175 rpm, 600 µs advance (8 degrees), 168 µs cut-off (2 degrees) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.28 Back emf and switching square waves at 2175 rpm, 600 µs advance (8 degrees), 168 µs cut-off (2 degrees) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
The back emf waveforms for the two PSpice models are similar to those produced at 1090rpm. The simple model produces a sinusoidal back emf, while the model using harmonic components to represent mutual effects produces a back emf with more of a flat top. As before, this is closer to the experimentally observed back emf, but this difference is expected to have little effect on current and voltage waveforms.

Figures 3.29 and 3.30 show the mutual inductance waveforms for the two PSpice models at 2175rpm.
Figure 3.29 $M$ and $dM/d\theta$ at 2175 rpm, 600 $\mu$s advance (8 degrees), 168$\mu$s cut-off (2 degrees) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.30 $M$ and $dM/d\theta$ at 2175 rpm, 600 $\mu$s advance (8 degrees), 168$\mu$s cut-off (2 degrees) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
Once more these mutual inductance waveforms are similar to those produced at 1090rpm, with very little difference between the two models. When differentiated though, it is clear that the harmonically represented mutual inductance is different to the sinusoidal representation in the simple model. It would be expected though that, as with the results at 1090rpm, this will only have a minor impact on current and voltage waveforms.

Figures 3.31 and 3.32 show the voltages produced across the field and armature windings for the two PSpice models at 2175rpm. Figure 3.33 shows the voltages produced by a TRW simulation at the same speed for comparison. (No TRW voltage results were available at 1090rpm).
Figure 3.31 Field and armature voltages at 2175 rpm, 600 μs advance (8 degrees), 168μs cut-off (2 degrees) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.32 Field and armature voltages at 2175 rpm, 600 μs advance (8 degrees), 168μs cut-off (2 degrees) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
Figure 3.33 Previously available TRW simulated winding and supply voltages, 2175 rpm warm conditions [7]
(Note: winding voltages include resistance drop in leads and end inductances.)

The PSpice voltage waveforms are again very similar in magnitude and shape, just as they were at 1090rpm. This time the armature voltage is around 3 to 5V and the field 2.5 to 5V, except for switching spikes.

When compared to the TRW results, the PSpice waveforms are similar in terms of both shape and magnitude. The TRW voltages are a little higher, with more constant values of field and armature of around 4 and 6V respectively, except just after switching. As in the PSpice model, the field winding goes briefly negative immediately after switching.

Figures 3.34 and 3.35 show current waveforms produced by the two PSpice models at 2175rpm. Figure 3.36 shows the current waveforms produced in the TRW simulation at 2175rpm and figure 3.37 shows experimental current waveforms for comparison at the same speed.
Figure 3.34 Field and armature currents at 2175 rpm, 600 $\mu$s advance (8 degrees), 168$\mu$s cut-off (2 degrees) obtained from the simple PSpice model representing mutual effects as sinusoidal waves.

Figure 3.35 Field and armature currents at 2175 rpm, 600 $\mu$s advance (8 degrees), 168$\mu$s cut-off (2 degrees) obtained from the more accurate PSpice model using harmonic components to represent mutual effects.
Figure 3.36 Previously available TRW simulated waveforms at 0.8 Nm torque, 2175 rpm, 600 μs advance (8 degrees), 168 μs cut-off (2 degrees), warm conditions [7]

Figure 3.37 Previously available TRW practical waveforms at 0.8 Nm torque, 2175 rpm, 600 μs advance (8 degrees), 168 μs cut-off (2 degrees) [7]

CH1 – Position
CH2 – DC ECU voltage (5V/div)
CH3 – Armature current (50A/div)
CH4 – Battery current (50A/div)
The two PSpice models again produce current waveforms that are very similar indeed with very little difference in both shape and magnitude. The field and armature currents peak at just under 90A for both models. This is of similar magnitude to the TRW simulation waveforms in figure 3.36, which peak a little higher at around 95A. However, the TRW simulation produces current waveforms with more variation at the peak level when compared to the PSpice waveforms, which have a flatter top. In between switching, the TRW waveforms drop as low as 65A and go as high as 95A, whereas the PSpice waveforms never drop below 80A in between switching. The experimental results in figure 3.37 are closer to the PSpice waveforms with their flatter tops and not much variation around 80A in between switching on the positive half of the armature current. However, more variation occurs on the negative half which is closer to the TRW simulation, although only drops to around 75A compared to 65A in the simulation. In general, the PSpice current waveforms are a good representation in terms of magnitude and shape, being comparable to both previous simulation and experimental results.

Figures 3.38 and 3.39 show the torque waveforms and average values produced by the two PSpice models at 2175rpm. Figure 3.40 shows torque waveforms produced by the TRW simulation at the same speed.
Average torque = 0.51Nm

Figure 3.38 Torque at 2175 rpm, 600 µs advance (8 degrees), 168µs cut-off (2 degrees) obtained from the simple PSpice model representing mutual effects as sinusoidal waves

Average torque = 0.52Nm

Figure 3.39 Torque at 2175 rpm, 600 µs advance (8 degrees), 168µs cut-off (2 degrees) obtained from the more accurate PSpice model using harmonic components to represent mutual effects
Figure 3.40 Previously available TRW simulated torque and average torque waveforms at 2175 rpm [7]

The PSpice torque waveforms are again similar to those produced at 1090rpm. Both are of similar magnitude peaking at around 1.1N. As expected from the lower speed torque results, the use of harmonics to model mutual effects produces a different shape of waveform, which is less sinusoidal than in the simple model. The torque in the more complex model also goes more negative, down to around -1.5N compared to -0.75N in the simple model. This is most likely down to subtle differences in switching as is the fact that the TRW simulated torque remains positive. Again, the original TRW data is not available and so may not be entirely accurate.

At this speed, the TRW simulation predicts higher peak and average torque than the PSpice simulation. When compared in ‘warm’ conditions (green line in figure 3.40) the TRW simulation peaks as high as 1.5N compared to 1.1N in the PSpice simulation. The TRW simulation produces a steady state average torque of around 0.75N, while this is more like 0.5N in the PSpice models. The PSpice models are unable to model the alternating shape of the waveforms in the TRW simulation. The harmonic PSpice model does produce different shaped waveforms on alternating...
cycles, but they are all of similar magnitude unlike the TRW model, which peak at 1.5N and then around 1.2N on alternate cycles.

It is clear that at this higher speed of 2175rpm, the PSpice model is not as close a representation of the TRW model as it is at the lower speed of 1090rpm. In terms of average torque the PSpice models produce around two thirds that of the TRW model. Some of this difference could be put down to inaccuracies, both in the TRW model and the PSpice models, in terms of switching patterns.

3.9 Conclusions from the PSpice Work

The above results show that the two PSpice models clearly produce very similar results as already stated. There are some differences evident in the shape of some of the waveforms due to the more accurate representation of mutual effects. However, the magnitudes compare favourably with the previous data made available by TRW; in particular the field and armature voltages and currents are very close. It can be noted that the torque produced in the TRW simulations is always positive, whereas the torque produced in both PSpice simulations does go negative at points and is generally smaller in magnitude. This is most likely to be down to differences in switching times as opposed to torque lost due to the assumption of constant self inductance. The results produced by TRW cannot be guaranteed as accurate due to the original data not being available. This can also be coupled with inaccuracies of the switching in PSpice, which will not be 100% accurate. The self inductances of the windings were found to produce very small amounts of positive and negative torque. The maximum error in the torque output of the model due to the assumption of constant self inductance was found to be 0.7%, which is negligible.

By modelling the mutual effects more accurately using harmonics to represent the mutual inductance waveforms, the simulation was made more accurate. However, this made it much more complex and time consuming to change parameters such as speed. The purpose of this model is for it to be simple, quick and easy to use so that an initial idea of performance can be gained for a new design. The number of turns in each winding can be varied and run at numerous speeds in just a few minutes. This would
take hours using finite element time stepping analysis. The PSpice model may not be as accurate as using finite element time stepping analysis, which would give a closer prediction of performance of a new design not yet built, but it will work as a good starting point. The PSpice model can be used initially to give a basic idea of how a new design will operate, allowing parameters to be changed quickly, and then time stepping FEA can be used to look more accurately at a chosen design. In this way, a lot of development time can be saved.

The simple PSpice model can now be used to predict the performance of the new designs developed for this application to allow an initial comparison with the current design. Time stepping finite element analysis can then be used to obtain more accurate predictions and determine the accuracy of the PSpice model.
Chapter 4
Finite Element Analysis of the Flux Switching Motor in Opera

4.1 Introduction

This chapter details the method of using static finite element analysis, introduced in chapter 2, to improve an existing motor design. The method involves investigating the geometric design of the laminations by varying parameters on the stator and the rotor with the aim to produce a higher torque output. The ultimate objective is to produce a new improved design with higher torque capability than the original.

4.2 Investigating the Motor Laminations

A prototype motor developed by Prof. C. Pollock in conjunction with TRW has already been presented in previous chapters. The aim of this research was to build upon that model and improve its performance. This chapter details the steps initially taken to improve upon the design using static FEA.

A control file was written to allow finite element analysis over 90 degrees rotation of the rotor in steps of 4.5 degrees for both negative and positive armature current. This control file set all the parameters determining the geometry of the lamination for the particular design. It also stated the winding configuration (21 field turns and 19 armature turns) and the amount of current input. The current value used was 12.5A in each field coil (21 turns) and +/-12.5A as appropriate in each armature coil (19 turns). As there are 4 of these coils in parallel in each winding, each slot area therefore has two coils going through it. So, each field slot has 42 turns with 25A and each armature slot has 38 turns with +/-25A. This current value input is taken by the Opera model and converted into the appropriate current density for each slot area. The control file was read into Opera in a .txt format and the finite element analysis was performed.

Various parts of the motor design were varied including the shape of the stator and rotor and the size of the airgap. Each variation was done in turn over 90 degrees of rotation and for positive and negative armature currents. A set of results was obtained.
for each varied parameter and these were read into Excel from the .csr text file produced by Opera. These results were then arranged into a suitable format and graphs were then produced for each set plotting torque per metre against rotor angle for each increment of parameter change. This made it possible to tell how much each parameter affected the torque produced and which value of that parameter (if any) produced the highest torque output. It was found that some parameters affected the torque quite considerably from relatively small changes while others had very little effect. The sets of results were analysed as each was completed so that they could be used by the next simulation to build on any improvements in torque production.

The average torque per metre for each parameter variation in each set of results was calculated within the same Excel spreadsheet. Average torque per metre against parameter value was plotted to show more accurately which value (if any) gave the overall highest torque value.

This method sought to work through a system of changing parameters one at a time in an organised method of trial and error. There are other methods available, such as using genetic algorithms to reach a solution [22-31] and these too were considered. However, the genetic algorithms approach is a more random way of generating a solution where, for each iteration, all parameters are randomly varied. This means that a huge number of unnecessary iterations would be produced by varying parameters that are known from initial intuitive tests to have little impact on the results. For example, parameters far away from the airgap were quickly identified as having very little effect on torque production. Hence, although this intuitive method may not produce a definitive optimal design, it is more efficient than the long winded approach taken by genetic algorithms which would require considerable setting up time and the production of vast amounts of data, much of which would be unnecessary.

The process is illustrated in the flow chart below:
Choose parameter to vary

Perform static FEA in Opera with a range of values for that parameter

Determine value of the parameter that produces most desirable torque output

Hold parameter at that value

Choose another parameter to vary

Perform static FEA in Opera with a range of values for that parameter

Does this parameter affect the results from previous parameters?

Yes

Vary previous affected parameters together in Opera to find an optimal combination

No

Have all parameters been investigated?

Yes

Design is optimal
4.3 Crossover Torque

The various torque patterns obtained were compared in terms of average and rms torque, peak torque and how much crossover torque they produced. Crossover torque is the torque where the positive and negative torque waveforms cross as switching occurs from one to the other. The crossover torque is important in that it has to be sufficient enough to allow the motor to start. It is the aim to maximise this crossover area under the two torque profiles in order to give maximum starting torque over as many degrees as possible so that there is more flexibility in terms of when switching occurs.

The amount of crossover torque that is produced is affected predominantly by the leading edge of the rotor in terms of its shape and how much it protrudes from the main rotor body. By observing flux plots from simulations performed in Opera, which show the lines of flux and what paths they take through the stator, rotor and airgap, it can be clearly seen why this is the case. The following diagrams in figure 4.1 show a simple 4/2 (stator/rotor) design with very different leading edges on the rotor. The lines of flux plotted on them show how the flux paths differ and how the rotor teeth are attracted (or not) to the corresponding stator teeth.

Figure 4.1 Simple flux switching motor diagrams illustrating the effect on flux of a protruding leading rotor edge
The design with the straight leading edge on the rotor illustrates how there may well be points during its rotation cycle when the rotor would not be able to start. This can be seen from the position shown in the diagram above, where there is very little attraction from the next stator pole to begin its rotation. The flux lines here are still widely spaced and hence attraction is just as likely from the stator pole the rotor is just passing. This is a significant problem here, because if the rotor stopped in one of these positions, then it would not be able to start again without manual intervention, which is obviously unacceptable. However, the design with the protruding pointed leading edge on the rotor has much more attraction to the next stator pole when in the same position as the straight leading edge design. The flux lines here are more concentrated on the leading edge and so the rotor will experience a greater pull to the next stator pole than from the pole it is currently passing. Hence this design would always be able to self start no matter what position the rotor finished in.

In the slightly more complex 8/4 (stator/rotor) tooth designs under investigation here, several examples were discovered where changes made to the rotor produced results where no crossover was present. An example can be seen below in the two torque profiles where figure 4.2 has no crossover and figure 4.3 does.

![Torque Profile Diagram](image)

**Figure 4.2** Example of the torque profile of a motor lamination design which does not produce any crossover torque
Figure 4.3 Example of the torque profile of a motor lamination design which does produce crossover torque

4.4 Comparing Torque Output for Parameter Variations

Changing different parameters within the design would affect the torque output in various ways and so in order to compare these variations fairly, it was necessary to consider how the current density within the windings was affected due to the winding areas being made larger or smaller. For example, the rotor outer diameter was varied and, initially, the current in the windings was kept constant and hence the current densities would change as the rotor got larger or smaller due to the fact that the stator winding areas would change in size accordingly. This is illustrated by the following equation:

\[ J = \frac{I}{A} \]  \hspace{1cm} (4.1)

Where current density, J is the current, I per unit area, A.

Hence, a decrease in area would increase the current density and vice versa.
This meant that as the rotor got larger, the torque would obviously get higher as the current density is increased due to smaller area. However, in practice there would be a trade off as the winding area is reduced and hence to more accurately model this, it was necessary to keep the current density the same for each parameter value. Two new variables were introduced into the design; densa and densf as the current density values of armature and field respectively. These values were set at 5e6 A/m² (roughly equivalent to the value of current density achieved by inputting current values of 12.5A in the original design) and used to replace the current density calculations, which previously used the current values input for field and armature. Two different .dem files were used for different polarities of armature current density to give the positive and negative results. By comparing torque output for varying different parameters at constant current density, a fair comparison could be achieved even if the size of the winding areas was altered.

This constant current density comparison was a good starting comparison for comparing the torque output as the design was altered. However, later on in the investigation, both copper and iron losses were taken into account. This is detailed in chapter 5 where the most promising designs are compared more closely. At this point it was necessary to compare the designs at constant copper loss, so that an overall comparison taking into account losses could be achieved. In order to do this, the value of $J^2A$ was kept constant across the designs and this is explained in chapter 5.
4.5 Finite Element Simulation Results

The initial configuration of a flux switching motor used for the starting point of the design can be seen in figure 4.4. The torque curves produced by this design for both positive and negative armature current can be seen in figure 4.5.

![Figure 4.4 The original motor design – the starting point for investigation](image)

![Figure 4.5 Torque per metre against rotor angle for positive and negative armature current for initial design](image)

The above torque graphs show how there is some crossover torque as the current is switched (in terms of positive torque). This design was then taken and a wide selection of the rotor parameters were varied to observe the impact on torque.
production. An example of the parameter being varied along with the graphical results produced can be seen below in figures 4.6 to 4.8.

Figure 4.6 Varying the arc/pitch of the rotor at the airgap (label PA01) from 0.2 to 0.6 in increments of 0.05 (i.e. starting with the configuration on the left and ending with the one on the right.) A zoomed in view is also shown, so it can be seen more clearly what the parameter variation is doing.
Figure 4.7 Plots showing torque production for variation of rotor arc/pitch at airgap (described as ROTOR_TOOTH_LEADING PA01) from 0.2 to 0.6 for positive and negative armature currents of +/-12.5A.
Figure 4.8 Plots of average torque produced (for positive half of torque curve) for variation of rotor arc/pitch at airgap (PA01) for positive and negative armature current +/-12.5A

Similar plots were produced for the variation of several other parameters to give some indication of which was most effective in changing the average torque produced by the motor.
The following pages show some of the other parameters whose values were varied in a similar way to the example given above. A zoomed in view is also shown to highlight more clearly what the parameter changes are doing.

A full list of all the parameters investigated with a description of the parameter, the values it was varied over and the effect each had on the torque produced is tabulated in the appendix (p.199).

Figure 4.9 PA02 Fraction of rotor radius to leading edge. This effectively determines how close to the airgap the pointed leading edge of the rotor is, from far away on the left to at the airgap on the right.
Figure 4.10 PA03 Profile of leading edge near airgap. This parameter determines the thickness of the point on the leading edge of the rotor, from blended to the rest of the leading edge on the right, to a well defined narrow point on the left.
Figure 4.11 PA04 Profile of leading edge midway between rotor base and airgap. This parameter determines to what extent the leading edge of the rotor protrudes further away from the airgap, from an extreme indentation on the left to an extreme protraction on the right.
Figure 4.12 PA05 Leading edge near rotor base. This parameter determines the position of where the rotor meets the base and how far advanced that point is to the next tooth, from a narrow position on the left to a more advanced and wider position on the right.
Figure 4.13 PA06 Profile of trailing edge near airgap. This parameter determines how protruded the trailing rotor edge is near to the airgap, from an indented position on the left to a protruded position on the right.
Figure 4.14 PA07 Profile of trailing edge near rotor base. This parameter determines how the trailing rotor edge is shaped near to the base, from a protruding position on the left (leading to an indent where the tooth meets the base) to an indent on the right.
4.5.1 Introducing a Step in the Rotor Leading Edge

The original design of the rotor included a protruding slope on the leading edge. As described previously this allows for a substantial crossover torque in order for the motor to self start and then maintain as high a level of torque as possible throughout the rotation cycle. The idea of replacing this slope with a step (as shown below in figure 4.16) hopes to bring about a more square shape to the torque profile. This means that, as the switch from positive to negative current occurs, the decrease in torque is less gradual and will be closer to the overlapping of two square waves which would be ideally desired. It was hoped that this would significantly increase the crossover torque and indeed the overall average torque throughout the cycle. Rotor designs of this style have been used before in reluctance drives, particularly by the inventor J. V. Byrne (among others), who has also introduced a number of holes in some rotor designs. [32-35]. A rotor design with a step in the leading edge can be seen below in figure 4.16.
Several variations were investigated for this type of rotor design. For example the size of the step was varied in terms of depth (ie distance from the airgap) and how far it protruded from the main body of the leading edge. Various other “cut-out” style designs were also considered, where the leading edge cut back on itself at varying angles and even leaving the leading edge straight with a cut out positioned further back on the rotor where it meets the airgap. Examples of these designs can be seen below in figure 4.17.

Figure 4.16 Motor design with a step in the leading edge of the rotor along with a close up view of the rotor leading edge

Figure 4.17 Two examples of the designs investigated using steps and cut-outs in the rotor
4.6 Results from Analysis Using Opera

Several sets of results were obtained using Opera, similar to those examples displayed earlier on in this chapter. Many different parameters were varied in multiple ways, using a combination of favourable values recognised from previous simulations to develop subsequent changes. These results demonstrated how the torque profile produced by a particular design can be altered significantly.

It is very clear that the parameters near the airgap have by far the greatest effect on the torque profile produced, while those near the base of the rotor have very little impact. This is demonstrated when the flux is plotted as lines of potential and these were observed for all the designs investigated. This shows how great the attraction between the significant teeth is (ie the pull of the rotor tooth to the energised stator tooth) and obviously any changes near the airgap will have an impact on this – particularly on the leading edge of the rotor and the edges of the stator teeth.

4.7 Comparing the Designs Investigated

The original design used as a starting point for this investigation has already been introduced. Ultimately, any potential designs indicating improvements have to be compared to the original. At this initial stage of development, the comparison concentrated on the torque output by the motor, with peak and average torque being obvious benchmarks as well as the crossover torque, which plays an important role also. Chapter 5 looks at comparisons in greater detail between potentially optimal designs chosen and takes into account copper and iron losses.

The section below shows the original design once more and displays the torque profile that it has produced. These torque results have been produced using Opera simulation using both constant current and constant current density, the reasons for which were described earlier in the chapter. Along with the torque profiles are also the simulated flux patterns produced in the field and armature, which have been used to produce a simulated back emf. This is then compared to an experimental back emf waveform. From the investigation, a potentially improved design was then chosen. This "stepped rotor" design was proposed as a more optimal solution as the investigation showed
that it produced higher peak and average torque as well as crossover torque. The results for this are displayed after the original design for comparison.

4.7.1 Original 4 Tooth Design

![Lamination profile of original 4 tooth design](image)

**Figure 4.18 Lamination profile of original 4 tooth design**

The following figure 4.19 shows torque per metre against rotor angle for constant field and armature current densities set at 5e6 A/m² and -5e6 A/m² (ie. both positive and negative cycles together) for 21 turns on the field and 19 turns on the armature.

![Torque per metre against rotor angle](image)

**Figure 4.19 Torque per metre against rotor angle for original design at constant current density of +/-5e6A/m²**
Figure 4.20 below shows the resultant torque pattern produced when switching occurs at 43.5 degrees. This angle is the optimized switching angle for this design, which maximizes the torque production and occurs where the two torque curves cross. The rms and average value of the waveform are also shown.

![Figure 4.20 Torque per metre against rotor angle switching at 43.5 degrees for original design at constant current density of +/-5e6A/m²](image)

- **rms torque per m (N)**: 19.7
- **average torque per m (N)**: 18.3

**Figure 4.20 Torque per metre against rotor angle switching at 43.5 degrees for original design at constant current density of +/-5e6A/m²**

Figure 4.21 shows torque per metre against rotor angle as in figure 4.19 above, except that these plots were obtained by using constant current values of 12.5A and -12.5A.
Figure 4.21 Torque per metre against rotor angle for original design at constant current +/-12.5A

Figure 4.22 below shows the resultant torque pattern produced when switching occurs at the optimized switching angle of 43.5 degrees for constant current. The rms and average value of the waveform are also shown.

- rms torque per m (N) 20.7
- average torque per m (N) 19.2

Figure 4.22 Torque per metre against rotor angle switching at 43.5 degrees approx for original 4 tooth design at constant current +/-12.5A
Figure 4.23 shows flux plots of field and armature when the field winding only is excited by using the current density of $5 \times 10^6$ A/m$^2$ and there is no current in the armature.

Figure 4.24 shows the armature flux waveform produced for a 30mm stack length and 19 armature turns, obtained simply by multiplying the armature flux plot from figure 4.23 by 0.03 and 19.

Figure 4.25 shows the back emf produced at a speed of 2175 rpm. This waveform is obtained by differentiating the flux wave in figure 4.24 with respect to a time base corresponding to the desired speed.

Coupling: approx. 71%

**Figure 4.23** Field and armature flux against rotor angle for original design with field excitation only
This simulated back emf can be compared to an experimental back emf taken at the same speed, 2175rpm (in a .csv format) shown below in figure 4.26. The shape of the two waveforms is similar, but not exact, with the experimental having a slightly higher peak at just under 4V compared to the simulation at 3.5V. The differences can be put down to limitations in the static model as rotational effects and eddy currents are not accounted for. There may also be inaccuracies in the way the back emf was measured, such as differences in exact speed.
4.7.2 “Optimal” 4 Tooth Design

This section displays the results produced by a stepped rotor tooth design, which was found to be most favourable (in terms of torque production) of all the potential designs investigated in the optimisation process. The geometric design can be seen in figure 4.27 below.

Figure 4.26 Original design experimental back emf at 2175 rpm

Figure 4.27 Lamination profile of “optimal” 4 tooth design
The following figure 4.28 shows torque per metre against rotor angle for constant field and armature current densities set at 5e6 A/m² and -5e6 A/m² (ie. both positive and negative cycles together) for 21 turns on the field and 19 turns on the armature.

Figure 4.28 Torque per metre against rotor angle for optimal design at constant current density of +/-5e6A/m²

Figure 4.29 below shows the resultant torque pattern produced when switching occurs at the optimized switching angle of 43.5 degrees.
rms torque per m (N) 20.15
average torque per m (N) 18.62

Figure 4.29 Torque per metre against rotor angle switching at 43.5 degrees for optimal 4 tooth design at constant current density of +/-5e6A/m²

Figure 4.30 shows torque per metre against rotor angle as in figure 4.28 above, except that these plots were obtained by using constant current values of 12.5A and -12.5A.

Figure 4.30 Torque per metre against rotor angle for optimal design at constant current +/-12.5A
Figure 4.31 below shows the resultant torque pattern produced when switching occurs at the optimized switching angle of 43.5 degrees for constant current. The rms and average value of the waveform are also shown.

![Torque per metre against rotor angle switching at 43.5 degrees for optimal 4 tooth design at constant current +/-12.5A](image)

- **Rms torque per m (N)**: 17.9
- **Average torque per m (N)**: 16.6

**Figure 4.31 Torque per metre against rotor angle switching at 43.5 degrees for optimal 4 tooth design at constant current +/-12.5A**

Figure 4.32 shows flux plots of field and armature when the field winding only is excited by using the current density of $5e6$ A/m$^2$ and there is no current in the armature.

Figure 4.33 shows the armature flux waveform produced for a 30mm stack length and 19 armature turns, obtained simply by multiplying the armature flux plot from figure 4.32 by 0.03 and 19.

Figure 4.34 shows the back emf produced at a speed of 2175rpm. This waveform is obtained by differentiating the flux wave in figure 4.33 with respect to a time base corresponding to the desired speed.
Coupling: approx. 67%

*Figure 4.32 Field and armature flux against rotor angle for optimal design with field excitation only*

*Figure 4.33 Optimal design armature flux 30mm stack 19 turns*
4.7.3 Comparison between Original and New Design

The "optimal" 4 tooth rotor design has improved on the original design in terms of both overall average torque and crossover torque to some extent when current density is kept constant. This can be seen below in figure 4.35 which shows the respective torque curves together for easy comparison at constant current density. However, these improvements are only fairly small. The torque curves for the optimal design show how the trailing edge of the curve has a gentler slope than the steeper trailing edges of the original design, especially for the "negative" current density waveforms. A steeper trailing edge is more advantageous in terms of crossover and indeed overall torque. It is clear though that the crossover is slightly better for the optimal design and the average torque is slightly greater (around 0.3N more). The optimal design also produces higher peak torque than the original for constant current density (about 4N higher). However, when current is kept constant, the original design still produces more torque overall, although this is not an entirely accurate comparison as the size of the winding slots has changed slightly.
Figure 4.35 Torque per metre against rotor angle for original and “optimal” designs at constant current density

There is of course a trade off between the overall average torque produced and the crossover torque. By bringing the leading edge of the rotor further forward, the crossover can be increased, while the overall torque drops away. It therefore depends on how much crossover is required as to how this parameter is set.

In terms of the field and armature flux plotted, the optimal design produces very similar results to the original, but has a slightly higher level of field flux and a slightly lower value of peak armature flux than the original. Hence, the percentage of coupling is lower for the optimal design (67% compared to 71%). The back emf waveforms are also very similar, with the optimal design having a very slightly higher peak back emf indicating slightly higher power output.
4.8 Conclusions

It was found that some of the parameters altered did not significantly affect the torque production while others played a key role in how much torque is produced.

It can be concluded that, in general, the leading edge of the rotor has a greater effect on torque production than the trailing edge and that the crucial parameters determining the amount of torque produced lie close to the airgap.

In particular it was found that altering the parameter PA01, the rotor arc/pitch at the airgap, had the greatest effect on torque production with a large variation being observed between the values producing highest and lowest torque. In contrast, varying the parameter PA07, profile of trailing edge near rotor base, had very little effect, as did the parameter PA05, leading edge near rotor base. It is clear that the different parameters affect the torque profile in different ways, for example, how steeply the curve rises or falls. It is important to note that the production of crossover torque is mainly affected by the rising edge of the torque curve and hence maximising the steepness of this part of the curve will produce the greatest crossover area. In designing the rotor it is essential that the crossover area is taken into account as well as the average running torque and thus a compromise must be made in choosing parameter values to satisfy both.

The lamination design with a step in the leading edge of the rotor was found to produce greater crossover, average (0.3N more) and peak (4N higher) torque than the original when compared at constant current density. This design will be investigated further in the next chapter.
Chapter 5
Developing the 12 Tooth Rotor Model

5.1 Introduction

This chapter builds on the work detailed in chapter 4 where an initial motor design was taken and then steps were made using finite element analysis to improve it. So far only fairly minor changes have been made to the lamination profile going as far as to introduce a ‘step’ in the rotor leading edge. In this chapter, a more radical change is made by investigating a change in the number of rotor teeth from 4 to 12. The same process is adopted starting with static finite element analysis to produce a number of 12 tooth rotor designs, which can be compared to the original 4 tooth design and the chosen 4 tooth ‘step’ design. The design selection process then goes a stage further to analyse the designs with most potential in greater detail. Here, the effects of both copper and iron losses are taken into account to arrive at a conclusion about which design is most suitable.

5.2 Why Choose 12 Teeth?

In order to make further improvement to the torque produced, a change in the number of rotor teeth from 4 to 12 was proposed. This would obviously alter the torque pattern produced significantly, with 3 times more repetition than the previous 4 tooth designs leading to a higher frequency of torque ripple. This configuration of 8 stator teeth and 12 rotor teeth was chosen as it provides the correct alignment of teeth as the rotor rotates. This means that as four of the stator poles become excited, four corresponding rotor teeth will be aligned while the remaining rotor teeth will not be aligned. This design is therefore still essentially a four pole machine which could not be achieved with other configurations such as 8/6, 8/8, 8/10. Combinations of rotor teeth to stator teeth have been discussed and investigated in previous work involving switched reluctance motors. [36-38]

By increasing the number of rotor teeth, the frequency of operation is also increased. This indicates that if the flux was the same as in the 8/4 motor then there would be a
larger back emf at the same speed and hence more power due to the higher frequency. A potential disadvantage of this increased frequency is that the rise time of the armature current becomes significant. Ideally, the current should rise to its maximum value as quickly as possible, which is determined by the values of inductance and resistance of the armature winding. The higher switching frequency may mean that the current does not have time to reach this maximum. This is described by the following equation:

\[ I(t) = I_{\text{max}} (1 - e^{-\frac{t}{\tau}}) \]  

(5.1)

Where \( \tau \) is the time constant; \( \tau = \frac{L}{R} \)

Another disadvantage is that in order to avoid excessive flux leakage, it is necessary to make the teeth smaller which results in less flux. The rotor teeth are automatically made smaller due to their increased number, but the stator teeth need to be reduced in size accordingly to avoid unwanted leakage flux. This leakage flux can be seen in the following diagram in figure 5.1 where the stator teeth are too large in comparison to the rotor teeth. This means that flux passes from one stator tooth to more than one rotor tooth and hence produces low or negative torque.
Rotor tooth has passed the energised stator tooth, but flux is still passing through it from the overly large stator tooth creating negative torque.

Figure 5.1 Diagram showing 12 tooth rotor lamination design with stator teeth that are too large resulting in leakage flux and negative torque.

The purpose of this investigation into a higher number of rotor teeth is to determine if a design is possible which ensures the flux decreases by less than the frequency increases. This would therefore produce more power.

The basic operation of the 12 tooth rotor machine can be seen below in the following flux plots at different steps of the excitation cycle. Initially, the current in the armature winding could be considered to be negative from 0-9 degrees and then after this, it is switched in the opposite direction (positive). The flux lines and flux density plots in figure 5.2 below show where the torque is at a maximum and when the current is ready for switching as the poles start to saturate.
Component: BMOD (Tesla)

0 degrees – negative

3 degrees – negative

6 degrees – negative

9 degrees – negative

12 degrees – positive

15 degrees – positive
5.3 12 Tooth Rotor Designs Chosen for Comparison

The three designs shown below in figures 5.3 to 5.5 are the designs chosen after an investigation to optimise torque production both in terms of overall torque and crossover torque. The profile of the rotor teeth was varied systematically in the same way as with the 4 tooth rotor optimisation. The torque profiles of the three designs vary, each having some advantages over the others and were chosen as they were deemed to give the best all round torque profiles of all the combinations investigated. It is also clear how the stator teeth have been made narrower to accommodate for the
additional rotor teeth and these were also optimised. All the chosen designs have the same stator teeth and it is only the rotor teeth that vary. The waveforms produced by the designs can be seen together in the comparison section that follows.

Figure 5.3 12 tooth rotor design – design 1

Figure 5.4 12 tooth rotor design – design 2
5.4 Comparison of 12 Tooth Designs

The following figures 5.6 to 5.8 show torque per metre against rotor angle for constant field and armature current densities set at $5 \times 10^6$ A/m$^2$ and $-5 \times 10^6$ A/m$^2$ (i.e., both positive and negative cycles together) for 21 turns on the field and 19 turns on the armature for the three 12 tooth designs.

Figure 5.6 Torque per metre against rotor angle for 12 tooth design 1 at constant current density of $5 \times 10^6$A/m$^2$
Figure 5.7 Torque per metre against rotor angle for 12 tooth design 2 at constant current density of 5e6 A/m²

Figure 5.8 Torque per metre against rotor angle for 12 tooth design 3 at constant current density of 5e6 A/m²
These torque plots show that designs 1 and 3 produce a more symmetrical torque curve with a more distinct peak than design 2. It is evident that all three 12 tooth designs produce higher peak torque than the original 4 tooth design, with design 3 peaking at around 40N compared to the original design which peaks at less than 30N. This is an increase of around 33%.

Figure 5.9 shows the resultant torque pattern produced by the three different designs when switching occurs at optimized switching angles for each design (12 degrees for designs 1 and 3 and 13.5 degrees for design 2) at constant current density of 5e6A/m². This allows a clear comparison of torque output between the three designs at constant current density. Table 5.1 then displays the RMS and average torque per metre values.

![Figure 5.9 Torque per metre against rotor angle, switching at approx 12 degrees, for the three 12 tooth designs at constant current density of 5e6A/m²](image)

<table>
<thead>
<tr>
<th></th>
<th>Design 1</th>
<th>Design 2</th>
<th>Design 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>RMS torque per m (N)</td>
<td>24.7</td>
<td>24.9</td>
<td>28.0</td>
</tr>
<tr>
<td>Average torque per m (N)</td>
<td>23.7</td>
<td>23.3</td>
<td>26.1</td>
</tr>
</tbody>
</table>

Table 5.1 RMS and average torque per metre values produced by the three 12 tooth designs at constant current density of 5e6A/m²
It can now be seen clearly that, when compared at constant current density, design 3 produces the highest torque, while designs 1 and 2 produce very similar torque, which is around 3N lower. Hence design 3 produces around 12% more torque.

Figures 5.10 to 5.12 below show torque per metre against rotor angle as in the figures shown above for each design, except that these plots were obtained by using constant current values of 12.5A and −12.5A.

![Graph showing torque per metre against rotor angle for 12 tooth design 1 at constant current of +/-12.5A](image)

**Figure 5.10 Torque per metre against rotor angle for 12 tooth design 1 at constant current of +/-12.5A**

![Graph showing torque per metre against rotor angle for 12 tooth design 2 at constant current of +/-12.5A](image)

**Figure 5.11 Torque per metre against rotor angle for 12 tooth design 2 at constant current of +/-12.5A**
These constant current torque plots show that design 3 is producing the highest peak torque per metre of around 25N when compared to the other two 12 tooth designs which both produce around 20N. This is 25% higher. These figures are however much lower than those produced at constant current density, where design 3 was producing 40N peak torque per metre. They are also lower than those produced at constant current for the original 4 tooth design, which produces just under 30N peak torque per metre (20% higher than 12 tooth design 3).

Figure 5.13 shows the resultant torque pattern produced by the three different designs when switching occurs at approximately 12 degrees for constant current of +/-12.5A. This allows a clear comparison of torque output between the three designs at constant current. Table 5.2 then displays the RMS and average torque per metre values.
Figure 5.13 Torque per metre against rotor angle, switching at approx 12 degrees, for the three 12 tooth designs at constant current of +/-12.5A

<table>
<thead>
<tr>
<th>Design</th>
<th>RMS torque per m (N)</th>
<th>Average torque per m (N)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Design 1</td>
<td>15.1</td>
<td>14.5</td>
</tr>
<tr>
<td>Design 2</td>
<td>15.3</td>
<td>14.2</td>
</tr>
<tr>
<td>Design 3</td>
<td>17.0</td>
<td>16.0</td>
</tr>
</tbody>
</table>

Table 5.2 RMS and average torque per metre values produced by the three 12 tooth designs at constant current of +/-12.5A

These constant current results again show that design 3 produces the highest peak and average torque per metre. This time design 3 is producing average and RMS torque per metre of 1.5-2N more than the other 12 tooth designs (around 10-13% higher). However, these figures are lower than for the original 4 tooth design at constant current, which produces an average torque per metre of 19.2N. This is 20% higher than 12 tooth design 3.

Figures 5.14 to 5.16 show flux plots of field and armature when the field winding only is excited by using the current density of $5 \times 10^6$ A/m² and there is no current in the
armature for each of the three designs. The percentage coupling between the field and armature windings is also shown.

Coupling: approx. 35%

**Figure 5.14 Field and armature flux against rotor angle for 12 tooth design 1 with field excitation only**

Coupling: approx. 36%

**Figure 5.15 Field and armature flux against rotor angle for 12 tooth design 2 with field excitation only**
Coupling: approx. 36%

Figure 5.16 Field and armature flux against rotor angle for 12 tooth design 3 with field excitation only

These coupling figures for the 12 tooth designs are much lower than those for the original and optimal 4 tooth designs, where the coupling was around 70%, approximately double that of the 12 tooth designs. In order to explain why the coupling in the 12 tooth designs is so low, the flux density can be plotted on the motor lamination design along with flux lines showing where the field flux is going. This can be seen below in figure 5.17.

Figure 5.17 Flux density and flux lines plotted on the lamination of 12 tooth design 3 to show saturation and flux leakage when only the field is excited.
The above plots show that a significant amount of flux is leaking into the adjacent stator teeth. This can be seen in figure 5.17 in the un-energised stator teeth (coloured green as they are not saturated) where there are three or four flux lines present when ideally there should be no flux lines. With field only excitation, the aim is to get a 4 pole magnetic field with no flux in the other 4 teeth. In this case as the 4 energised stator teeth saturate, the flux leaks through the other 4 unsaturated teeth and hence this flux does not link the armature winding. The higher number of rotor teeth in this design mean that they cannot give a low enough alignment with the un-energised stator teeth. Due to this, the flux is just as likely to go through the gap between the unaligned rotor tooth (ie. diagonally) into the un-energised stator tooth as it is to go across the air gap between the energised teeth. These effects mean that a lower percentage of the field flux is coupled to the armature, resulting in the low coupling figures produced in figures 5.14 to 5.16. In order to improve this, it would be necessary to make the stator teeth narrower than the gap between the rotor teeth. However, this would make the teeth very narrow indeed, creating even more saturation and increasing iron loss as well as becoming structurally weak. This is therefore not a viable solution.

Figure 5.18 shows the armature flux waveforms produced by the 12 tooth designs for a 30mm stack length and 19 armature turns. These were obtained by multiplying the armature flux plots from figures 5.14 to 5.16 by 0.03 and 19 for each design.

![Armature flux for 30mm stack length and 19 armature turns for the three 12 tooth designs](image)
The armature flux waveforms for the 12 tooth designs are very similar, peaking at around 1.4mWb/m/turn. Due to the much lower coupling in these designs, the armature flux is much lower than that for the original 4 tooth design, which peaks at around 4.4mWb/m/turn. This is over 3 times greater than for the 12 tooth designs.

Figure 5.19 shows the back emf's produced at a speed of 2175rpm for each 12 tooth design. These waveforms were obtained by differentiating the flux waveforms from figure 5.18 with respect to a time base corresponding to the desired speed.

The 12 tooth back emf waveforms are similar in shape, with design 1 having a more rounded peak than designs 2 and 3. These are all in contrast to the original 4 tooth design, which exhibits a back emf with a distinctive flat top. The shape of the back emf is dependent on the shape of the motor laminations.
5.5 12 Tooth Design Summary

Table 5.3 brings together the 12 tooth designs for comparison along with the original and optimal 4 tooth designs. It shows how the torque ripple varies in each design by detailing the average to minimum and average to maximum torque per metre. The peak torque values and crossover torque areas are also displayed at constant current density.

<table>
<thead>
<tr>
<th>Design</th>
<th>Average to maximum torque per m (N)</th>
<th>Average to minimum torque per m (N)</th>
<th>Peak torque per m (N)</th>
<th>Area of crossover torque (N.degrees)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4 tooth original</td>
<td>13.7</td>
<td>9.9</td>
<td>28.3</td>
<td>17.5</td>
</tr>
<tr>
<td>4 tooth optimal</td>
<td>11.4</td>
<td>12.4</td>
<td>31.2</td>
<td>32</td>
</tr>
<tr>
<td>12 tooth design 1</td>
<td>13.5</td>
<td>10.8</td>
<td>34.5</td>
<td>27.3</td>
</tr>
<tr>
<td>12 tooth design 2</td>
<td>14.6</td>
<td>12.4</td>
<td>35.7</td>
<td>21.3</td>
</tr>
<tr>
<td>12 tooth design 3</td>
<td>19.9</td>
<td>15.7</td>
<td>41.8</td>
<td>12.6</td>
</tr>
</tbody>
</table>

Table 5.3 Average to minimum and average to maximum torque values to illustrate the extent of torque ripple in each design at constant current density of 5e6A/m². Peak torque per metre and crossover area is also shown.

The above table shows that the optimal 4 tooth and the three 12 tooth designs produce higher peak torque than the original design at constant current density. In terms of overall torque ripple, the two 4 tooth designs are roughly equal with total variations of 23.6 and 23.8 N from minimum to maximum. 12 tooth design 1 is also close to these figures with a total variation of 24.3N. The other 12 tooth designs display higher variation, with design 2 at 27N and design 3 being much higher at 35.6N minimum to maximum. If the average to maximum torque is viewed as a percentage of the maximum torque, then the designs with the highest percentage are the original 4 tooth and 12 tooth design 3, which both vary by around 48%. The lowest percentage variation from average to maximum is the optimal 4 tooth design at 36.5%, with 12 tooth design 1 and design 2 at 39% and 41% respectively. From these figures it can be concluded that the design suffering from least torque ripple is the optimal 4 tooth, while 12 tooth design 3 exhibits the most. Lower torque ripple is obviously desirable in order to produce a smoother torque output.
The area of crossover torque was calculated by assuming the crossover area is triangular. The total crossover for each design is then the total crossover for one electrical cycle (ie. two crossover areas added together). The results of this show that the optimal 4 tooth design has the greatest crossover area (32 N.degrees) and 12 tooth design 3 the least (12.6N.degrees). The other two 12 tooth designs do however produce greater crossover torque than the original 4 tooth design, with 12 tooth design 1 producing over 50% more. The 12 tooth design 3 would still produce enough crossover torque to self start, but the leeway for switching points would be reduced.

Again, it is possible with all three of these 12 tooth designs to vary the amount of crossover by bringing forward the leading edge of the rotor to increase crossover and decrease the overall torque and vice versa. Hence, these designs can be altered slightly according to how much crossover and how much overall torque is desired.

In order to determine which design is indeed the best option it is necessary to consider other factors in a more detailed comparison of all 5 designs.
5.6 Comparison of 4 Tooth and 12 Tooth Rotor Designs

Table 5.4 below shows the rms and average torque values for each of the designs for comparison. The first half of the table shows the rms and average torque produced for a constant current density of 5e6 A/m². The second half of the table shows the rms and average torque produced for a constant current of + or - 12.5A. These values were obtained by calculating the rms and average values of the switched waveforms shown for each design above.

<table>
<thead>
<tr>
<th></th>
<th>4 tooth Original</th>
<th>4 tooth Optimal</th>
<th>12 tooth Design 1</th>
<th>12 tooth Design 2</th>
<th>12 tooth Design 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>RMS torque per m (N)</td>
<td>19.75</td>
<td>20.15</td>
<td>24.71</td>
<td>24.94</td>
<td>28.04</td>
</tr>
<tr>
<td>Average torque per m (N)</td>
<td>18.34</td>
<td>18.62</td>
<td>23.72</td>
<td>23.35</td>
<td>26.15</td>
</tr>
<tr>
<td>J²A (A²/m²)</td>
<td>2.7x10⁹</td>
<td>2.95x10⁹</td>
<td>3.425x10⁹</td>
<td>3.425x10⁹</td>
<td>3.425x10⁹</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th></th>
<th>4 tooth Original</th>
<th>4 tooth Optimal</th>
<th>12 tooth Design 1</th>
<th>12 tooth Design 2</th>
<th>12 tooth Design 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>RMS torque per m (N)</td>
<td>20.71</td>
<td>17.92</td>
<td>15.09</td>
<td>15.27</td>
<td>17.04</td>
</tr>
<tr>
<td>Average torque per m (N)</td>
<td>19.22</td>
<td>16.61</td>
<td>14.50</td>
<td>14.257</td>
<td>16.04</td>
</tr>
<tr>
<td>J²A (A²/m²)</td>
<td>2.09x10⁹</td>
<td>1.91x10⁹</td>
<td>1.65x10⁹</td>
<td>1.65x10⁹</td>
<td>1.65x10⁹</td>
</tr>
</tbody>
</table>

Table 5.4 RMS and average torque values for all designs

The above table shows that when current density is constant, 12 tooth design 3 produces the highest average and RMS torque per metre. This is around 42% higher than the original 4 tooth design in terms of average and RMS. However, at constant current, the original 4 tooth design produces the highest average and RMS torque per metre. This is around 16% higher than the optimal 4 tooth and 20% higher than 12 tooth design 3, which still produces the highest average and RMS torque per metre of the 12 tooth designs.

The table also displays the J²A value for each design. Another way of comparing the designs would be to keep this value constant for all of them and then compare the torque. This would mean that the copper loss is the same for all the designs. This will be investigated later on in this chapter.
The designs can also be compared at constant torque and noting what current is required in each design to achieve that torque value. Table 5.5 below compares the designs when the average torque per metre value is kept constant at 19.2N. This is the average torque per metre value for the original 4 tooth design in table 5.4, which was obtained by using 12.5A.

<table>
<thead>
<tr>
<th>Input current (A)</th>
<th>4 tooth original</th>
<th>4 tooth optimal</th>
<th>12 tooth design 1</th>
<th>12 tooth design 2</th>
<th>12 tooth design 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>12.5</td>
<td>13.95</td>
<td>14.65</td>
<td>14.9</td>
<td>13.85</td>
<td></td>
</tr>
</tbody>
</table>

**Table 5.5 Comparison of current input to achieve average torque per metre of 19.2N**

Table 5.5 shows that in order to achieve the same average torque per metre as the original 4 tooth design, the other designs all require more input current. 12 tooth design 2 requires an extra 2.4A (over 19% extra), while the optimal 4 tooth design requires 1.45A more and 12 tooth design 3 a similar 1.35A (around 11% more). This higher input to achieve the same output indicates that all the new designs have a lower efficiency than the original design. In chapter 6, the efficiency of the best new design is compared to the original in a more detailed calculation incorporating losses.

Figure 5.20 below shows the torque curves for the original and optimal 4 tooth designs on the same graph for constant current densities of 5e6 A/m².

Figure 5.21 shows the same data for constant current of +/- 12.5A.
When compared, the torque curves for the two 4 tooth designs are of a very similar shape. At constant current density, the optimal 4 tooth design has a slightly higher peak of just over 30N compared to the original design, which peaks at around 28N and
has a squarer shape. At constant current, the original design has a higher peak torque per metre, which again is a more square shape. The optimal design is now peaking around 26N whereas the original peaks at nearer 29N. In terms of crossover, the two designs appear very similar. However, when looked at in closer detail at constant current density in table 5.3, it was shown that the optimal 4 tooth design has a considerably larger crossover area.

Figure 5.22 below shows the torque curves for the original 4 tooth and 12 tooth design on the same graph for constant current densities of 5e6 A/m². These have been plotted over one electrical cycle for a direct comparison. Figure 5.23 shows the same data for constant current of +/- 12.5A.

Figure 5.22 Torque per metre against rotor angle for original and 12 tooth design 1 at constant current density
The plots produced by 12 tooth design 1 at constant current and constant current density are very different. When compared to the original 4 tooth design at constant current density, 12 tooth design 1 produces considerably higher torque per metre, peaking at around 35N when compared to the original design which peaks at around 28N. Like the optimal 4 tooth design, this 12 tooth design has a pointed peak when compared to the flat top of the original design. At constant current, 12 tooth design 1 produces much lower torque per metre, peaking at just over 20N compared to a peak of around 29N for the original design. It is very clear at constant current that the original design produces the greater torque. However, in both cases, 12 tooth design 1 obviously produces a much larger crossover area, which was confirmed in table 5.3.

Figure 5.24 below shows the torque curves for the original 4 tooth and 12 tooth design 2 on the same graph for constant current densities of 5e6 A/m². Again, this plotted over one electrical cycle for direct comparison.

Figure 5.25 shows the same data for constant current of +/- 12.5A.
When compared to the original design, 12 tooth design 2 produces a similar outcome to 12 tooth design 1. This being that at constant current density, the 12 tooth design produces a higher peak torque than the original (around 35N compared to 28N), but at
constant current it peaks much lower at around 20N compared to the original at around 29N. Again, the 12 tooth design produces an obviously greater crossover area in both cases.

Figure 5.26 below shows the torque curves for the original 4 tooth and 12 tooth design 3 on the same graph for constant current densities of 5e6 A/m², plotted over one electrical cycle. Figure 5.27 shows the same data for constant current of +/- 12.5A.

Figure 5.26 Torque per metre against rotor angle for original and 12 tooth design 3 at constant current density

Electrical degrees
This comparison between the original 4 tooth design and 12 tooth design 3 is similar to the other 12 tooth comparisons. This time though, it is clear that design 3 produces much greater peak torque at constant current density with a peak at over 40N compared to the original at 28N. 12 tooth design 3 also produces a higher peak torque at constant current than the other two 12 tooth designs, with a peak at around 25N as opposed to 20N. This is still lower than for the original design, which peaks at 29N. Once more, the 12 tooth design produces a more pointed peak than the flat top of the original design. Crossover torque area appears similar, but table 5.3 showed that 12 tooth design 3 did in fact produce a smaller crossover area than the original design.

In general, the 12 tooth designs look more promising as alternatives to the original design when they are compared at constant current density. This of course could be considered to be the fairest comparison due to the much larger slot area of the 12 tooth designs. However, table 5.5 showed that more current was required for the 12 tooth designs to produce the same torque output as the original design, indicating lower efficiency. Further comparison will now be made by looking at losses.
5.7 Losses

In order to determine which of the designs performs best, it is necessary to compare the losses in each and then weigh this against any improvement in torque production. A simple improvement in terms of torque is no use if the losses this entails significantly outweigh it and produce a less efficient machine.

Windage and friction losses will of course contribute to the overall loss, but the most significant losses in the flux switching motor for this application will be copper and iron losses.

5.7.1 Copper Losses

The copper loss in the machine can be calculated as shown in the equations below which show how the copper loss relates to the slot area for both constant current (I) and constant current density (J). The copper loss has been calculated for per unit values compared to the original 4 tooth design, which allows a direct comparison.

Copper loss, \( P_{cu} = I^2 R \) \hspace{2cm} (5.2)

\( R = \rho l / A \) \hspace{2cm} (5.3)

\( P_{cu} = I^2 \rho l / A \) \hspace{2cm} (5.4)

\( I = JA \) \hspace{2cm} (5.5)

\( P_{cu} = J^2 A^2 \rho l / A = J^2 \rho l A \) \hspace{2cm} (5.6)

So, for constant I; \( P_{cu} \propto l/A \) \hspace{2cm} (5.7)

For constant J; \( P_{cu} \propto A \) \hspace{2cm} (5.8)

If the areas of field and armature slots for the original 4 tooth design are taken as base values then the other designs can be compared to the original in terms of copper loss by dividing the base by the optimal design for constant current and by dividing the optimal design by the base for constant current density;
Original design;  
Field slot area = 8.71e\(^{-5}\) m\(^2\)  
Armature slot area = 1.08e\(^{-4}\) m\(^2\)  

So, if copper losses are taken as 100% for the original design;

4 tooth optimal;  
Field slot area = 9.92e\(^{-5}\) m\(^2\)  
Armature slot area = 1.18e\(^{-4}\) m\(^2\)  

For constant I;  
Field = 8.71e\(^{-5}\) / 9.92e\(^{-5}\) = 88%  
Armature = 1.08e\(^{-4}\) / 1.18e\(^{-4}\) = 92%  

For Constant J;  
Field = 9.92e\(^{-5}\) / 8.71e\(^{-5}\) = 114%  
Armature = 1.18e\(^{-4}\) / 1.08e\(^{-4}\) = 109%  

Hence, when current is kept constant, the copper losses will be lower in the optimal 4 tooth design than in the original, whereas the opposite will be the case when the current density is kept constant.

12 tooth designs;  
Field slot area = 1.28e\(^{-4}\) m\(^2\)  
Armature slot area = 1.37e\(^{-4}\) m\(^2\)  

For constant I;  
Field = 8.71e\(^{-5}\) / 1.28e\(^{-4}\) = 68%  
Armature = 1.08e\(^{-4}\) / 1.37e\(^{-4}\) = 79%  

For Constant J;  
Field = 1.28e\(^{-4}\) / 8.71e\(^{-5}\) = 147%  
Armature = 1.37e\(^{-4}\) / 1.08e\(^{-4}\) = 127%  

Hence, when current is kept constant, the copper losses will be lower in the 12 tooth designs than in the original and optimal 4 tooth designs, whereas the opposite will be the case when the current density is kept constant.

However, by making a comparison with constant copper loss (i.e. constant J\(^2\)A), a more accurate picture of which design performs best can be obtained.
5.7.2 Constant Copper Loss Comparison

For this comparison, static FE was used to obtain torque waveforms using current density values according to the slot area of the design. An initial current density value was chosen for the original 4 tooth design which enabled a current of 80A – the average current value produced in the TRW simulations at 2175rpm. From this, $J^2A$ was calculated and then this value was kept constant for all the other designs by inputting the necessary J value in the field and armature according to the slot area. Hence, the copper losses were kept constant. The results can be seen in the following figures.

![Graph showing torque per metre against rotor angle for original 4 tooth design at constant copper loss ($J^2A$)](image)

average torque per metre 36.6N

**Figure 5.28 Torque per metre against rotor angle for original 4 tooth design at constant copper loss ($J^2A$)**
Figure 5.29 Torque per metre against rotor angle for optimal 4 tooth design at constant copper loss ($J^2A$)

average torque per metre $33.1N$

Figure 5.30 Torque per metre against rotor angle for 12 tooth design 1 at constant copper loss ($J^2A$)

average torque per metre $37.7N$
Figure 5.31 Torque per metre against rotor angle for 12 tooth design 2 at constant copper loss ($J^2A$)

average torque per metre 37.0N

Figure 5.32 Torque per metre against rotor angle for 12 tooth design 3 at constant copper loss ($J^2A$)

average torque per metre 42.1N
The 12 tooth design with the highest average torque (design 3) can be compared directly with the original 4 tooth design at constant copper loss by plotting the torque curves over one electrical cycle, as shown in figure 5.33 below.

![Torque curves for original 4 tooth and 12 tooth design 3 compared at constant copper loss](image)

Original 4 tooth average torque per metre 36.6N
12 tooth design 3 average torque per metre 42.1N

Figure 5.33 Torque curves for original 4 tooth and 12 tooth design 3 compared at constant copper loss

These results at constant copper loss clearly show that the 12 tooth design 3 produces both the highest peak and average torque values of all the designs. Therefore, this design still performs the best after taking into account losses due to the winding copper. However, it is also necessary to take into account the iron losses before the 12 tooth design 3 can be determined to be the best design. The next section starts to analyse this.
5.7.3 Flux Density in Each Machine

The flux density in each design was looked at carefully in 4 specific areas. This becomes important when calculating iron loss, as will become clear in the next section.

The following plots in figures 5.34 to 5.38 show the average flux density across an armature and field back iron and across a stator and rotor tooth cross section plotted against the rotor angle in degrees as the rotor rotates through 180 degrees for the 5 designs. The plots were produced using constant current density switching between 5e6A/m² and -5e6A/m² in the armature at approximately every 45 degrees for the 4 tooth rotor designs and approximately every 15 degrees for the 12 tooth rotor designs.

![Figure 5.34 Average flux densities across field and armature back irons and stator and rotor teeth for original 4 tooth design at constant current density](image)

Figure 5.34 Average flux densities across field and armature back irons and stator and rotor teeth for original 4 tooth design at constant current density
Figure 5.35 Average flux densities across field and armature back irons and stator and rotor teeth for optimal 4 tooth design at constant current density

Figure 5.36 Average flux densities across field and armature back irons and stator and rotor teeth for 12 tooth design 1 at constant current density
Figure 5.37 Average flux densities across field and armature back irons and stator and rotor teeth for 12 tooth design 2 at constant current density

Figure 5.38 Average flux densities across field and armature back irons and stator and rotor teeth for 12 tooth design 3 at constant current density
It can be noted from these flux density plots that the 12 tooth designs generally have lower flux density, particularly in the back irons, indicating that they are less close to saturation than the 4 tooth designs. There is also significantly less ripple in the field back iron for the 12 tooth designs.

5.7.4 Iron Losses

A significant proportion of the total losses dissipated in the machine can be down to iron loss. The iron loss dissipated in an electrical machine may be separated into Hysteresis and eddy current losses. [39]. Hysteresis losses occur due to non-linearity and differences in the magnetisation and demagnetisation of all Ferro magnetic materials. [40]. Eddy current losses are produced by alternating magnetic flux inducing an EMF in the iron core, which causes current to flow in any available closed path. This results in a magnetic flux, which opposes the applied magnetising field and leads to a reduced flux capacity in the machine. [41]

Iron losses are usually calculated using the Steinmetz equation shown in its basic form below in equation 5.9.

\[
Q = \eta B^{1.6} \quad (5.9) \quad [42]
\]

Where:

\( Q \) is the loss of energy per cycle per volume, \( B \) is the maximum induction and \( \eta \) is the hysteresis coefficient.

However for the flux switching machine, as with the switched reluctance machine, the flux waveforms are non-sinusoidal. Hence, different parts of the magnetic circuit have different flux waveforms [43]. Therefore, using the Steinmetz equation, as the following method involves, has its limitations.

The hysteresis losses can be calculated using the iron loss prediction model described below, which assumes a sinusoidal flux density variation.

\[
P_{HYS} = K_H \ Vol \ f \ (\Delta B_m)^n \quad (5.10) \quad [40]
\]
Where:

\( K_H \): Hysteresis constant for a given iron type, and given range of flux density.

\( \text{Vol} \): Volume of iron specimen \([\text{m}^3]\).

\( f \): Frequency \([\text{Hz}]\).

\( n \): Steinmetz index.

\( \Delta B_m \): Maximum flux density \([\text{Tesla}]\).

The eddy current losses can be calculated using the following model. This takes the approach of using the rate of change of flux density in different sections of the machine [44]. This method is well suited to non-sinusoidally distributed flux density [40]

\[
P_{\text{EDDY}} = \frac{\text{Vol} \, \tau^2 \left( \frac{dB(t)}{dt} \right)^2}{12 \rho} \quad (5.11) \quad [40]
\]

Where:

\( \tau \): Thickness of lamination.

\( \rho \): Resistivity \([\Omega\text{m}]\).

\( B(t) \): Time varying flux density in the iron core \([\text{Tesla}]\).

Using these prediction models, the hysteresis and eddy losses have been calculated for the 5 designs to provide a total iron loss figure at two different speeds; 1000rpm and 2175rpm. This was done by splitting each design up into 4 main areas; field back iron, armature back iron, stator tooth and rotor tooth. The average flux densities were calculated across each of these areas as shown in the above plots. The necessary values were then calculated and input into the above equations. This was straightforward to find the hysteresis losses, but to obtain the eddy current losses, the flux density waveforms were curve fitted to a Fourier approximation and this was then differentiated at each angular step and the eddy losses then calculated for that position. The eddy losses for each position were then added up to give the total eddy loss over 360 degrees. The iron losses at 1000rpm and 2175rpm are shown in tables 5.6 and 5.7 respectively.
<table>
<thead>
<tr>
<th>Design</th>
<th>Area</th>
<th>Hysteresis (w)</th>
<th>Eddy (w)</th>
<th>Total (w)</th>
<th>Total iron loss (w)*</th>
</tr>
</thead>
<tbody>
<tr>
<td>4 tooth original</td>
<td>Armature back iron</td>
<td>0.05</td>
<td>3.39</td>
<td>3.45</td>
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<tr>
<td></td>
<td>Field back iron</td>
<td>0.04</td>
<td>1.33</td>
<td>1.38</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.11</td>
<td>0.94</td>
<td>1.06</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.04</td>
<td>0.43</td>
<td>0.47</td>
<td>29.63</td>
</tr>
<tr>
<td>4 tooth optimal</td>
<td>Armature back iron</td>
<td>0.04</td>
<td>2.49</td>
<td>2.53</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.03</td>
<td>0.75</td>
<td>0.79</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.14</td>
<td>1.18</td>
<td>1.32</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.03</td>
<td>0.36</td>
<td>0.39</td>
<td>25.37</td>
</tr>
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<td>12 tooth design 1</td>
<td>Armature back iron</td>
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<td>6.04</td>
<td>6.10</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.06</td>
<td>0.36</td>
<td>0.41</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.18</td>
<td>3.99</td>
<td>4.17</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.02</td>
<td>0.27</td>
<td>0.30</td>
<td>62.93</td>
</tr>
<tr>
<td>12 tooth design 2</td>
<td>Armature back iron</td>
<td>0.06</td>
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<td>5.72</td>
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</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.05</td>
<td>0.36</td>
<td>0.42</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.17</td>
<td>3.81</td>
<td>3.98</td>
<td></td>
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<tr>
<td></td>
<td>Rotor tooth</td>
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<td>0.27</td>
<td>0.30</td>
<td>59.98</td>
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<td>12 tooth design 3</td>
<td>Armature back iron</td>
<td>0.06</td>
<td>5.86</td>
<td>5.92</td>
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<td>Field back iron</td>
<td>0.06</td>
<td>0.47</td>
<td>0.53</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.18</td>
<td>4.02</td>
<td>4.20</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.03</td>
<td>0.30</td>
<td>0.33</td>
<td>63.32</td>
</tr>
</tbody>
</table>

*Total iron loss is hysteresis + eddy loss for 4 x armature back iron + 4 x field back iron + 8 x stator tooth + 4 or 12 x rotor tooth

Table 5.6 Iron losses for each design at 1000rpm
<table>
<thead>
<tr>
<th>Design</th>
<th>Area</th>
<th>Hysteresis (w)</th>
<th>Eddy (w)</th>
<th>Total (w)</th>
<th>Total iron loss (w)*</th>
</tr>
</thead>
<tbody>
<tr>
<td>4 tooth original</td>
<td>Armature back iron</td>
<td>0.12</td>
<td>16.06</td>
<td>16.17</td>
<td>136.48</td>
</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.09</td>
<td>6.31</td>
<td>6.40</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.25</td>
<td>4.46</td>
<td>4.71</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.08</td>
<td>2.04</td>
<td>2.12</td>
<td></td>
</tr>
<tr>
<td>4 tooth optimal</td>
<td>Armature back iron</td>
<td>0.09</td>
<td>11.79</td>
<td>11.87</td>
<td>116.14</td>
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<td></td>
<td>Field back iron</td>
<td>0.07</td>
<td>3.56</td>
<td>3.63</td>
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<tr>
<td></td>
<td>Stator tooth</td>
<td>0.30</td>
<td>5.58</td>
<td>5.88</td>
<td></td>
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<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.07</td>
<td>1.70</td>
<td>1.77</td>
<td></td>
</tr>
<tr>
<td>12 tooth design 1</td>
<td>Armature back iron</td>
<td>0.13</td>
<td>28.58</td>
<td>28.70</td>
<td>292.23</td>
</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.13</td>
<td>1.68</td>
<td>1.81</td>
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</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.38</td>
<td>18.88</td>
<td>19.26</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.05</td>
<td>1.29</td>
<td>1.34</td>
<td></td>
</tr>
<tr>
<td>12 tooth design 2</td>
<td>Armature back iron</td>
<td>0.12</td>
<td>26.80</td>
<td>26.92</td>
<td>278.39</td>
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<td>Field back iron</td>
<td>0.12</td>
<td>1.72</td>
<td>1.84</td>
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</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.37</td>
<td>18.03</td>
<td>18.40</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.05</td>
<td>1.29</td>
<td>1.35</td>
<td></td>
</tr>
<tr>
<td>12 tooth design 3</td>
<td>Armature back iron</td>
<td>0.13</td>
<td>27.71</td>
<td>27.84</td>
<td>293.94</td>
</tr>
<tr>
<td></td>
<td>Field back iron</td>
<td>0.13</td>
<td>2.22</td>
<td>2.34</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Stator tooth</td>
<td>0.39</td>
<td>19.02</td>
<td>19.41</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rotor tooth</td>
<td>0.06</td>
<td>1.44</td>
<td>1.50</td>
<td></td>
</tr>
</tbody>
</table>

*Total iron loss is hysteresis + eddy loss for 4 x armature back iron + 4 x field back iron + 8 x stator tooth + 4 or 12 x rotor tooth

Table 5.7 Iron losses for each design at 2175rpm

The results displayed in the tables above were obtained using the following values:

Resistivity = 5.2e-7
Hysteresis constant ($K_H$) = 125
Steinmetz index = 2.05
Lamination thickness = 2e-4
Frequency = (speed*number of rotor poles)/60

It can be clearly seen that the 12 tooth models exhibit much higher iron loss than both the 4 tooth models. The iron losses for the 12 tooth designs are around twice that for the 4 tooth designs. This is the case even though the flux density plots show that the flux density in the 4 areas investigated is significantly lower for the 12 tooth designs indicating that they are not as close to saturation. It seems that the key factor here is the frequency of the rotor poles passing the stator poles. In the 12 tooth designs, the frequency is obviously higher due to there being 3 times as many rotor teeth as in the
4 tooth designs and this appears to cause the losses to double. For all the designs it is clear that the dominant component of the iron losses is the eddy current loss, with this being far higher than the hysteresis losses.

There are also two clear areas; the armature back iron and the stator tooth, where the iron losses are much greater in the 12 tooth designs. This is obviously down to the narrowing of the stator teeth combined with the enlarged slot areas, which brings about significantly increased eddy current losses. The solution here may be to increase the width of the stator tooth away from the air gap, so that it narrows towards the end and also increase the width of the armature back iron. This should improve the iron loss without compromising the torque output as the stator teeth are still narrow, as is necessary, at the airgap. This concept is detailed in the following section as a suggestion for further work.

5.7.5 Further Work – Potential Improvements to the 12 Tooth Design

A number of designs were investigated with the intention to reduce iron loss, and in particular eddy loss within the armature back iron and stator teeth, while not compromising torque production. From this investigation, the findings indicated that the proposed 12 tooth design could be improved upon with more work in the future.

The design shown below in figure 5.39 looked to increase the width of the stator teeth and the back irons with the intention to reduce the large iron losses associated with the 12 tooth designs. The teeth were tapered towards the air gap in order to keep them narrow enough at this point to limit flux leakage (as described earlier in this chapter). As a result of this, the slot areas were obviously reduced in size considerably. This would therefore lead to greater copper loss. Hence there is clearly a trade off, which any further work in this area would have to investigate – is it worth reducing the iron loss in this way if any benefit from reduced copper loss is lost?
Torque curves were obtained for this design at constant current of 12.5A and constant current density of 5e6A/m² for comparison with all the other designs. These can be seen in the following figures;

Average torque per metre = 16.45N

Figure 5.40 Torque per metre against rotor angle for 12 tooth design with wider stator teeth and back irons at constant current +/-12.5A
Average torque per metre = 16.30N

**Figure 5.41 Torque per metre against rotor angle for 12 tooth design with wider stator teeth and back irons at constant current density +/-5e6A/m²**

The above results show that this type of design does have some potential for further investigation in any future work. The torque produced is still lower than the original 4 tooth design at constant current with an average torque per metre of 16.45N compared to 19.2N and peak of 25N compared to 30N. However, it is slightly higher than the 12 tooth design 3, which has an average torque per metre of 16N at constant current. When compared at constant current density, the torque is much lower than the 12 tooth design 3 with an average torque per metre of 16.3N compared to 26.1N and a peak of 25N compared to 42N. It is again lower than the original 4 tooth, which has an average torque per metre of 18.3N and peak of 28N at constant current density. This is to be expected as the slot areas in this 12 tooth design have been reduced significantly.

One of the crossover points also exhibits very low torque when compared to previous designs. It is clear therefore that the alterations to the stator teeth have had a negative impact upon the torque produced. Any future work building on this suggestion would have to consider if the improvement to iron loss that this would bring about is worth the detrimental impact on the torque as well as increased copper loss. Future work
could investigate this type of design further to determine if the torque produced by this
design here can be improved upon.

5.8 Conclusions

In terms of the flux plots, the field and armature flux are much lower for the 12 tooth
designs compared to the 4 tooth designs. The 12 tooth designs have very similar flux
plots, with design 1 having a slightly higher field flux and hence a lower percentage of
coupling than the others. The 12 tooth designs have coupling percentages of around
half of those of the 4 tooth designs (35% compared to 70%). This is down to the
saturation of the stator teeth in the 12 tooth design and the leakage of flux.

The back emf’s of the 12 tooth designs are of a different shape to the 4 tooth designs,
with design 2 having the most pointed peak back emf. The 12 tooth designs’ back
emf’s do not exhibit the flat tops produced by the 4 tooth designs, which is mainly
down to the shape of the laminations. This is a disadvantage as a flat topped back emf
is desireable for more efficient torque production.

The 12 tooth designs exhibit improved torque production in terms of overall average
torque per metre for when current density is kept constant. The 12 tooth design 3
clearly produces the most torque with around 43% more average torque than the
original design. Design 1 and design 2 produce around 29% and 27% more average
torque respectively. When current is kept constant however, this is not the case and
the original design produces the best torque output, although this is not strictly a fair
comparison due to the slot areas being considerably larger on the 12 tooth designs.

Further adjustments could be made to the 12 tooth design to increase or decrease the
amount of crossover torque depending on exactly how much is required, which would
have the subsequent effect of decreasing or increasing the overall average torque
produced.

The 12 tooth designs have a higher frequency of torque ripple than the 4 tooth designs
as expected due to the increase in tooth number. The optimal 4 tooth design produced
the lowest torque ripple with a percentage variation from average to maximum of

156
36.5%. 12 tooth design 3 had the highest, which in terms of a percentage was the same as the original 4 tooth design at around 48%.

In terms of losses, copper losses are greater for the 12 tooth designs when compared at constant current density, with the copper losses being 37% higher on average (27% higher in the armature and 47% higher in the field) at constant current density. However, the larger slot areas in the 12 tooth designs mean that copper loss is lower than the 4 tooth designs when compared at constant current. At constant current, copper loss is 32% lower in the field and 21% lower in the armature for the 12 tooth designs when compared to the original 4 tooth design. The torque curves obtained at constant copper loss for all the designs showed that 12 tooth design 3 produced the highest peak and average torque per metre, with around 15% more average torque than the original 4 tooth design.

The iron loss varies for the different 12 tooth designs, but on average appears to be around twice that of the 4 tooth designs, which are similar. These higher losses are down to the reduced width of the stator teeth and back irons and the increase in the number of rotor teeth leading to an increased frequency of operation. This high iron loss is the main drawback of the 12 tooth designs.

As a suggestion for future work, a 12 tooth design was proposed with wider stator teeth and back irons, which would allow a reduction in iron losses. The torque production in this type of design was compromised and it would obviously have higher copper losses due to smaller slot areas. Future work could build on this to investigate if it is possible to improve the torque output, but retain the advantage of lower iron loss. It would then be a question of whether this benefit outweighs the detrimental affect of higher copper losses.

It can be concluded that 12 tooth design 3 is the most promising of the new designs in terms of torque production when compared to the original design. However, the iron loss produced in this 12 tooth design could prove to be too large for any overall benefit to be significant.
In the next chapter, 12 tooth design 3 will be investigated further. PSpice will be used to determine an optimum winding configuration and then a dynamic comparison can be made with the original 4 tooth design using time stepping finite element analysis. This will produce more accurate results than the static analysis performed so far. The data from this can then be used in conjunction with iron and copper loss data, at a number of operating speeds, to determine if the 12 tooth design has improved upon the original in any areas.
Chapter 6
Using the Rotating Machine Program to Perform Time Stepping
Finite Element Analysis of 4 Tooth and 12 Tooth Motors

6.1 Introduction

This chapter introduces a time stepping method of modelling the motor using finite element analysis. Time stepping methods have been used in previous work effectively to model dynamic conditions in permanent magnet DC, induction and permanent magnet synchronous machines.[45-50]. The flux switching motor has also been modelled using time stepping methods by TRW. [7

6.2 Using the Rotating Machine Program (RM) to Perform Time Stepping FEA

Opera’s RM time stepping FEA program has been described in chapter 2, which gave some background on what it does and how useful it can be with regard to motor design in terms of its advantages and limitations.

The RM program has been used to look at the current waveforms generated by the motor designs in order to allow comparison between the original 4 tooth design and the optimised 12 tooth design and relate these simulated results to experimental results obtained from the original motor to determine the accuracy of the model.

Initially, the original design was investigated and was set up to enable a rotating machine simulation. The windings were configured so that they were all in series in a bifilar winding arrangement. This was achieved using four different external circuits corresponding to each part of the switching cycle. Two of these circuits contained a representation of the freewheel diode to allow the current to return to the supply at the necessary point during the switching cycle. As the switching cycle progressed, the external circuits were used to replace each other as required.

The actual configuration of the motor places 4 windings in parallel each with 21 (field) or 19 (armature) turns. Hence, this model is somewhat different as the windings
are arranged in series. This is due to the way the RM software works and how the external circuits operate as they are exchanged at various points throughout the switching cycle as described above. It was found after some initial research using the RM that it was very difficult if not impossible to configure the windings in parallel. Vector Fields (the designers of Opera) were also contacted regarding this matter, but their support staff could not reach a solution. So, the windings were arranged in an equivalent series configuration. In order to obtain comparable results to the data already available, some research was performed on how exactly the windings should be represented. The result of this was to use half the turns in the RM than was actually required. So, in the actual motor there are 4 parallel windings each of 21 and 19 turns, while to represent this in the RM, there are 4 series windings each with 10.5 and 9.5 turns. The “half a turn” may seem odd, but the RM software was discovered to work with this fine and enabled results to be produced that compared very favourably with the previously available data. This can be seen below in figures 6.1 to 6.10 where the field and armature currents and torque produced can be compared between the RM simulation, previous TRW simulation and experimental data.

Figure 6.1 RM time stepping simulated field and armature current waveforms for the original 4 tooth design at 1090rpm
Figure 6.2 TRW simulated waveforms for the original 4 tooth design at 1.0Nm torque, 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree), warm conditions [7]

Figure 6.3 Practical waveforms for the original 4 tooth design at 1.0Nm torque, 1090rpm, 800μs advance (5 degrees), 168μs cut-off (1 degree) [7]
Figure 6.4 RM time stepping simulated torque for the original 4 tooth design at 1090rpm

Figure 6.5 Previously available TRW simulated torque and average torque waveforms for the original 4 tooth design at 1090 rpm [7]
Figure 6.6 RM time stepping simulated field and armature current waveforms for the original 4 tooth design at 2175rpm.

Figure 6.7 Previously available TRW simulated waveforms at 0.8 Nm torque, 2175 rpm, 600 μs advance (8 degrees), 168 μs cut-off (2 degrees), warm conditions [7].
Figure 6.8 Previously available TRW experimental waveforms for the original 4 tooth design at 0.8 Nm torque, 2175 rpm, 600 µs advance (8 degrees), 168 µs cut-off (2 degrees) [7]

Figure 6.9 RM time stepping simulated torque for the original 4 tooth design at 2175 rpm
The above results show a good comparison between the RM simulation, the previously simulated data and the experimental data. In terms of magnitude, the RM results are within 10% of the previously available simulated and experimental results. The clear difference being that the torque produced in the RM goes negative while the other simulation always has positive torque. This can be put down to a number of factors. The main one being that the TRW data cannot be guaranteed to be accurate and small changes in the switching angle can lead to considerable variation in torque production. The accuracy of the switching angles used in the RM to perform the time stepping analysis is also limited. These inaccuracies coupled together mean that it is difficult to replicate the torque curves accurately by switching at exactly the same times.

Another noticeable difference between the time stepping simulation results produced by the RM and TRW’s results is that the starting transients are different. This can be seen when comparing figures 6.4 and 6.5 and figures 6.9 and 6.10. This is due to the difference in the angle of rotation that the simulations start at, which is arbitrary. For
example, the torque waveform produced by the RM in figure 6.4 shows that the rotor has started closer to magnetic zero than that used to produce the TRW torque waveform in figure 6.5. The waveforms from both simulations steady after this initial transient and so the first half cycle or full cycle should be ignored and comparison made once the waveforms have stabilised.

In using the RM to produce the above time stepping results, it was established that the simulation was capable of producing reliable and comparable results to the previous work – both simulated and experimental. This meant that it could be used to predict what would be reasonably accurate waveforms for the 12 tooth rotor design and hence enable some meaningful comparison between this and the original 4 tooth design.

The RM software is of course fairly complicated to set up and also takes a long time to produce the results. The PSpice simulation developed earlier is simpler to use and less time consuming. Hence, preliminary modelling work could be carried out in PSpice, with the results being verified by the RM. In this way the 12 tooth design could be modelled extensively using the two different simulations and allow the accuracy of both to be tested against previous data available for the 4 tooth motor.

6.3 Using PSpice Simulation to Determine Winding Configuration for the 12 Tooth Design

In order for the 12 tooth design to produce comparable torque to the original 4 tooth design, the number of turns in the armature and field windings had to be changed. Initially, the 12 tooth design had been developed using the 21 field and 19 armature turns used in the original 4 tooth design. However, it can be seen from the previous results produced in the RM that the current in the 12 tooth windings is much lower than that for the 4 tooth design and hence the torque it produces is lower too. By reducing the number of turns in the windings, a greater current can be achieved due to lower resistance and inductance and therefore more torque will be produced.

The optimal number of turns to produce maximum torque can be found using trial and error by reducing the number of turns one at a time. However, doing this in the RM
would be a very lengthy process and so here the PSpice model has a significant advantage. Torque and copper loss results can be quickly obtained for various winding and speed combinations in the PSpice model for the 12 tooth design.

However, in doing this it is important to note another limitation of the PSpice model. This being that the PSpice model is a linear model, which does not take into account the effects of saturation. The RM model does account for saturation and so these effects will impact upon the torque that is produced in that model. Hence, it was quickly found that as the number of turns used was decreased in the PSpice model, the current continued to rise and higher and higher torques were output until only a very small number of turns were in use. At such low numbers of turns, the RM model results obviously did not agree with these PSpice results due to the saturation effects present.

In order to avoid unrealistic results from PSpice, it was decided to keep the current produced in the windings to around 100A at 1000rpm. This is around the same current produced by the previous data available for the original 4 tooth design at 1090rpm and so this was a good starting point. It was quickly found using PSpice that this combination of current and speed could be achieved by reducing the original turns by one on each winding, so the configuration was now 20 field turns and 18 armature turns.

The resistance and inductance values used in PSpice for the 12 tooth model were obtained using a previously available spreadsheet provided by TRW used to calculate winding parameters. This allowed parameters from the new 12 tooth design to be entered including the number of rotor teeth and the significant change of larger slot areas on the stator. This obviously changed the thickness of the wire used from the original 4 tooth design, and so this coupled with a slightly lower number of turns changed the resistance values of the windings quite considerably. Table 6.1 below shows the resistance and inductance values that were used for the original 4 tooth and 12 tooth designs for comparison.
<table>
<thead>
<tr>
<th>Design</th>
<th>No. of turns</th>
<th>Resistance (mΩ)</th>
<th>Inductance (μH)</th>
<th>No. of turns</th>
<th>Resistance (mΩ)</th>
<th>Inductance (μH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Original 4 tooth</td>
<td>21</td>
<td>32</td>
<td>79.1</td>
<td>19</td>
<td>40</td>
<td>67.4</td>
</tr>
<tr>
<td>12 tooth</td>
<td>20</td>
<td>17.9</td>
<td>62.52</td>
<td>18</td>
<td>29.9</td>
<td>49.31</td>
</tr>
</tbody>
</table>

Table 6.1 Resistance and inductance values used for the original 4 tooth and 12 tooth designs

The PSpice circuit for the 12 tooth motor with 20 field and 18 armature turns can be seen below in figure 6.11 and is set up to simulate at a speed of 1000rpm.
Figure 6.11
12 tooth rotor PSpice circuit with 20 field and 18 armature turns at 1000rpm
6.4 Verifying the PSpice Results Using the RM

With the number of turns chosen using PSpice, the circuit simulation was run at speeds of 1000, 1500, 3000 and 4500rpm. The results obtained were clearly reasonable in terms of current and torque shape and magnitude when compared to previous results for the 4 tooth motor. It was now necessary to compare these to results produced using the RM at the same speeds to prove that the two simulations concurred. This would mean that the PSpice simulation was performing as desired by producing initial quick results that were close to the more extensive RM simulation and thus allowing easy parameter manipulation.

The results produced by PSpice and the RM at the speeds mentioned above are displayed together on the following pages for easy comparison in figures 6.12 to 6.23.
Figure 6.12 Field and armature current waveforms produced in PSpice at 1000rpm for the 12 tooth rotor design

Figure 6.13 Field and armature current waveforms produced in the RM at 1000rpm for the 12 tooth rotor design
Figure 6.14 Comparison of torque waveforms produced in PSpice and the RM at 1000rpm for the 12 tooth rotor design

The above waveforms show a close comparison between the PSpice and RM time stepping simulation in terms of the magnitude of the currents. It could be considered from these plots that the PSpice simulation is switching slightly too soon due to the higher current peak just before switching when compared to the RM waveforms. The torque waveforms show that this is probably the case due to the negative torque produced by the RM time stepping simulation. If the RM simulation was switched earlier then it is more likely to be closer to the PSpice torque waveform, which has a higher peak and does not go negative. This illustrates that it is difficult to accurately find the correct switching point when comparing simulations. It is made harder still in the RM because it is so time consuming.
Figure 6.15 Field and armature current waveforms produced in PSpice at 1500rpm for the 12 tooth rotor design.

Figure 6.16 Field and armature current waveforms produced in the RM at 1500rpm for the 12 tooth rotor design.
Again, these waveforms show that the RM time stepping simulation should be switched a little earlier to replicate the PSpice results more accurately and avoid the negative torque. Apart from this though, it is evident that the two simulations produce results which are reasonably close in terms of magnitude and shape of the waveforms for both current and torque.

At these lower speeds it can be noted that the higher peak torque produced in PSpice could also be due to the fact that at low speed the saturation within the motor is greater. The RM models this saturation and so this is taken into account within its results, whereas the PSpice model is linear and does not account for this. So, when saturation is higher this will impact on the comparison between results more significantly.

**Figure 6.17 Comparison of torque waveforms produced in PSpice and the RM at 1500rpm for the 12 tooth rotor design**
Figure 6.18 Field and armature current waveforms produced in PSpice at 3000rpm for the 12 tooth rotor design.

Figure 6.19 Field and armature current waveforms produced in the RM at 3000rpm for the 12 tooth rotor design.
These results at 3000rpm show that this time, the RM time stepping simulation is no longer producing negative torque. This could be down to the switching angle used here being closer to ideal for this design. The PSpice torque waveform is now going slightly negative. It can be noted at this point that the results displayed from the three speeds so far show that the PSpice torque waveforms drop to roughly zero each time. The cause of this is mainly down to the switching point and how close this is to the ideal position. The current waveforms here are less alike in terms of shape, but are very similar in magnitude, peaking at around 50A.
Figure 6.21 Field and armature current waveforms produced in PSpice at 4500rpm for the 12 tooth rotor design.

Figure 6.22 Field and armature current waveforms produced in the RM at 4500rpm for the 12 tooth rotor design.
The results produced at 4500rpm show again that the RM time stepping simulation is producing all positive torque indicating a switching point closer to ideal. Its peak magnitude is close to the PSpice simulation, within about 10%. The PSpice torque is once more going negative, which is most likely to be down to the switching angle not being at the ideal position as discussed. It appears that as speed is increased the two simulations start to produce current waveforms that are less alike in terms of shape (the field current remains more constant in the RM simulation). They are still comparable in terms of magnitude, being within 10% of each other and this time peaking just under 40A. This difference in shape does not seem to impact too much on the torque waveforms produced.

Although there are clearly some differences between the PSpice and RM simulations, the results obtained demonstrate that on the whole the PSpice simulation is an accurate predictive tool. Its main drawbacks are the need to use a constant self inductance value and the inability to model saturation due to being a linear model. Despite this though, it has proved to be a much simpler and quicker way of obtaining results when compared to the RM. In a matter of minutes the winding configuration
can be changed (by altering inductor and resistor values) and the speed of rotation can also be changed giving great flexibility. Hence a large number of parameter combinations can be investigated in a time that would take the RM simulation a good number of hours just to run a couple of configurations. This time-saving advantage outweighs any disadvantage in terms of accuracy as PSpice can be used to get a rough idea of how one optimised configuration will work and then this can be modelled more accurately in the RM.

6.5 Comparing the Original 4 Tooth Design with the New 12 Tooth Design Using the RM

In the previous chapter, static finite element analysis was used to compare proposed designs, leading to the choice of a new 12 tooth design. The static FE indicted that the 12 tooth design produced more torque, but with considerably greater iron losses. The RM can now be used to perform a time stepping analysis and further results then obtained using PSpice in the process. These results can be compared to those produced for the original 4 tooth design to build upon the static work. The dynamic effects taken into account by the RM should allow a more accurate picture of how the two designs compare in terms of torque production.

The following figures show torque waveforms produced in the RM for both the original 4 tooth and new 12 tooth designs.
Figure 6.24 Comparison of torque produced in RM for original 4 tooth and new 12 tooth designs at 1000rpm

Figure 6.25 Comparison of torque produced in RM for original 4 tooth and new 12 tooth designs at 1500rpm
Figure 6.26 Comparison of torque produced in RM for original 4 tooth and new 12 tooth designs at 3000rpm

Figure 6.27 Comparison of torque produced in RM for original 4 tooth and new 12 tooth designs at 4500rpm
The torque curves above show a direct comparison between the original 4 tooth motor and the new 12 tooth design at the 4 selected speeds. It can be seen that at the lower speeds of 1000 and 1500rpm, the 12 tooth torque output is a lot closer to the 4 tooth torque than at the higher speeds of 3000 and 4500rpm. This is also evident from table 6.2 below which shows the average torque produced by the two designs for the 4 speeds.

<table>
<thead>
<tr>
<th>Motor design</th>
<th>Speed (rpm)</th>
<th>Average torque (Nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>original 4 tooth</td>
<td>1000</td>
<td>1.26</td>
</tr>
<tr>
<td></td>
<td>1500</td>
<td>1.15</td>
</tr>
<tr>
<td></td>
<td>3000</td>
<td>0.69</td>
</tr>
<tr>
<td></td>
<td>4500</td>
<td>0.41</td>
</tr>
<tr>
<td>12 tooth</td>
<td>1000</td>
<td>0.90</td>
</tr>
<tr>
<td></td>
<td>1500</td>
<td>0.76</td>
</tr>
<tr>
<td></td>
<td>3000</td>
<td>0.37</td>
</tr>
<tr>
<td></td>
<td>4500</td>
<td>0.20</td>
</tr>
</tbody>
</table>

Table 6.2 Comparison of average torque values for the two motor designs at the selected speeds.

From table 6.2, it can be seen that the torque produced by the 12 tooth design is around 70% of that produced by the 4 tooth design at 1000rpm. At the highest speed of 4500rpm though, the 12 tooth design is producing around half the torque of the 4 tooth. Hence, it is clear from these results that the original 4 tooth motor is far superior in terms of torque production when compared to the new 12 tooth design. However, it may be that, despite the 12 tooth design also having much higher iron losses, it does have an advantage in terms of efficiency due to lower copper losses. The winding configuration for the 12 tooth motor has been determined using PSpice and so now the copper losses for the two designs can be calculated and compared. It would be expected that the 12 tooth design, with its larger slot areas and fewer turns, has lower copper loss. This may outweigh its higher iron loss.

Copper loss can be calculated from the equations below;

\[ P_{CU(field)} = I_f^2 R_f \]  \hspace{1cm} (6.1)

\[ P_{CU/arm} = I_A^2 R_A \]  \hspace{1cm} (6.2)
Output power is calculated as follows;

\[
P_{\text{out}}(W) = \frac{\text{average}_\text{torque}(Nm) \times 2\pi \times \text{speed}(rpm)}{60}
\]  \hspace{1cm} (6.4)

Efficiency can then be calculated:

\[
\text{Efficiency} = 100 \times \frac{P_{\text{out}}}{P_{\text{out}} + P_{\text{iron}} + P_{\text{CU}}}
\]  \hspace{1cm} (6.5)

Table 6.3 compares power output of the 4 and 12 tooth designs along with their copper and iron losses and gives the efficiency for the 4 speeds investigated.

<table>
<thead>
<tr>
<th>Motor Design</th>
<th>Speed (rpm)</th>
<th>Power Out (W)</th>
<th>Iron Loss (W)</th>
<th>Copper Loss (W)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Original 4 tooth</td>
<td>1000</td>
<td>132.07</td>
<td>29.63</td>
<td>794.19</td>
<td>13.82</td>
</tr>
<tr>
<td></td>
<td>1500</td>
<td>180.33</td>
<td>65.58</td>
<td>629.56</td>
<td>20.60</td>
</tr>
<tr>
<td></td>
<td>3000</td>
<td>216.20</td>
<td>258.00</td>
<td>281.05</td>
<td>28.63</td>
</tr>
<tr>
<td></td>
<td>4500</td>
<td>193.18</td>
<td>577.26</td>
<td>142.04</td>
<td>21.17</td>
</tr>
<tr>
<td>12 tooth</td>
<td>1000</td>
<td>97.34</td>
<td>63.32</td>
<td>389.04</td>
<td>17.71</td>
</tr>
<tr>
<td></td>
<td>1500</td>
<td>122.57</td>
<td>140.83</td>
<td>282.53</td>
<td>22.45</td>
</tr>
<tr>
<td></td>
<td>3000</td>
<td>116.15</td>
<td>556.72</td>
<td>90.91</td>
<td>15.21</td>
</tr>
<tr>
<td></td>
<td>4500</td>
<td>95.37</td>
<td>1247.66</td>
<td>51.32</td>
<td>6.84</td>
</tr>
</tbody>
</table>

Table 6.3 Comparison between original 4 tooth motor design and new 12 tooth design, showing output power, losses and efficiency at the 4 speeds.

Table 6.3 shows that the output power produced by the 12 tooth design is much lower than for the original 4 tooth design at the same speeds. Again, this output power is around 70% of the 4 tooth at the lower speeds and around 50% at the higher speeds as the torque figures suggested.
As discussed in chapter 5, the iron losses in the 12 tooth design are much greater than those in the 4 tooth design. The 4 tooth iron loss is around 46-48% of the 12 tooth iron loss at all the speeds investigated.

However, as expected, the copper loss is considerably lower in the 12 tooth design. At 1000rpm, the 12 tooth copper loss is around half that of the 4 tooth design. As the speed is increased, this margin is increased so that at the two higher speeds of 3000 and 4500rpm the 12 tooth copper loss is around 32% and 36% of the 4 tooth copper loss respectively.

This is countered by the very large iron loss figures produced by the 12 tooth design, meaning that an increase in efficiency is not observed. It is noticeable though that at the lower speeds, the 12 tooth design is more efficient, with its lower copper loss obviously outweighing its higher iron loss.
6.6 Conclusions from the RM Time Stepping Results

The torque waveforms produced by the RM clearly show that the original 4 tooth design produces greater torque than the new 12 tooth design. This is in contrast to what was indicated by the static finite element analysis work described in previous chapters when the designs were compared at constant copper loss and current density. With regard to improving the performance of the motor this is obviously disappointing.

It can be concluded from the RM results that the dynamic factors (that the static analysis does not account for) play an important part in the production of torque. The static results indicated that the 12 tooth design would produce more torque than the original 4 tooth motor (when compared at constant current density and copper loss). However, the more complex and dynamically accurate time stepping analysis has shown this not to be the case.

The dynamic effects present in the actual motor as the rotor rotates can be split into first and second order effects. The first order effects include inductance, back emf, time constants and rise times. The second order effects include eddy currents and hysteresis skin effect. The RM time stepping simulation can model the first order effects, but the second order effects are harder to model accurately. In this case therefore, the difference between the time stepping and static models is down to these first order effects. The static model obviously does not account for variations in the inductance of the windings, the back emf present in the armature and of course time constants and rise times. It has been noted before that one of the limitations of the PSpice model is that the self inductance values are constant, when this is not the case in the actual motor.

When comparing output and losses from the two designs it became clear that the 12 tooth design may have considerably higher iron losses (up to double those of the 4 tooth), but the 4 tooth has much higher copper loss (up to 3 times that of the 12 tooth). This meant that the 12 tooth design was more efficient at lower speed (although it produces less power). At higher speed the iron losses of the 12 tooth design became very high so that it was no longer outweighed by the lower copper loss and the
efficiency dropped very low (less than 7% at 4500rpm). These figures therefore indicate that the 12 tooth design does have some advantage over the 4 tooth design at lower speeds.

In order to reduce the iron losses in the 12 tooth design, the lamination thickness could be reduced to reduce the large eddy current loss. This would obviously be more costly to produce thinner laminations, but would allow the benefits of the 12 tooth design, mentioned above, to be exploited. This may prove to be an overall improvement on the original design. However, in order to determine if the 12 tooth design has any significant benefit over the original, it would be necessary to build a prototype and run experimental tests to verify the simulation results. Only then can a true comparison be made between the designs and a conclusion drawn on which is the best.
Chapter 7
Conclusions and Further Work

7.1 Introduction

This chapter details the objectives at each stage of the research and the conclusions reached in each chapter before pulling the investigation together to reach an overall conclusion.

7.2 Chapter 1 – Introduction

This chapter introduced the thesis and described the layout in terms of the general content of each chapter. It defined the aims and objectives of the work and the main conclusions that could be drawn from it.

7.3 Chapter 2 – Background

In this chapter, the research conducted into why this work was necessary was presented. This looked at the use of electronic driver aids in the automotive industry and how brushless motors in particular are becoming more important as vehicles become more reliant on their electronic systems. Devices such as anti-lock braking systems and traction control were singled out as areas where brushless motors may play an important part in the future. The chapter then went on to introduce the flux switching motor, a new type of brushless motor. A conclusion was reached that many features of this motor such as its robust nature, reliability and low cost would be suitable to automotive applications where these attributes are crucial. Hence, research was necessary to produce a method of accurately modelling the motor and allow an optimal design to be achieved for use in this field.

7.4 Chapter 3 - Developing a Circuit Simulation of the Flux Switching Motor in PSpice

This chapter looked at the initial modelling process of accurately representing the motor and its power electronics using a circuit simulation technique. A simple motor
model was produced using PSpice circuit simulation. This model was then developed to take into account the effects of mutual coupling between the windings. The concept behind this was to produce a relatively simple model where the configuration of the motor windings (i.e. number of turns) and the speed of operation could be varied easily and quickly. This would then allow many sets of parameter variations in a short space of time in order to arrive at an optimal winding configuration. The PSpice model could be used to obtain quick initial results to give a rough idea of how a motor design would perform before using the more complex and time consuming, although more accurate, time stepping analysis detailed in chapter 6.

A more complex PSpice model was investigated later in the chapter using harmonics to model the mutual inductance effects more accurately. However, it was found that the results from the simple model were a close enough representation when compared to simulation and experimental data that was already available. Hence, a more complicated model was not necessary.

The main limitation of the model was the assumption that the self inductances of the windings remain constant. This is not the case in practise, but was done in the PSpice model to keep it as simple as possible without using complex look-up tables. It was found that this limitation was very small and affected the torque output from the model by only 0.7% maximum.

The simple PSpice model managed to produce results for the original 4 tooth design that were reasonably close to previous simulation results produced by TRW and also experimental data already available. The PSpice current waveforms were within 10% of the TRW results at both the speeds investigated. At the lower speed of 1090rpm, the average torque produced by the simple PSpice model was within 10%. This was not as good at the higher speed of 2175rpm, where it only produced around two thirds of the TRW simulation. This could be put down to inaccuracies in switching patterns between the two simulations, which would be more influential at higher speed.

The model was a useful tool in the development of a new motor design by allowing some quick and easy initial predictions of performance.
7.5 Chapter 4 - Finite Element Analysis of the Flux Switching Motor in Opera

In this chapter the design process of altering the motor lamination profile was detailed. This involved varying many parameters affecting the geometry of both the stator and the rotor in order to produce greater torque output. It became clear as parameters were investigated that it was those nearest the airgap on both stator and rotor that affected torque production most significantly. The amount of protrusion on the leading edge of the rotor also greatly affected the amount of crossover torque produced. This crossover torque was defined as the amount of torque produced as the current in the armature winding was switched and was important to allow the motor to self start. If the torque dropped to zero at the switching point i.e. no crossover, then there may be positions that the rotor could finish in where starting next time would be impossible. It was discovered that there was a trade off between the peak torque and the crossover torque, where maximising the peak torque decreased the crossover and vice versa.

The introduction of a 'step' on the leading edge of the rotor aimed to improve the crossover torque produced without too much detriment to the peak torque. This design proved to be the most promising and produced greater crossover, average (0.3N more) and peak (4N higher) torque than the original design at constant current density. This 'optimal step' design was taken forward for further investigation in the next chapter.

7.6 Chapter 5 - Developing the 12 Tooth Rotor Model

This chapter took the investigation a stage further by significantly changing the topology of the motor. The number of teeth on the rotor was increased to 12 from 4 in the idea that this would increase the frequency of operation, allowing a higher back emf and hence more power. It was found that there was a trade off in that the smaller teeth on both stator and rotor meant higher saturation and more flux leakage. This was shown by much lower coupling between the field and armature windings than in the original 4 tooth design. The coupling in the 12 tooth designs being around half that in the 4 tooth designs (35% compared to 70%).

A number of 12 tooth designs were investigated with the three most promising being singled out for more detailed comparison with the original 4 tooth design and the
'optimal step' design from chapter 4. One of these 12 tooth designs, design 3, was found to exhibit the most potential by producing a high peak torque per metre (41.8N). This was 13.5N and 10.6N greater than the original and optimal 4 tooth designs respectively when compared at constant current density. It also produced a high enough crossover torque to enable self starting, despite this being lower than the 4 tooth designs (around 5 N.degrees lower than the original and 19.4 N.degrees lower than the 4 tooth optimal design). The higher frequency of operation resulted in significantly more torque ripple, but a higher average torque was achieved than for the 4 tooth designs. This average was 7.8N higher than the original 4 tooth design and 7.5N higher than the optimal 4 tooth design when compared at constant current density.

When compared at constant average torque per metre, it was found that more current was required to achieve an average of 19.2N for the 12 tooth designs. The original required a current input of 12.5A while the 12 tooth design 3 required 13.85A to achieve this average torque output. This indicates that the 12 tooth design requires higher power to generate the same torque and is hence less efficient.

The proposed designs were then compared in more detail by taking into account copper and iron losses. By simulating the designs using finite element analysis at the same copper losses (i.e. $J^2A$ being equal for all designs) it was found that 12 tooth design 3 produced the highest torque output both in terms of peak and average torque. Its peak torque per metre was around 4N higher than the original 4 tooth design, while its average torque per metre was 5.5N greater. However, it was also found that the 12 tooth designs, as predicted, suffered from significantly greater iron losses due to narrower teeth. These losses in the 12 tooth designs were more than double those of the 4 tooth designs. The iron loss predictions indicated that these losses would outstrip the benefit of more torque for 12 tooth design 3.

A suggestion for further work was made at the end of this chapter by presenting a 12 tooth design concept with wider stator teeth and back irons in an attempt to reduce iron loss. Some initial investigation showed that this design may have potential. The torque produced was still lower than the original design in terms of both peak and average torque and it also suffered from low crossover torque. Any future work based
around this type of design would have to determine if the torque output can be improved and how significant the reduction in iron loss is when compared to the 12 tooth designs proposed in this work.

7.7 Chapter 6 - Using the Rotating Machine Program to Perform Time Stepping Finite Element Analysis of 4 Tooth and 12 Tooth Motors

This chapter endeavoured to take the modelling process another stage further by introducing a more complex method, which could potentially produce more accurate predictions of motor performance. The method used a time stepping approach to take account for dynamic effects present in the operation of the motor. The static finite element analysis used in the previous chapters could obviously not model these effects.

Firstly, the RM was used to model the original 4 tooth motor and ultimately produced results that were comparable (within 10%) to previous TRW simulations and experimental data. Hence the RM was capable of producing an accurate prediction of motor performance in terms of current and torque and could therefore be used to compare the proposed designs. The disadvantage of this simulation was that it was very time consuming and rather complex to set up when compared to PSpice and the static finite element analysis.

The next part of the investigation made use of the PSpice simulation developed in chapter 3 to optimise the winding configuration of the chosen 12 tooth design (design 3) and provide some initial results. These results were then replicated using the rotating machines solver (RM). This proved that the PSpice model had a relatively good accuracy, certainly at higher speeds, with the results being within 10% of the RM results. It was clear at the lower speeds that the linearity of the PSpice model affected its accuracy as the RM could model saturation effects which PSpice could not. This was also coupled with the difficulty in replicating the exact switching angles between the models. Although a limitation here, the PSpice simulation could be concluded a good enough representation to be useful as a development tool.
The RM was then used to compare the original 4 tooth design to the new 12 tooth design in terms of torque production at a number of different speeds. The results from this were surprising in that they contradicted the static finite element results and clearly showed that the original 4 tooth design produced greater peak torque than the 12 tooth design. This was significant at higher speeds, where the 12 tooth design produced around half the average torque of the original 4 tooth design. At the lower speeds the 12 tooth design was producing more like 70% of the torque of the original design.

Therefore it can be concluded that the dynamic effects present play a significant part in the production of torque. A static finite element analysis may be a good starting point, but ultimately the more complex and lengthy process of using time stepping methods are required to model the necessary dynamic effects.

7.8 General Conclusion to the Work

The objectives of this research aimed to look at different ways of modelling the flux switching motor using circuit simulation, static and time stepping finite element analysis. The ultimate objective was to improve upon the design of the flux switching motor using these simulation tools to model it accurately and arrive at an optimised design.

It can be concluded that the first of these objectives has been met. A new PSpice circuit simulation of the motor and its power electronics has successfully been developed. This model is simple to use and can produce very quick results enabling rapid investigation into configuration changes. The results are accurate within reason when compared to TRW data taking into account the possible inaccuracy of switching patterns. At lower speed the PSpice model produced results within 10% of the TRW results. Although this was less accurate at the higher speed (66% of the TRW average torque figure) it fulfils its main use as an initial predictive tool.

The rotating machines time stepping finite element analysis has been developed for use with the flux switching motor using external circuits to model the winding parameters. This simulation model has allowed a more accurate dynamic
representation of the motor, which takes into account factors that the static analysis is incapable of. Although a time consuming and complex model to use, it has proved to provide some accurate results when compared to previous simulation and experimental data (within 10%). When used in conjunction with the PSpice model, they combine to make a useful method of developing potential new motor designs.

The research indicates that the 12 tooth rotor model has improved upon the original design in some areas, but is not as good in others. The proposed 12 tooth design performed well using static finite element analysis and appeared to produce more torque than the original 4 tooth design even when copper loss was taken into account (an increase of around 15%). However, it was always clear that the iron losses were significantly higher (more than double) in all the 12 tooth designs investigated due to the narrower teeth. The impact of this indicated that the higher torque produced would be out-weighed by these greater losses i.e. a 15% increase in torque at constant copper loss, but a 100% increase in iron loss.

In terms of efficiency, the much lower copper losses in the 12 tooth design mean that it was more efficient than the original design at lower speeds (around 4% more efficient at 1000rpm and around 2% at 1500rpm). At higher speeds the iron loss became so great that it was the dominating factor and hence the 12 tooth design was less efficient here than the original (around 13% lower efficiency at 3000rpm and around 14% at 4500rpm).

When the RM was used to compare the designs it also became clear that, when accounting for dynamic affects, the 12 tooth design actually produced less torque than the original 4 tooth design (only half the torque at higher speeds and 70% at lower speeds). Hence, it proved that the flux in the smaller rotor teeth decreased, due to saturation and flux leakage, more greatly than the frequency increased. The higher frequency of operation would in theory create a higher back emf and hence more power, but this was countered by the need for narrower teeth and thus more saturation and leakage. The designs proposed during this work could not overcome this problem to produce a net gain of power. However, it was shown that the overall losses in the 12 tooth design were lower than in the original 4 tooth design at the lower speeds.
under 1500rpm. Hence the losses in the 12 tooth design are smaller if the current is less.

**7.9 Author’s Contribution to Knowledge**

This work has made use of a novel approach to motor design by combining different methods already in use. Static finite element analysis was used to provide data, which was then fed into a PSpice circuit analysis. An optimum motor design was then produced and the PSpice results were then compared to time stepping FEA, which also allowed for more accurate investigation. The PSpice simulation served as a quick and easy to use initial model, providing a good indication of how a design would perform. This could then be verified using the more complex and time consuming time stepping FEA.

The work also involved the investigation of an 8/12 flux switching motor. The flux switching motor has not been investigated previously in this configuration of 8 stator teeth and 12 rotor teeth. The advantages and disadvantages of this type of configuration have been highlighted and can now be considered for future motor topologies.

**7.10 Further Work**

The tools developed throughout this research can be used to aid further work in developing flux switching motor designs. The PSpice model could potentially be developed further to include iron and copper loss calculation. More work could be carried out in looking at further geometric lamination designs for both 4 and 12 rotor teeth. These could include more radical cut-out rotor and stator shapes and possibly even the use of square or round holes in the teeth. A design was proposed in chapter 5 with the aim of reducing the high iron losses found in the 12 tooth designs by increasing the thickness of the stator teeth and back irons. This type of design could be investigated further in the future in order to determine if its torque output can be improved. It would also be necessary to establish if this can be done while the benefit of lower iron loss remains and if this benefit outweighs the higher copper loss due to smaller slot areas.
References


Appendix

The tables on the following pages detail the parameters investigated in the stator and rotor lamination profiles carried out in Opera as documented in chapter 4. The name of each parameter along with a brief description, the range of values over which they were varied, and the effect they had on torque production are included.
<table>
<thead>
<tr>
<th>Name</th>
<th>Description</th>
<th>Range of Values</th>
<th>Effect on Torque</th>
</tr>
</thead>
<tbody>
<tr>
<td>ARM_SLOTS_PA01</td>
<td>Point on top of armature slot pole shoes</td>
<td>0.04 – 0.2</td>
<td>Minimal impact on torque, with higher values producing slightly higher average.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA02</td>
<td>Point on top of armature slot pole shoes</td>
<td>0.1 – 0.3</td>
<td>Minimal impact on torque, with higher values producing slightly higher average.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA03</td>
<td>Point on top of armature slot pole shoes near stator pole</td>
<td>0.2 – 0.4</td>
<td>Higher values produced slight increase in average torque. Minimal impact on</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>crossover torque.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA04</td>
<td>Point midway between airgap and back iron</td>
<td>0.3 – 0.5</td>
<td>Higher values produced slight increase in average torque. Minimal impact on</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>crossover torque.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA05</td>
<td>Point midway between airgap and back iron</td>
<td>0.3 – 0.5</td>
<td>Higher values produced slight increase in average torque. Minimal impact on</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>crossover torque.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA06</td>
<td>Point on stator tooth near back iron</td>
<td>0.3 – 0.5</td>
<td>Minimal impact on torque.</td>
</tr>
<tr>
<td>ARM_SLOTS_PA07</td>
<td>Point on stator tooth near back iron</td>
<td>0.2 – 0.4</td>
<td>Minimal impact on torque.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA01</td>
<td>Point on top of field slot pole shoes</td>
<td>0.04 – 0.2</td>
<td>Minimal impact on torque, with higher values producing slightly higher average.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA02</td>
<td>Point on top of field slot pole shoes</td>
<td>0.1 – 0.3</td>
<td>Minimal impact on torque, with higher values producing slightly higher average.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA03</td>
<td>Point on top of field slot pole shoes near stator pole</td>
<td>0.2 – 0.4</td>
<td>Higher values produced slight increase in average torque. Very little impact on</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>crossover.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA04</td>
<td>Point midway between airgap and back iron</td>
<td>0.2 – 0.46</td>
<td>Higher values increased average torque. Very little impact on crossover.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA05</td>
<td>Point midway between airgap and back iron</td>
<td>0.2 – 0.4</td>
<td>Higher values produced slight increase in average torque. Minimal impact on</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA06</td>
<td>Point on stator tooth near back iron</td>
<td>0.2 – 0.4</td>
<td>Minimal impact on torque.</td>
</tr>
<tr>
<td>FIELD_SLOTS_PA07</td>
<td>Point on stator tooth near back iron</td>
<td>0.2 – 0.4</td>
<td>Minimal impact on torque.</td>
</tr>
<tr>
<td>GEOMETRY_PA02 &amp; PA03</td>
<td>Width of stator teeth immediately at airgap</td>
<td>0 – 1.0</td>
<td>Lower values produce higher torque. High values produce very little crossover.</td>
</tr>
<tr>
<td>DIMENSIONS_PA02</td>
<td>Length of airgap</td>
<td>0.3 – 1.4</td>
<td>Lower values increase torque significantly, in terms of average and crossover.</td>
</tr>
</tbody>
</table>
## Parameters Investigated in the Rotor

<table>
<thead>
<tr>
<th>Name</th>
<th>Description</th>
<th>Range of Values</th>
<th>Effect on Torque</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rotor arc/pitch at air gap</td>
<td></td>
<td>0.2 - 0.6</td>
<td>Middle values produced highest average torque and greatest crossover.</td>
</tr>
<tr>
<td>Fraction of rotor radius to leading edge between midway and airgap</td>
<td></td>
<td>0.8 - 1.0</td>
<td>Lower values produce greatest average torque. Middle values produce slightly greater crossover.</td>
</tr>
<tr>
<td>Profile of leading edge just back from airgap</td>
<td></td>
<td>0.66 - 0.9</td>
<td>Minimal effect on both average and crossover torque.</td>
</tr>
<tr>
<td>Profile of leading edge midway between rotor base and airgap</td>
<td></td>
<td>0.76 - 0.96</td>
<td>Higher values produce slightly higher average torque. No significant impact on crossover.</td>
</tr>
<tr>
<td>Leading edge near rotor base</td>
<td></td>
<td>0.1 - 0.6</td>
<td>No significant impact on average or crossover torque.</td>
</tr>
<tr>
<td>Profile of leading edge near airgap</td>
<td></td>
<td>0.2 - 0.4</td>
<td>Significant effect on torque. Middle values produced highest average torque, while highest value produced greatest crossover.</td>
</tr>
<tr>
<td>Profile of leading edge near rotor base</td>
<td></td>
<td>0 - 0.5</td>
<td>Middle values produced higher average torque. Low values produce slightly higher crossover.</td>
</tr>
<tr>
<td>Diameter of rotor base</td>
<td></td>
<td>0.4 - 0.8</td>
<td>Middle values produced slightly higher average torque. Lower value increases crossover by minimal amount.</td>
</tr>
<tr>
<td>Leading edge at base of rotor</td>
<td></td>
<td>0.1 - 0.6</td>
<td>No significant impact on average or crossover torque.</td>
</tr>
<tr>
<td>Rotor arc/pitch at air gap</td>
<td></td>
<td>0.2 - 0.6</td>
<td>Middle values produced highest average torque and greatest crossover.</td>
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<td>Fraction of rotor radius to leading edge between midway and airgap</td>
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<td>0.66 - 0.9</td>
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<tr>
<td>Profile of leading edge midway between rotor base and airgap</td>
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<td>Higher values produce slightly higher average torque. No significant impact on crossover.</td>
</tr>
<tr>
<td>Profile of trailing edge just back from airgap</td>
<td></td>
<td>0.1 - 0.3</td>
<td>Lower values produced slightly higher average torque. Little impact on crossover.</td>
</tr>
<tr>
<td>Profile of trailing edge midway between rotor base and airgap</td>
<td></td>
<td>0.1 - 0.3</td>
<td>Virtually no effect on average torque or crossover produced.</td>
</tr>
<tr>
<td>Profile of trailing edge near base of rotor</td>
<td></td>
<td>0.1 - 0.6</td>
<td>No significant impact on average or crossover torque.</td>
</tr>
<tr>
<td>Profile at base of rotor</td>
<td></td>
<td>0.1 - 0.6</td>
<td>No significant impact on average or crossover torque.</td>
</tr>
<tr>
<td>Rotor outer diameter</td>
<td></td>
<td>36 - 46</td>
<td>Higher value produced greatest average torque. Minimal impact on crossover.</td>
</tr>
</tbody>
</table>