

Chapter 3 – Machine design

This chapter is divided into three sections. In the first section the detailed design plan for the prototype is given. The second section contains the stator design as well as the verification of the design. The last section contains the rotor design. The rotor design consists of the individual PMSM, IM and the combined LS PMSM rotors, each in its own subsection.

3.1 General information

In this section the design specifications and method for prototype LS PMSM will be discussed. The design process is indicated with the aid of a flow diagram. The various sub-design sections will also be listed along with a flow diagram

3.1.1 General specifications

The general specification of a motor is the information obtained from the motor manufacturers or from the datasheet. Table 3.1 lists these specifications.

Table 3.1: General specifications

Machine type	LS PMSM
Topology	Cylindrical shape radial flux
Rated power	7.5 kW
Rated line voltage	Three phase 525 V @ 50 Hz
*Rated current	10 A
Preferred line connection	Star
Number of poles	4
Duty cycle	S1 (continuous operation)
*Efficiency	95%
*Full load torque	± 48 N.m
*Power factor	>0.95

** Values are estimated in accordance with same power rating induction motor. Values will be determined in later stages of design and testing*

3.1.2 Design process

Figure 3.1 below is a diagram of the design process. The first step is to design the stator for both the PMSM and IM parts of the LS PMSM. After this is done the focus moves to the PMSM's rotor design, dynamic modelling and optimisation (indicated by the dashed arrow). The PMSM's rotor design must be done in such a way that it will be possible to accommodate a cage inside the rotor. After the PMSM's design is finalised, the IM- and combined rotor will be designed.

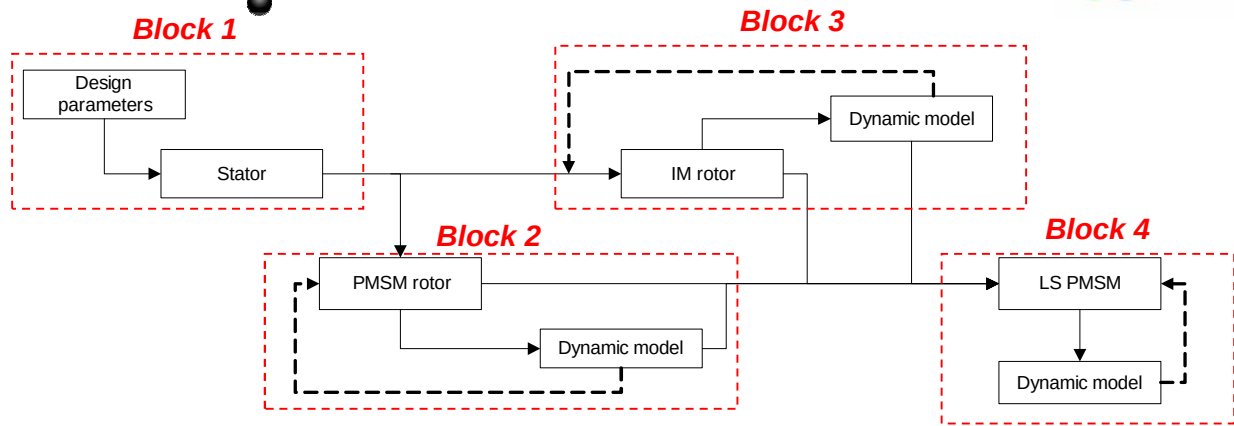


Figure 3.1: Extended design process

Information and results gained from both the PMSM and IM rotor design will then be used in the first iteration design of the LS PMSM prototype rotor. On completion of this design, a dynamic model needs to be compiled. The dynamic model will aid in adjusting the design until the desired results are met. After the LS PMSM's rotor and stator design is complete, the prototype design will be manufactured and tested.

The first step in Block 1 is to determine the design parameters; this sets the boundaries for the design, helps to clarify what needs to be done and establishes design focus points. This step will be discussed in depth in Section 3.1.3. In the second step of Block 1 a common stator needs to be designed. The term common refers to the usability of the stator. This means that both the PMSM rotor and IM rotor must be able to operate in this stator design; minor adjustments to the design may however be made. Figure 3.2 shows the generic stator design process, which was adapted from [1]. The original diagram in [1] is for the design of a general purpose IM and only focused on the design of the stator by using analytical equations with no simulations. The diagram below combines analytical calculations with simulations to produce the best-suited stator for the LS PMSM prototype.

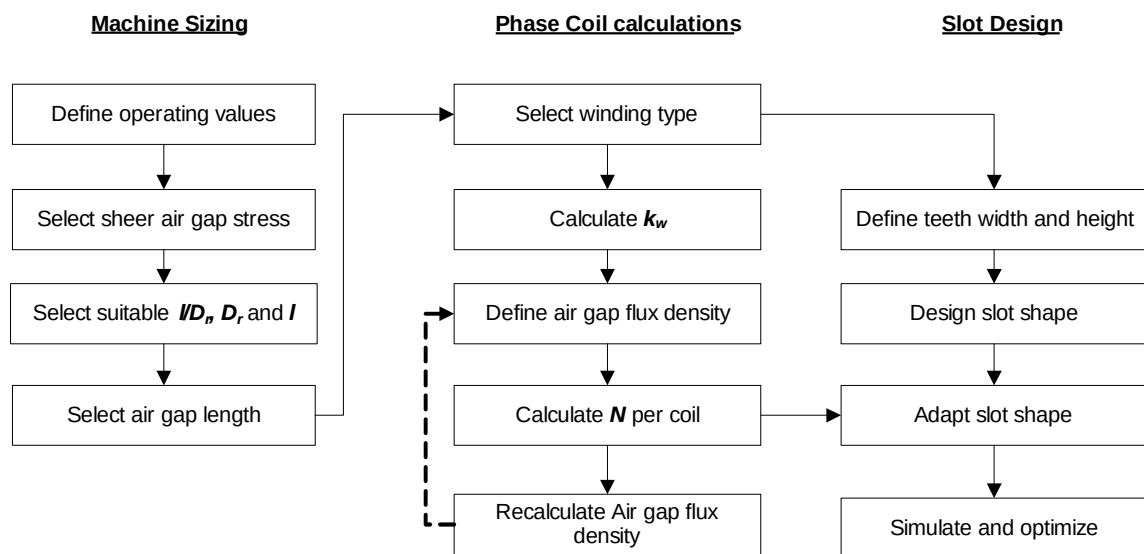


Figure 3.2: Common stator design [1, 3]

As seen in Figure 3.2, the stator design process is divided into three phases: machine sizing, phase coil calculations and slot designs. The in-depth design process and calculations for each phase will be presented in Section 3.3.

After the stator design is completed (Block 1), the next step is to design the individual rotor components as in Block 2 and Block 3. Both the blocks consist of the design and modelling of each component. The modelling will be done using the stator designed in Block 1. The design methodology and process for both the IM and PMSM were compiled of [1-3] and are indicated in Figure 3.3 and Figure 3.4. The final design will then be analysed and verified against the simulation results. Results gained from Block 2 and 3 will also be compared to the LS PMSM model.

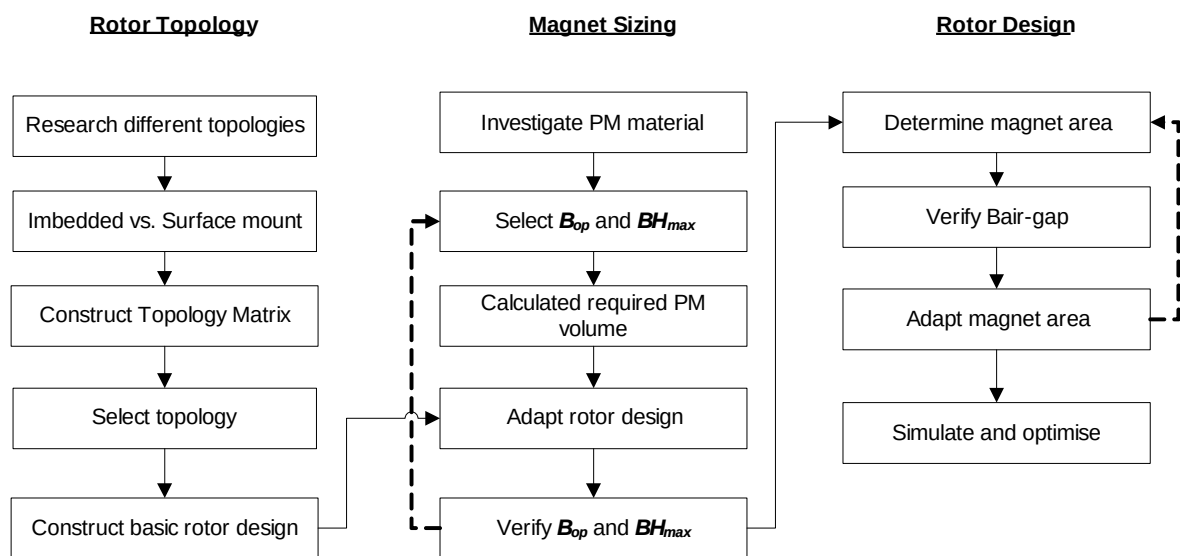


Figure 3.3: PMSM design process [1-3]

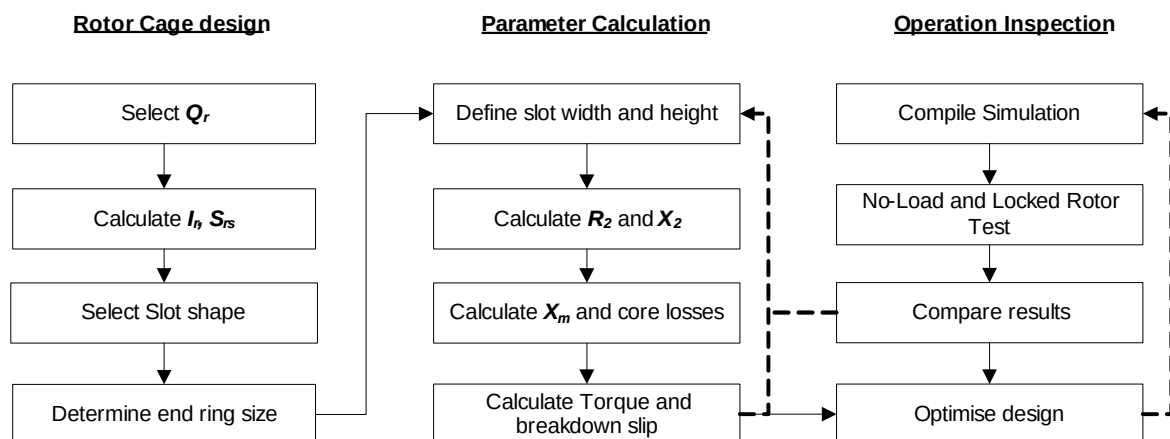


Figure 3.4: IM design process [1-3]

3.1.3 Determine design parameter

There are two types of applicable parameters that need to be selected or defined before starting machine design, namely fixed and free parameters. Fixed parameters are determined by design specifications, free

parameters can be selected by the designer [1]. In this section both of the parameters are discussed and listed; if applicable at this time, the rest of these parameters will be selected. The list below contains the base machine design parameters [1, 3].

- Electrical parameters (Power, Voltage, Current, Frequency)
- Number of pole pairs
- Machine duty cycle
- Stator stack outer diameter
- Stator stack inner diameter
- Air gap length
- Air gap diameter.

Each of the parameters must be evaluated to determine if the parameter is a free or a fixed parameter. If there is no clear indication to what the parameter must be, the parameter can be seen as a free parameter and can be determined by the designer. As each machine design is unique, the designer must determine what parameters are free or fixed for every machine design. Table 3.2 contains the parameters and if needed the reason for the classification.

Table 3.2: Classification of base design parameters

Parameter	Free/Fixed	Comment
Electrical parameters	Fixed	Specified in Chapter 1
Number of pole pairs	Fixed	Specified in Chapter 1
Machine duty cycle	Fixed	S1- Continuous running motor with fixed load.
Stator stack outer diameter	Free	Is partially fixed to standard frame sizes
Stator stack inner diameter	Free	
Air gap length	Free	
Air gap diameter	Free	

Once the base parameters classification has been selected, the free parameters need to be determined. The free parameters should be specified before commencing with the design. However, if needed during the design, a decision can be made to change a fixed-free parameter to aid the design.

3.2 Motor sizing

In this section methods of determining the size of the motor will be handled. The size of the motor is directly linked to the frame size for the motor. The sizing of the motor will aid in fixing the final free parameters.

3.2.1 Frame possibilities

Industrial machine are fitted into standardised cast iron frames. All machines of the same operating principal and power level are fitted into the same frame size. More information regarding this can be found in Appendix B. The LS PMSM prototype will be fitted into a standard frame. Table 3.3 below contains the standard IEC frame of induction motors in the LV LW range.

Table 3.3: IEC standard IM frames [1, 12]

Power (kW)	Frame size	Frame diameter (A,mm)	Frame length (BB,mm)
7.5 kW	132M or 132L	216	225, 250
5.5 kW	132 S	216	187
4 kW	112M	190	117
3 kW	100L	160	173

From the four frame sizes selected to be the possible frame, the biggest frame is that of the IM counterpart. As part of the design specifications the motor may not be bigger than that of the IM counterpart. The smaller size frames are also of interest as they can be a frame option. In general a PMSM can be 25%-40% smaller than an IM of the same power rating [13]. The size reduction can be mainly accredited to the lower operating temperatures of a PMSM. The lower operating temperature is due to the absence of rotor currents in the PMSM, thus the stator can handle more current and in turn consume more power [13]. As an LS PMSM operates as a PMSM at rated speed, the assumption can be made that it is possible to design the LS PMSM to fit in a smaller frame. Before an adequate frame can be selected the stator length and inside diameter need to be determined.

3.2.2 Selecting an adequate frame size

The first step in determining the main dimensions of the machine is selecting an adequate tangential stress (σ), also known as the shear air gap stress value. The tangential stress of a machine is the main torque producing component when it acts upon the rotor surface.

The tangential stress of the machine is an empirically determined value and can be selected by the designer as long as it falls within the given boundaries. The tangential stress boundaries are determined by using the highest and lowest linear current density (A) and air gap flux density (B_δ) values as well as the specified power factor [1].

$$\sigma = \frac{A \hat{B}_\delta \cos \varphi}{\sqrt{2}} \quad (3.1)$$

Equation 3.1 is derived from the Maxwell stress theory with respect to the tangential component that produces torque when acting upon the rotor surface. This equation assumes the both the A and B_δ to be sinusoidal thus obtaining the average tangential stress. A creates the tangential field strength (H_{tan}) in an electrical machine thus $A = H_{tan}$. A more in-depth explanation is provided by [1].

Table 3.4 contains these boundaries. As an LS PMSM is not included in the data found in [1], no tangential stress values were found to aid in the design. Thus it was decided to select corresponding values from Table 3.4. The non-salient pole synchronous machine's values was not used as an LS PMSM mainly consists of a PMSM and an IM

Table 3.4: Tangential stress values of AC machines [1]

Tangential stress (σ , Pa)	Induction machine	PMSM	Non-salient synchronous machine	LS PMSM
Minimum	12 000	21 000	17 000	21 000
Average	21 500	36 000	36 000	28 000
Maximum	33 000	48 000	59 500	33 000

The boundary values selected for the tangential stress of an LS PMSM must be within the region that overlaps. This region is between the minimum value of the PMSM and the maximum value of the IM. Another method can also be used to calculate the average between the two maximum values and two minimum values. The two averages may then be used as boundary conditions. Further investigation is needed to determine if this assumption is valid.

As stated, the tangential stress of the motor is the acting rotor surface torque component. The rated torque (T_{rated}) of a machine can be calculated by

$$T_{rated} = \sigma r_r S_r \quad (3.2)$$

with r_r the radius of the rotor in metres and S_r the surface of the rotor in m^2 . By submitting the equation used to calculate the surface of a cylinder the torque can be calculated by

$$\begin{aligned} T_{rated} &= \sigma r_r (2\pi r_r l) \\ T_{rated} &= \sigma 2\pi r_r^2 l \\ T_{rated} &= \frac{\sigma \pi D_r^2 l}{2} \end{aligned} \quad (3.3)$$

with D_r being the diameter of the rotor and l the active length of the stator/ rotor both in metres. The rated torque of the LS PMSM is

$$T_{rated} = \frac{P}{\omega_r} = \frac{P}{\frac{2}{p} 50f} \quad (3.4)$$

with P the rated power, ω_r the rotor speed in rad/s, f the stator frequency and p the number of pole pairs, thus the rated torque is calculated as 47.75 N.m

In Equation 3.2, there are now only two unknown variables, as the tangential stress of the motor only needs to be within the selected boundaries. The two variables are the stator inside diameter and the active length of the stator. To determine the variables, one must be chosen and the other calculated. To do this, Equation 3.3 needs to be rewritten as

$$\sigma = \frac{2T_{rated}}{\pi D_r^2 l} \quad (3.5)$$

The variable sweep must be done for both values to determine if the selected value falls within the tangential stress boundaries. Both the sweeps will be plotted on a graph. To aid in the motor sizing process, an Excel sheet was developed to help with the parameter identification as well as determining the smallest frames (as in Table 3.3) that could be used for the LS PMSM prototype.

It was determined that a 112M frame was the smallest frame that could be used. A 100L frame could not be used as it would not provide a valid solution that would fit into the frame. The 112M frame will be used for the first iteration in determining the D_r and l . The frame's inside volume is $\pm 40\%$ smaller than the 132M frame.

To aid in the selection of the two variables the empirical length/diameter relation ratio rule (X)

$$\begin{aligned} X &\approx \frac{l}{D_{si}} \\ X &= \frac{\pi}{2p} \sqrt[3]{p} \\ X &= 1 \end{aligned} \quad (3.6)$$

can be used, with D_{si} the inside stator diameter and p the number of pole pairs [1]. There are however some machines where this rule cannot be applied such as non-radial machines, DC machines and synchronous machines with $p = 1$.

For the first inside stator diameter (D_{si}), two thirds of the outside diameter (D_{so}) was used. Table 3.6 contains the values selected and the minimum and maximum lengths calculated with their corresponding ratios.

Table 3.5: Variable sweep of D_{si}

Selected D_{si} (mm)	Calculated l_{min} (mm)	Calculated l_{max} (mm)	$X \approx \frac{l}{D_{si}} \approx 1$	
			X_{min}	X_{max}
125	59	92	0.47	0.74
120	63	100	0.53	0.83
115	70	109	0.61	0.95
110	76	120	0.7	1.09
105	83	131	0.8	1.25

The same process was followed to identify l . The first value for l was also two thirds of the frame length. Table 3.6 contains the results and the corresponding diametric boundaries.

Table 3.6: Variable sweep of l

Selected l (mm)	Calculated D_{si_min} (mm)	Calculated D_{si_max} (mm)	$X \approx \frac{l}{D_{si}} \approx 1$	
78	108	136	0.72	0.56
95	98	123	0.96	0.77
105	93	117	1.13	0.9
110	91	114	1.2	0.95
115	89	112	1.3	1.02

From Table 3.6 and Table 3.7 the following can be concluded: for the tangential stress of the LS PMSM to be within the selected boundaries, the length of the stator as well as the inside diameter must be between 90mm to 120mm. When selecting the two values, the corresponding boundaries need to be recalculated to determine if the value can be used. This is a preliminary sizing and the selected values may be changed at a later stage if the design requires it. It was decided to choose values near the lower end of the tangential stress boundaries. By doing this, the rotor volume ($\pi D_r^2 l$) is bigger. This will aid in the design of the rotor. A rotor with a smaller diameter may use less material but has a more complex design.

The selection was made with the aid of Figure 3.5. The red and blue lines are the tangential boundary conditions. The green hyperbole represents the calculated tangential stress value with D_{si} fixed at 114 mm and varying l . The orange hyperbole represents the calculated tangential stress value with l fixed at 110 mm and varying D_{si} . The intersection between the hyperbole and straight lines represent the corresponding variables boundaries. The arrows indicate the two selected values. Both selected values fall within the tangential stress margins.

At this point in the design a 112M frame would be adequate to house both the stator and rotor, as both the frame's diameter and length are sufficient. If it is needed to increase the outer diameter of the stator (D_{so}) during the stator design, or the length of the end-windings is greater than the allowed length, a 132 frame will be used.

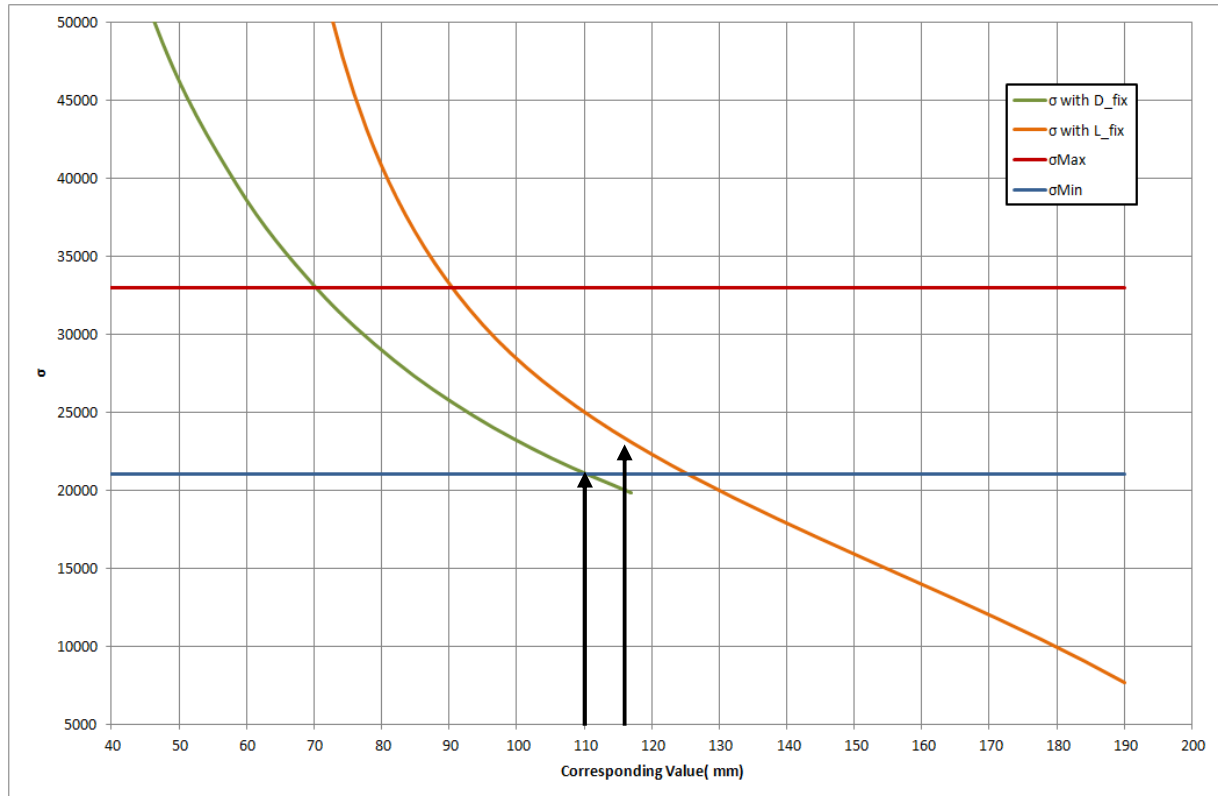


Figure 3.5: Selection of main machine dimensions with Excel tool

The final main dimension that needs to be defined is the air gap length (δ , distance between D_{si} and D_r). The air gap is one of the crucial parameters of an electrical machine and greatly influences the machine's characteristics. Even though this parameter's importance is so high, no theoretical optimum solution has been derived for this length as the air gap length influences many other operating characteristics. Thus empirical equations are used to select this parameter. This method calculates the air gap length as a function of the machine's power. For a 50Hz electrical machine the air gap length is calculated by

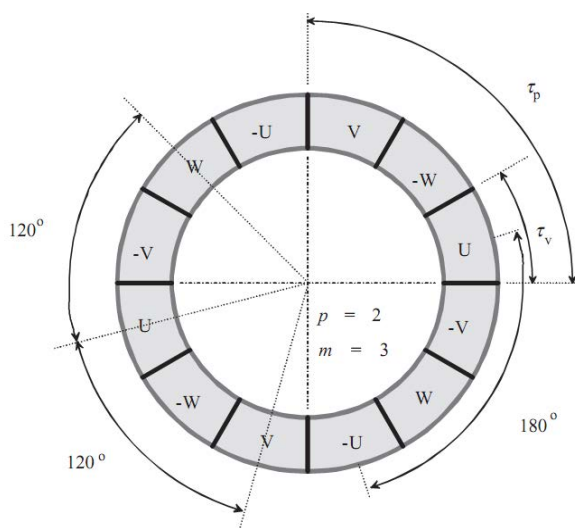
$$\begin{aligned} \delta &= 0.18 + 0.0006P^{0.4} \quad \text{when } p > 1 \\ \delta &= 0.39 \approx 0.4 \text{ mm} \end{aligned} \quad (3.7)$$

with P the power of the machine in Watts, p the pole pairs and δ the air gap length in mm [1]. The smallest manufactured air gap is 0.2mm; this however is usually used in high-speed and very heavy duty machines as this increases the machine manufacturing time and cost drastically. An air gap of 0.5mm is used on general purpose LV machines. This was also measured on opening different electrical machines

Table 3.7: Final main dimensions of the LS PMSM

Selected l (mm)	Selected D_{si} (mm)	Air gap δ (mm)	Tangential stress
110	114.5	0.5	21262.73

Before any focus can be placed on the physical design of the stator, a good understanding of the stator topology needs to be in place. In electrical machine design there are several stator topology configurations. The topology configuration is influenced by the design parameters as set in Chapter 1 and Section 3.1.2 and by the designer's preferences. For the design of the LS PMSM prototype's stator, a 3-phase, 4-pole topology must be used as indicated in Figure 3.6. The figure is a front view of a stator.



In Figure 3.6, U, V and W represent the three electrical phases of stator. The negative sign indicates that the conductors in that space (phase zone distribution (τ_v)) are installed so the current flows in the opposite direction. The phase zone distribution forms part of the pole pitch (τ_u) of the stator. A four-pole machine has four pole pitches and in each of these areas, each phase has to be present. This can be either positive or negative with respect to the coil conductors. Thus for a three phase machine a pole pitch contains three phase zone distribution sections. Furthermore within the phase zone distribution is the slot pitch area (τ_s).

The slot pitch is the area allocated in the stator for the stator slot. The size of this parameter is dependent on the stator slot count.

In the figure it can also be seen that the layout of each pole is the same (U,V,W), but when moving from one pole pitch to the next, the phase zone distribution needs to be 180° electrical degrees apart. By keeping this statement in mind it is clearly visible that this stator topology is symmetrical with an 180° shift. The main difference between the two topologies is the number of pole pitches. Table 3.8 below contains the formulas to calculate the stator parameters. In the table p represents the pole pairs, D_{si} the stator inner diameter, m the number of phases and Q_s the amount of stator slots.

Table 3.8: Stator parameter formulas

Parameter	Symbol	Equation in degrees	Equation in metres
Pole pitch	τ_p	$\tau_p = \frac{360}{2p}$	$\tau_p = \frac{\pi D_{si}}{2p}$
Phase zone distribution	τ_v	$\tau_v = \frac{\tau_p}{m}$	$\tau_v = \frac{\tau_p}{m}$
Slot pitch	τ_u	$\tau_u = \frac{360}{Q_s}$	$\tau_u = \frac{\pi D_{si}}{Q_s}$

Now that the stator zoning has been defined, the active length of the stator is next. The active length of a machine is also referred to as the stator stack or just the stack. The length of the stack is defined as the distance from the first lamination to the last lamination in metres and is indicated by l . This however is not the equivalent length of the stator. To calculate the equivalent length of the stator the effect of field fringing has to be taken into account. Field fringing occurs at the end of the stack or in ventilation ducts (only in very large radial flux machines). To understand the effects of field fringing on the equivalent length (l') of the machine, the focus must first be placed on how and where it occurs. Figure 3.7 below illustrates this.

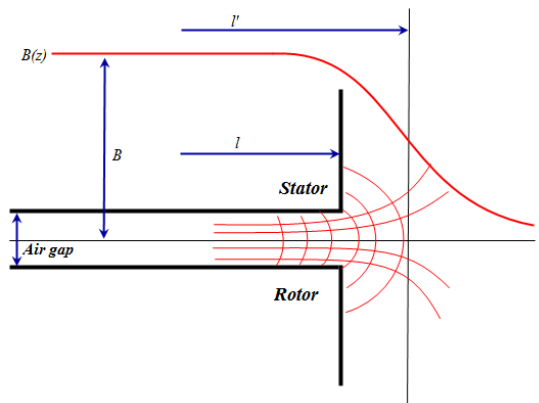


Figure 3.7: Field fringing effect at the end of the stator stack [1]

In the figure l is the physical length of the stator stack, l' the equivalent stator length, and the δ air gap length. The flux density (B) of the machine is a function of the length of the stator in the z direction thus $B=B(z)$. The assumption is usually made that the flux density is constant over the z , but the flux density decreases as the end at the stator stack is reached. Flux in the end sections of the stator leaves the stator along the x and y axis at the point where $z = l$. This effect on the flux density of the machine is known as the field fringing effect and is also indicated in the above figure. The majority of machine calculations use the equivalent length of the stator. This length is approximated by Equation 3.8 [1,3].

$$l' \approx l + 2\delta \quad (3.8)$$

Thus the approximated equivalent length of the stator is dependent on the air gap length. If a small air gap length is selected (as in most designs), the difference between the equivalent length and the stack length can be seen as the same value. Thus it is acceptable to use the stator stack length as the value used in calculations. From this point on, in the event the equivalent length of the stator is used in a formula, the stator stack length can be used.

$$l' \approx l \quad (3.9)$$

Both the radial and axial stator parameters have now been defined. This equation and these assumptions will thus be applicable from this point on.

3.3.2 Process of design

In this section the design process as illustrated by Figure 3.2 will be discussed in depth in a step by step manner. The needed design parameters were formulated in Section 3.1.2; thus the next phase in the stator design is to calculate the needed conductors per coil slot as well as the size of the needed slot.

3.3.2.1 Determining winding type and turns per coil

The main goal of this section is to clearly define the needed steps to determine the coils per coil slot. This will then be used to calculate the needed stator slot area. The method followed in this section is as in [1], and the additional literature used will be referenced accordingly.

Before starting with the stator design, the pole and slot pitch have to be calculated first. The pole pitch for the machine is 90° or 0.09 m and the phase zone destination is 30° or 0.03 m. The slot pitch can only be calculated once the slot count is known.

Select winding type and calculating the winding factor (k_w)

This is the most important and time consuming step in designing the stator. The main forming equation used to define the stator slot count (Q_s) is done with

$$\begin{aligned} Q_s &= 2pmq \\ Q_s &= 12q \end{aligned} \quad (3.10)$$

where q is the slots per phase zone and p the number of pole pairs. This is used in conjunction with k_{wv} (v^{th} winding factor harmonic) to select a suited design for the stator. k_{wl} is used to calculate the turns per coil at a later stage, the winding factor is used to increase the needed turns in the coil due to the selected stator winding configuration.

By calculating and using k_{wv} , the best suited stator can be selected. k_{wv} are composed of three sub-components namely distribution factor (k_{dv}), pitch factor (k_{pv}) and skewing factor (k_{sqv}). The v in each factor is v^{th} harmonic; for motor design it is only necessary to inspect up to the 66th harmonic [1,3]. To calculate the k_{wv} of the stator the product of k_{dv} , k_{pv} , k_{sqv} must be calculated, thus

$$k_{wv} = k_{dv} k_{pv} k_{sqv} \quad (3.11)$$

The individual factors are discussed and the equations used to calculate them are derived in Appendix C. The individual components fundamental component is also calculated in the appendix and listed in Table 3.9

Table 3.9: Compilation of various winding factors for the LS PMSM

Q_s	q	$kd1$	$kp1$	$ksq1$	$kw1$	Comment
24	2	0.9659	1	0.9886	0.9549	
24	2	0.9659	0.9659	0.9886	0.9223	Short pitched
36	3	0.9597	1	0.9949	0.9548	
36	3	0.9597	0.9848	0.9949	0.9403	Short pitched
48	4	0.9576	1	0.9971	0.9548	
48	4	0.9576	0.9914	0.9971	0.9466	Short pitched

Six possible winding layouts for the LS PMSM stator winding layout are considered and listed in Table 3.16. Each of the winding factors will be inspected and eliminated on grounds of the slot harmonics up to

the 66th harmonic and its fundamental component. Another important factor to consider is the physical size the selected winding will take up, the required fill factor and the manufacturing. The first step in selection is to compare a double layer stator to a short pitched double layer. This is done with the aid of the harmonic plots of the two options with the same q value.

Figure 3.8 to Figure 3.10 contains the slot harmonic plot comparison between a short pitched double layer and a non-short pitched double layer with the same slot count. The order of the slot harmonics is calculated with $\pm 2mqc + 1$ with c being 1, 2, 3,... [1].

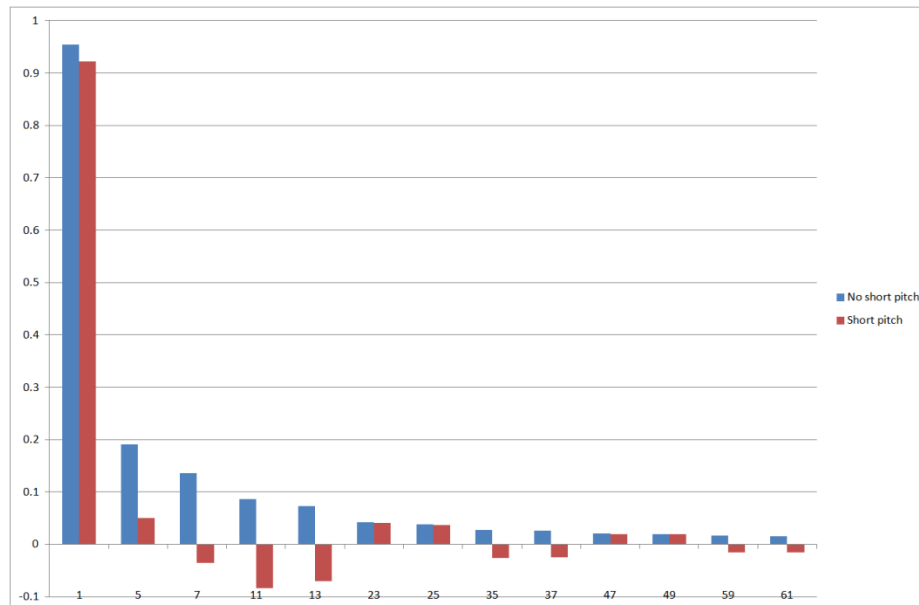


Figure 3.8: 24 double layer vs. short pitched 24 double layer

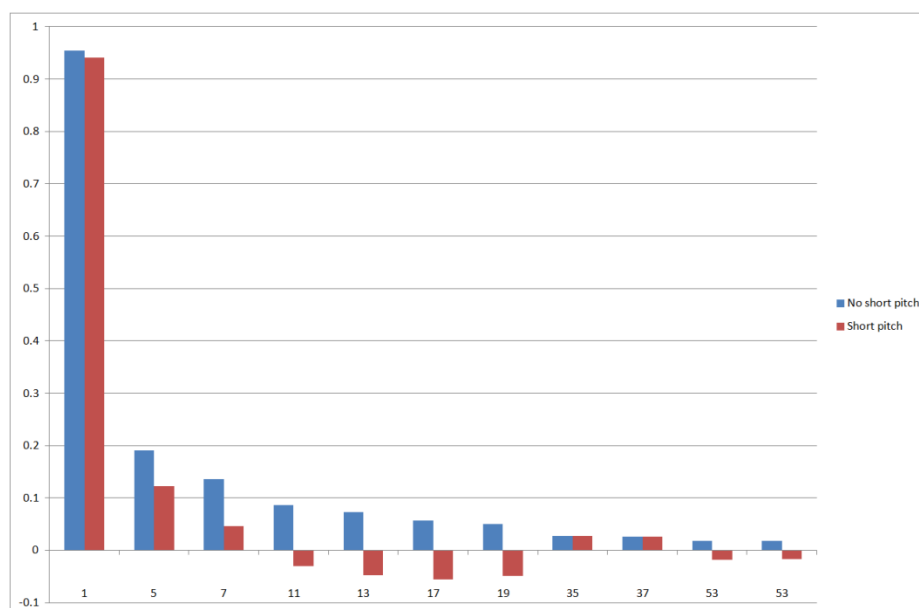


Figure 3.9: 36 double layer vs. short pitched 36 double layer

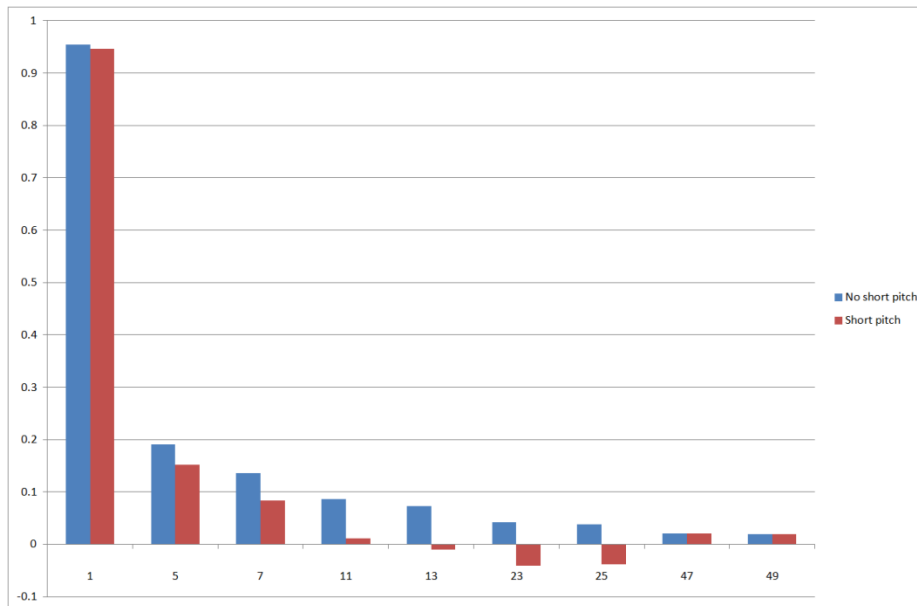


Figure 3.10: 48 double layer vs. short pitched 48 double layer

In the figures it is clear that by incorporating short pitching in the design the slot harmonics is greatly reduced. The reduction is the highest at 5th, 7th, 9th, 11th and 13th harmonics in all three figures. Although the fundamental component is also reduced in all three cases, the highest fundamental component reduction is for the 24 slot short pitched stator at 3.2%. Figure 3.11 contains the percentage slot reduction of each stator configuration due to short pitching.

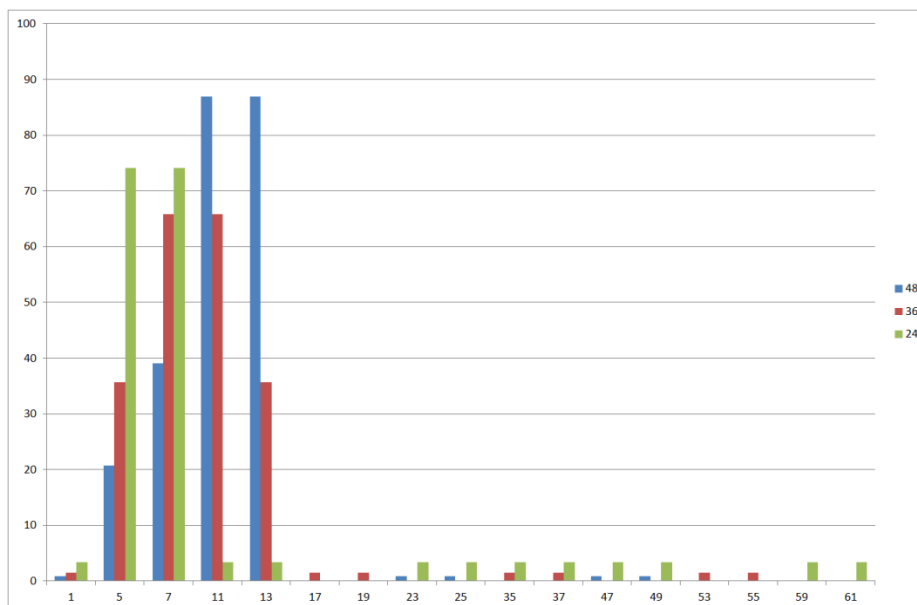


Figure 3.11: Percentage slot harmonic reduction due to short pitching

From an efficiency point of view a short pitched stator configuration is more efficient, thus by incorporating short pitching; a more efficient and more stable machine will be constructed. Figure 3.12 contains the slot harmonic plots of the three short pitched stators that are considered for the prototype.

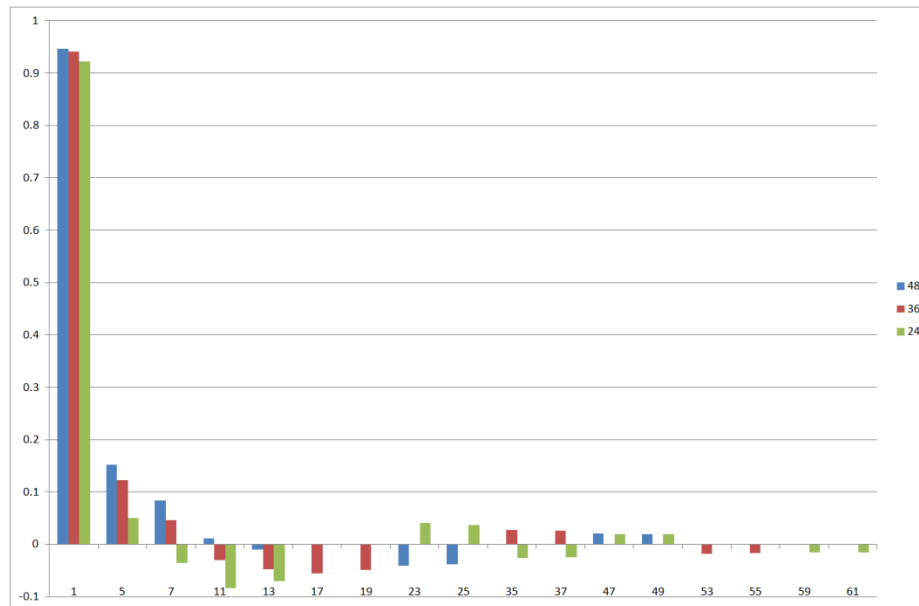


Figure 3.12: The remaining three winding factor harmonic plots

To select between the three remaining winding configuration is not simple and is not a single solution. From the 17th harmonic upwards there is no notable difference in magnitude between the different coinciding slot harmonics but the 48 slot stator tends to zero faster. When focus is place on the fundamental component of each stator, the 24 slot pitched stator is much lower than the 36 and 48 slot short pitched stators and is eliminated on this grounds. Another advantage that a stator with more slots has are an added reduction in cogging torque [14] and a smoother flux and mmf waveform.

From a manufacturing and size point of view, a short pitched 48 slot double layer stator will be more expensive to manufacture. Furthermore the stator inner diameter of the prototype is smaller than that of an IM with the same power rating, the available slot area is lower and this may introduce additional problems during the electromagnetic design of the stator. Another argument against the 48 double layer stator configuration is that its harmonic plot is very similar to that of a 36 double layer stator. The 48 slot stator slot harmonic component is also higher at the lower harmonics. Thus for the prototype a short pitched 36 slot double layer winding configuration will be used.

Defining Air gap flux density (B_δ)

The air gap flux density (B_δ) of an electrical machine must directly correlate with the selected tangential stress of the machine. Keeping this in mind, B_δ for a PMSM is between 0.85 – 1.05 T and for an IM it is between 0.7 – 0.9 T. For machines in the size range of the LS PMSM prototype, the machine's stator current density can be at the higher end of the values in Table 3.10. This is because the internal temperature of the machine is lower and is better regulated using forced air ventilation. This is not the case with larger machines. By selecting a higher current density, a lower B_δ value can be selected to provide the calculated tangential stress value.



Table 3.10: Stator current density for both IMs and PMSMs

	Induction machine	PMSM
Linear Current Density (A) kA/m	30 - 65	35 - 65
Current Density (J) A/mm ²	3 - 6	3 - 8

Before the final B_δ value can be determined, a value must be selected with aid of the tangential stress formula

$$\sigma = \frac{AB \cos \varphi}{\sqrt{2}} \quad \text{with } \cos \varphi = 0.9$$

thus

$$B = \frac{\sqrt{2}\sigma}{A \cos \varphi} \quad (3.12)$$

For the linear current density to fall within range of the boundaries set in Table 5.16, B_δ is selected as 0.85 T.

Calculating N_s

Now that the winding configuration and the air gap flux density have been selected the number of coils per phase (N) winding must be calculated. This can be done by adapting the fundamental equation for electromotive force

$$e = -k_w N \frac{d\Phi}{dt} = -k_w N \Phi$$

$$N = \frac{-e}{k_w \Phi} \quad (3.13)$$

where e is the applied voltage in motor mode or induced electrical voltage in generation mode, k_w the winding factor and Φ the magnetic flux in the machine [1]. Equation 3.13 can be adapted even further for rotating machines by rewriting it as

$$N_s = \frac{\sqrt{2}E_m}{\omega_e k_{w1} \Phi_m}$$

$$N_s = \frac{\sqrt{2}E_m}{\omega_e k_{w1} \alpha B_\delta \tau_p l} \quad (3.14)$$

with N_s the number of coil turns in the phase windings, E_m as the main electromotive force in volts, ω_e the electric angular velocity in rad/s and α is the coefficient for the arithmetical average of the flux density in the x -direction. The value for α in this case is selected as 0.75 and the final value for α is dependent on the

waveform of the air gap flux density. α is then calculated as a function of the air gap flux density waveform length over the pole length and the value of α seldom exceeds 0.85. The final value will be calculated once the rotor topology and magnet sizing is done. Small deviations on α do not have a great influence on N_s .

As the motor will be connected in Star configuration, the phase voltage is calculated by $525/\sqrt{3} = 303\text{V}$. Table 3.11 contains the relative information needed for use in Equation 3.14.

Table 3.11: Values for calculations of N_s

Symbol	Value	Description
E_m	303V	Phase voltage
ω_e	100π rad/s	Electrical angular velocity
k_{wl}	0.9403	Winding factor
α	0.8	Pole arch flux distribution coefficient
B_δ	0.85 T	Air gap flux density
τ_p	0.089 m	Pole pitch
l'	0.115 m	Active length

By substituting the values in Table 3.11 in Equation 3.14 the number of turns per coil is calculated as 208.42. This value has to be rounded to the closest value capable of producing an integer value with Equation 3.15 [1]. Thus N_s is rounded to 216 as this gives 36 turns per slot.

$$z_{Q_s} = \frac{2am}{Q_s} N_s \quad (3.15)$$

In Equation 3.15, a represents the amount of parallel paths in a single coil, in this case $a = 1$.

The new value of N_s must then be used to check if the value of B_δ is still within range. The new value of B_δ is 0.868 T which is an increase of less than 3%. This is acceptable, so the amount of turns per coil is 216.

Calculating slot area.

To calculate the required slot size to house the turns per slot, two aspects must be kept in mind. The first is the space needed for the slot area and the second aspect is the fill factor that needs to be added to facilitate the gaps caused by the shape of the conductors.

The required area for a single conductor is dependent on the selected current density (J_s). This can be calculated using:

$$S_{cs} = \frac{I_s}{J_s} \quad (3.16)$$

with the estimate rated current calculated as

$$I_s = \frac{P}{m\eta V_{ph} \cos \varphi} \quad (3.17)$$

$$I_s = 9.65 \approx 10A$$

To select the value for the current density the following aspects must be kept in mind: the machine cooling method, the power rating of the machine and physical size of the machine. The selected value of J_s must be within the boundaries of Table 3.10 but can be selected at the higher end because smaller size machines, at a low power rating tend to regulate temperature better. This is because industry manufactured frames incorporate forced air cooling over the frame. This allows designers to select a higher current density value for the machine. The added advantages for selecting the current density value as high as possible is the reduction on manufacturing cost as less copper is used to wind the machine. By selecting the stator current density as 6.5 A/mm^2 the needed copper area per conductor is

$$S_{cs} = \frac{I_s}{J_s}$$

$$S_{sc} = 1.54 \approx 1.6 \text{ mm}^2$$

With the aid of S_{sc} , the needed wire radius (r) can be calculated and a suitable wire can selected. The wire radius is calculated as 0.714 mm, thus the diameter is 1.43mm. This value can now be compared to the available options as provided by SWG (Standard Wire Gage) or any available diameter provided by a wire manufacturer. Table 3.12 contains the possible SWG wire diameters.

Table 3.12: SWG

SWG	Diameter (mm)
14	2.032
15	1.829
16	1.626
17	1.422
18	1.219

SWG 17 will be used to wire the stator because it has the diameter closest to the calculated. As the diameter of SWG 17 is smaller than the calculated diameter the current density has to be recalculated. The new current density is 6.3 A/mm^2 and is still close enough to 6.5 A/mm^2 as there is only a 3% reduction from the previously selected value.

Another technique often used in the design and manufactures of stators is to use multiple wires per coil turn. By incorporating this technique there are a couple of advantages introduced in the design and



manufacturing. The first is quite obvious: a wire with a thinner diameter bends more easily than a thicker wire. Thus the coil is easier shaped and modified during manufacturing [16]. The second advantage aids in a better stacking factor for the same slot. This is because of the reduction of “dead space” between the round conductors. This can be seen in Figure 3.13. The copper wire in both a) and b) covers the same area. For the same copper area but with multiple wires, a smaller slot can be designed as the conductors can be placed closer together. A fine balance between the amount of multiple strands and ease of manufacturing must be considered, as a high increase in strands may be difficult to handle and stack during manufacturing.

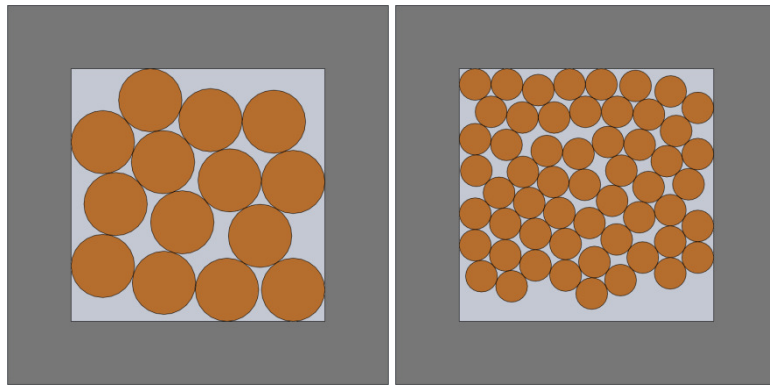


Figure 3.13:a) 14 coil turns with a single wire. b) 14 coil turns with 4 wires (56 wires)

For the multiple wires per coil turn, two 1 mm diameter wires will be used. This diameter was selected from the result found in the physical inspection of multiple LV machines below 7.5 kW. All these machines had two or three wires per coil turn ranging with diameters between 0.7 – 1 mm. Thus by introducing this design technique in the stator it would relate to the industry design trend. The current density for the two 1mm wires is

$$J_s = \frac{I_s}{2S_{cs}} = \frac{I_s}{2\pi r^2}$$

$$J_s = 6.36 A/mm^2$$

The current density with the multiple wires is still below the original value. By selecting the diameter at 0.9 mm the current density is $7.8 A/mm^2$ which is too high for the stator.

The next step in the design is to calculate the copper area in a single slot. As there are 36 turns in a single slot with 72 conductors the copper area in a slot is $S_{72SC} = ar^2\pi \approx 60 mm^2$

As shown in Figure 3.15, the slot area must be greater than the copper area. This is to accommodate the dead space between conductors. The slot area is calculated by incorporating a filling factor (ff) which is the relationship between the copper area and the slot area

$$ff = \frac{S_{sc}}{S_{ss}} \quad (3.18)$$

but by rewriting Equation 3.18 the required slot area can be calculated by

$$S_{ss} = \frac{S_{sc}}{ff_{ss}}$$

The value of ff_{ss} must be selected by the designer with respect to the size of the machine. The value of ff_{ss} can be between 0.5 to 0.6 for LV LW machines and 0.3 to 0.45 for higher power rating machines. The smaller the value of ff_{ss} , the larger the slot is. This is to accommodate the larger areas between the conductors as they have a larger diameter. For the LS PMSM the fill factor is selected as 0.5, thus the required slot area is 150 mm²

3.3.2.2 Selecting stator slot and calculating stator parameters

After the required slot area is calculated the next step is to select a slot shape for the stator. There are various slot types available that can be used. Figure 3.14 contains the most commonly used slot types.

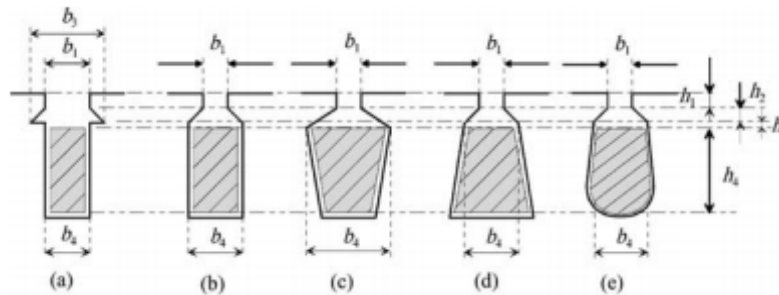


Figure 3.14: Stator slot shapes possibilities [1]

In the figure, (b-e) is most commonly used in LV machines. Selecting the best suited slot is usually the designer's preference but (c) and (e) are the most widely used in small machines [24]. By selecting a slot with a round top end (e), more windings can be fitted in the same slot than in a block shape (a-d). This is because there is less dead space between the conductor and slot side. The other advantage a round ended slot has is a lower flux density value at the outer end of the slot. Flux density tends to be higher at sharper edges than at rounder edges. This can be seen in Figure 3.15 where a round ended slot is compared to a square shaped slot.

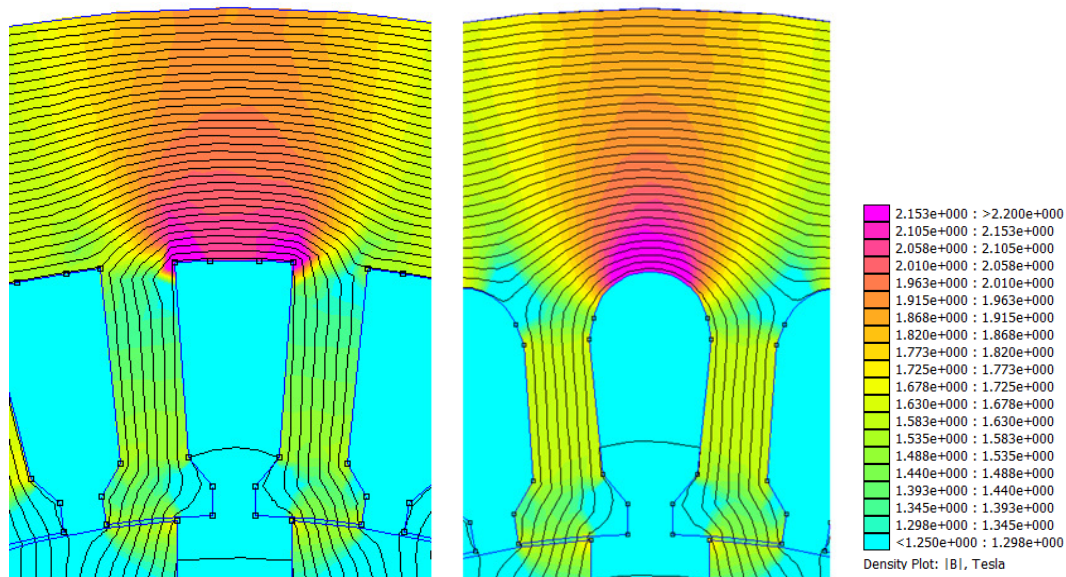


Figure 3.15: Slot shape comparison in terms of flux density

The flux density boundary for the figure is $1.25T - 2.2T$. The slot area of the two slots as well as the tooth width is more or less the same. It can be seen that the saturated area in purple above the slot of the square shape slot is much bigger and covers a larger area than the round ended slot. Thus for the prototype machine the round ended slots will be used as they will have lower core losses as the saturated areas will be smaller.

Once the slot shape has been selected the slot can be designed for the area calculated with Equation 3.18.

Defining slot height and teeth width

The height and width of the stator teeth influences the flux density in the stator. By selecting the values of the stator slot height (h_s) and tooth width (b_{st}) incorrectly the stator can be saturated in either the back bone and/or the teeth, thus increasing the losses of the machine. The required slot area must also be kept in mind as this directly influences both h_{sy} and b_{st} . To reduce the number of iterations needed to determine the best distance to flux density relationship, the flux density values can be selected first and h_{sy} and b_{st} can be calculated. Usually the slot's height spans half of the stator yoke. Figure 3.16 indicates the position of varies stator variables. h_s and b_4 are slot forming variables. Both these variables are influenced by h_{sy} and b_{st} . b_4 however is dependent on the slot shape and whether parallel or non-parallel tooth design is used as h_{sy} and b_{st} are the areas in the varies areas where the flux density will be the highest.

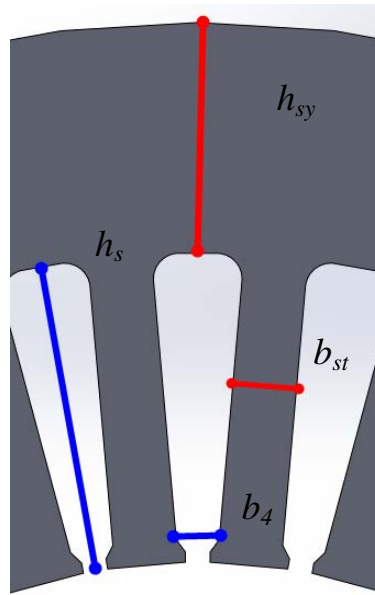


Figure 3.16: Stator tooth and slot height dimensions

The values of the tooth flux density (B_{st}) and back bone or yoke flux density (B_{sy}) are listed in Table 3.13. The values for an LS PMSM are the overlapping values between the IM and PMSM.

Table 3.13: Stator tooth and yoke flux density values [1]

	Induction machine	PMSM	LS PMSM
Stator yoke (B_{sy})	1.4 - 1.7	1.0 – 1.6	1.4 – 1.6
Stator Tooth (B_{st})	1.5 – 2.2	1.6 – 2.0	1.6 - 2.2

To calculate the length of b_{st} the tooth slot pitch ratio can be used as the flux flowing through the slot pitch is equal to the flux flowing in the stator tooth, thus

$$\begin{aligned}
 \Phi_m &= \Phi_{st} \\
 B_\delta \tau_u &= B_{st} b_{st} \\
 \frac{b_{st}}{\tau_u} &= \frac{B_\delta}{B_{st}}
 \end{aligned} \tag{3.19}$$

with Φ_m the main machine flux and τ_u the slot pitch [3]. To calculate h_{sy} the following can be used [3]

$$h_{sy} = \frac{B_\delta \tau_p / \pi}{B_{sy}} \tag{3.20}$$

The calculated value must be deducted from D_{so} to determine the slot height.

Designing slot shape

There are various different aspects that need to be taken into account when designing a slot. The most important is the slot area as this is the space needed for the windings. The second is the flux leakage due to the slot as this impacts the performance and more specific the starting torque of the machine [24]. Some other aspects are parallel vs. non-parallel teeth and open vs. closed slots. These choices impact the slot forming parameter as indicated in Figure 3.17

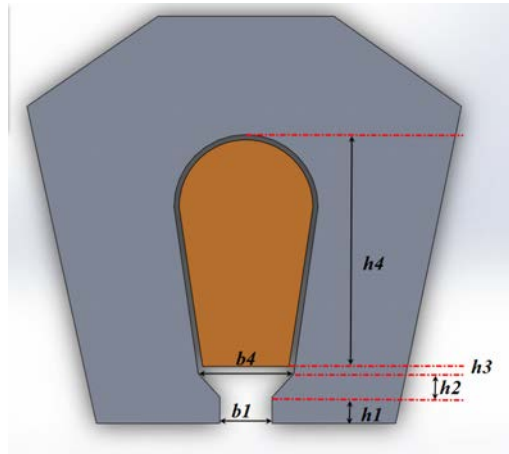


Figure 3.17: Slot forming parameters

The copper area in the figure represents the area where the windings will be placed in the slot and h_3 is the space between the windings and the slot shoe.

To calculate the leakage flux of the slot, the slots performance factor λ_{su} must be calculated. The equation used to calculate λ_{su} is influenced by the winding configuration and the slot shape [2, 24]. To calculate λ_{su} for a double layer round slot, the coil short pitching also needs to be included. λ_{su} for the selected stator design is calculated

$$\lambda_{su} = k_1 \frac{h_4}{3b_4} + k_2 \left(\frac{h_3}{b_4} + \frac{h_1}{b_1} + \frac{h_2}{b_4 - b_1} \ln \frac{b_4}{b_1} \right) \quad (3.21)$$

with

$$k_1 = 1 - \frac{9}{16} \varepsilon \quad k_2 = 1 - \frac{3}{4} \varepsilon \quad \varepsilon = 1 - \frac{y}{y_{Qs}} \quad (3.22)$$

y represents the calculated coil span where y_{Qs} equals the actual coil span due to coil short pitching. λ_{su} is directly proportional to the leakage inductance of the slot (L_{su}).

Simulating and optimising stator

Once the slot is designed, the flux density values need to be verified to determine if it's within the selected values. If it is not, the slot forming parameters needs to be adapted until it is as selected, keeping

in mind the required minimum slot area as calculated. During this phase, the saturation distribution in the back must also be reduced.

The peak flux density values are situated on the slot edge of the back yoke. To optimize the slot shape the peak flux density destitution value will be optimised. This is done by changing the slot shape on the back yoke side to reduce the area of the distribution.

As the selected slot shape contains a half radius at the back yoke the shape is altered by dividing the half radius into two quarter radii. The slot height is kept constant and the quarter radius values are reduced. This introduces a flat edge on the back yoke side of the slot similar to that of Figure 3.16.

Calculating stator parameters

Once the design is finalised the stator parameters can be calculated. These parameters are the stator winding resistance (R_l) and the inductance (L_l)

The stator coil resistance is calculated by [1, 24]

$$R_l = \rho_{Co} l_c \frac{N_s}{aS_{sc}} K_r \quad (3.23)$$

with ρ_{Co} the resistivity of copper, a the number of parallel strands per turn, K_r the a.c to d.c resistance ratio due to skin effect and l_c the coil length in metres calculated with

$$l_c = 2l + 2.4l_{yQs} + 0.1 \quad (3.24)$$

K_r for LW is equal to 1 as the skin effect in these machines can be neglected as the field penetration depth is very low. The following can be used to verify if the skin effect will have an influence on the resistance [24]

$$\delta_{Co} = \sqrt{\frac{\rho_{Co}}{\mu_o \pi f}} \leq D_{wire} \quad (3.25)$$

by using the above equation the skin depth for the copper wire is calculated as 0.094 mm which is much smaller than 1 mm thus proving that the skin effect is very small in comparison to the wire's diameter.

The stator leakage inductance is influenced by different aspects of the stator core and is highly dependent on the specific design. To calculate L_l , the performance factor of each of these aspects must be calculated and added together to form the main performance factor λ_{sm} . The performance factors that form λ_{sm} are:

- Slot leakage (λ_{su})
- End winding leakage (λ_{sw})
- Skew leakage (λ_{ssq})
- Zig-zag leakage (λ_{szz})

The Zig-zag leakage inductance is a combination of the air gap, tooth tip and differential leakage inductances [24]. The skew inductance can only be calculated once the main inductance is calculated thus λ_{ssq} must not be added to the other performance factors [1, 3, 24]. Once λ_{sm} is calculated L_l and X_l is calculated with

$$L_l = 2\mu_0 \frac{N_s l}{pq} \lambda_{sm} \quad (3.26)$$

$$X_l = \omega_e L_l$$

This is only the inductance without the skewing inductance. The skewing inductance is calculated with

$$L_{ssq} = L_l \lambda_{ssq} \quad (3.27)$$

and must be added to L_l for the stator leakage inductance. Table 3.14 contains each performance factor's equation. It also indicated if the equation is type specific or can be used in general

Table 3.14: Leakage inductance performance equations [1, 3, 24]

Performance factor	Equation	Remarks
Slot leakage (λ_{su})	$\lambda_{su} = k_1 \frac{h_4}{3b_4} + k_2 \left(\frac{h_3}{b_4} + \frac{h_1}{b_1} + \frac{h_2}{b_4 - b_1} \ln \frac{b_4}{b_1} \right)$	Only for round slots as in Figure 5.24. K_1 and K_2 is calculated by 5.26
End winding leakage (λ_{sw})	$\lambda_{sw} = 0.34 \frac{q}{l} (l_{y_{qs}} - 0.64\tau_p)$	Double layer distributed winding
Zig-zag leakage (λ_{szz})	$\lambda_{szz} = \frac{5\delta/b_1}{5 + 4\delta/b_1}$	
Skew leakage (λ_{ssq})	$1 - k_{sq1}^2$	k_{sq1} is skewing winding factor

3.3.3 Finalising the stator design

In this section the stator design will be finalised as all the relevant information has been provided to do so. Table 3.15 contains information, regarding choices made earlier in the chapter, needed to complete the stator.

Table 3.15: Stator design information

Symbol	Value	Description
D_{si} (mm)	114.5	Inside stator diameter
D_{so} (mm)	220	Outside stator diameter
τ_u (mm)	8.9	Pole pitch
l	115	Active length
Q_s	36	Number of stator slots
k_{wl}	0.9403	Stator winding factor
S_{ss} (mm ²)	150	Stator slot area

To complete the stator design the slot shape has to be finalised. To do this B_{sy} and B_{st} need to be selected to determine the height and width of the slot. The values of the flux density need to be between the boundaries set in Table 3.13. Table 3.16 contains the values with regards to the slot height and tooth width.

Table 3.16: Stator slot height and width parameters

	Yoke	Tooth
B (T)	1.4	1.7
h_{st} & b_{st} (mm)	35.5	4.5

Figure 3.18 indicated the area allocated to the slot, not the actual slot area or design, the height and width is limited by the values in Table 3.16.

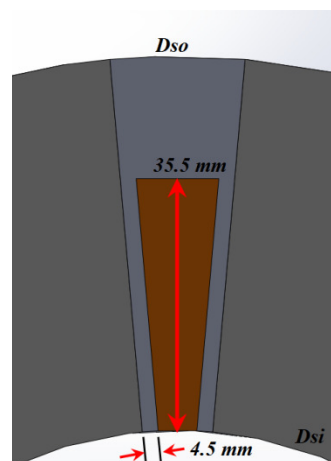


Figure 3.18: Calculated slot space in slot pitch

The initial allocated slot area was larger than 150mm², thus it was decided to reduce the slot height to 30 mm and tooth width to 6.4mm which made the slot area close to the required value. The slot shape was designed with focus placed on limiting λ_{su} as calculated in Table 3.14. A high λ_{su} increases the leakage inductance which in turn decreases starting torque of the machine. The final slot design and dimensions are listed in Table 3.17. There reduced values will lead to a decreased flux density and must be

recalculated; this will be done later in the chapter. It was decided to use parallel stator teeth for the slot design as S_{ss} was already reduced due to changing the slot height.

Table 3.17: Final stator slot design and dimensions

	Parameter	Value (mm)
	b_1	2.4
	b_4	4.2
	h_1	1.3
	h_2	1.3
	h_3	0.5
	h_4	26.5

As all the needed information regarding the factors that influence R_1 and L_1 has been set, the next step is to calculate these parameters. To calculate the both the stator resistance and inductance an Excel sheet was constructed. The sheet uses the equations listed in Table 3.14 and Equation 3.22 to 3.27. The results are listed in Table 3.18.

Table 3.18: Calculated stator parameters

Parameter	Value
R_1	1.40544 Ω
Slot leakage (λ_{su})	2.8836
End winding leakage (λ_{sw})	0.24197
Zig-zag leakage (λ_{szz})	0.1785
X_1 without skewing	2.343 Ω
Skew leakage (L_{ssq})	75.424 μH
X_1	2.3668 Ω

3.3.3.1 Stator CAD

To finalise the stator, Solid Works[®] was used to design the 3D model of the stator stack according to the slot dimensions and the other design parameters selected in this chapter.

Figure 3.19 illustrates the final 3D CAD. Although it is not visible on the figure, the core is made from laminated steel. The specific material will be selected later in this chapter. The figure contains the following (left to right): a projected view of the full stator stack, a front view of a single lamination and a sectional view of the full stator. In the sectional and projected view the skewed slots are clearly visible.

The sectional view of the stator was done along two different planes. The horizontal view was done in the centre of the slot and the vertical view was done along the centre of a tooth. The sectional view was done to aid in the visualisation of the stator skewing.

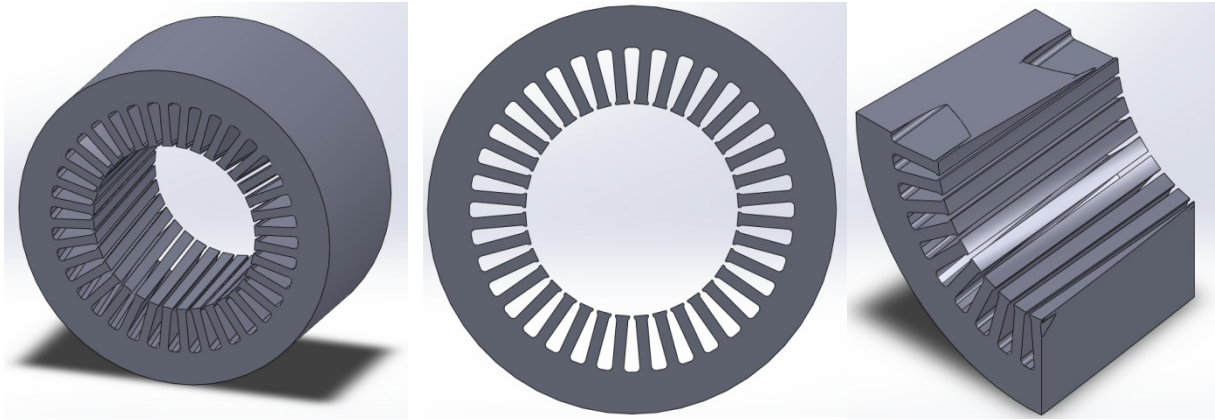


Figure 3.19: 3D CAD stator design

3.3.3.2 Lamination material

For manufacturing of the machine, available lamination material was limited were as manufacturers in South Africa. The available material for the laminations was the M400-50A and M530-50A. Figure 3.20 below contains the BH curve of both lamination materials. It can be seen that the M530-50A has a slightly higher saturation point at 1.4 T whereas the M400-50A saturates at 1.3 T making the M530-50A steel a better choice. However the iron core losses (P_{Fe} , eddy currents and hysteresis losses) per lamination differ as ρ differs. Figure 5.21 contains P_{Fe} per stator lamination for both materials. P_{Fe} was calculated by

$$P_{Fe} = \frac{V_{lam} (\pi f d_{lam} B_{max})^2}{6\rho} \quad (3.28)$$

with V_{lam} the lamination volume, d_{lam} the lamination thickness and B_{max} the maximum peak flux density in the lamination.

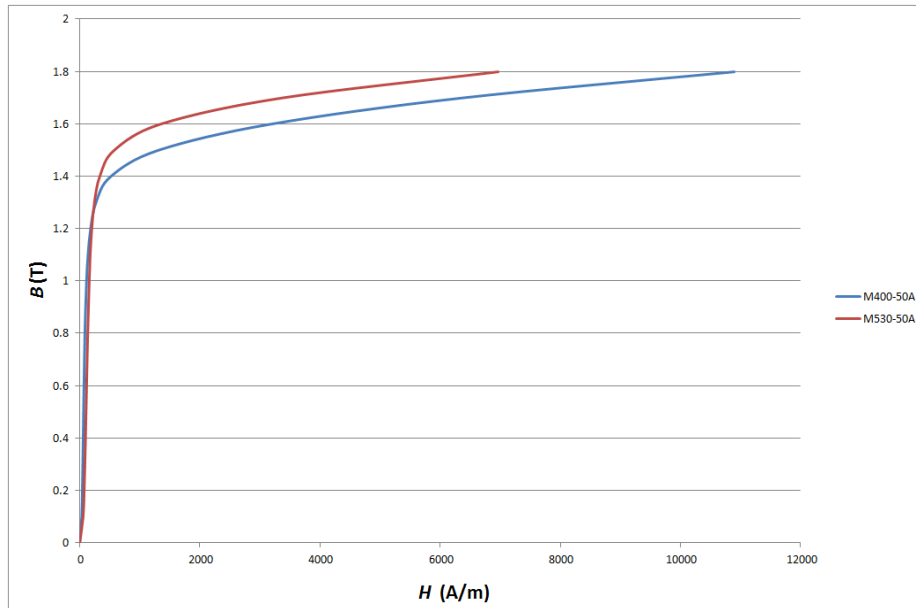


Figure 3.20: BH curve of M400-50A and M530-50A

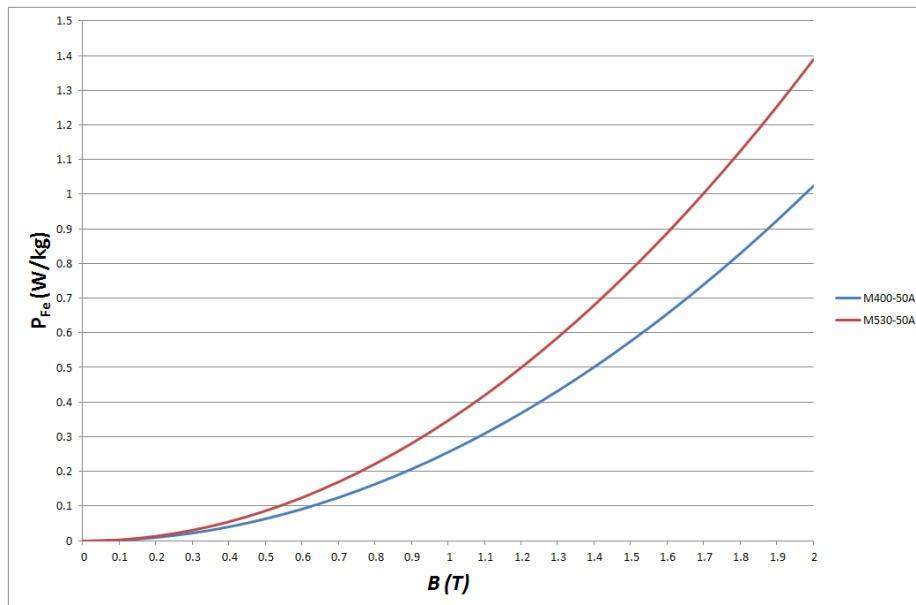


Figure 3.21: Stator core losses of M400-50A and M540-50A

From Figure 3.21 it can be seen that for both material type the losses are very low for a single lamination. However when B_{max} ($\pm 1.8 T$) is used to calculate the loss of the entire stack, the difference between the losses is more or less 100 W/kg; thus the decision is made to use the M400-50A steel for the laminations.

3.3.3.3 Verifying the stator design.

To verify the stator design, the selected flux density values in both the stator yoke and teeth will be compared to the results from the FEM simulation model. Both these values need to be within an acceptable limit of 10%. As only the stator design is complete at this stage an adapted model must be used to verify the design. The model that will be used is a quarter machine section due to symmetry. Solid

Works was used to design a model that can be used in FEMM. It was decided to use a basic surface mount rotor design. This is because the PM is placed on the surface of the rotor and the air gap flux density is easily manipulated to ensure the correct air gap flux density value. Once the final rotor design is completed the stator will be rechecked to ensure that the selected values in the stator are still obtained with the final rotor. The stator verification is done in this section to ensure that the design is adequate when the rotor is finalised.

Figure 3.22 below is the Solid Works model that was used to verify the stator design. The two quarter sections are exactly the same with regards to rotor pole to stator position. In Model A, the flux density was taken between two d -axis and Model B between two q -axis.

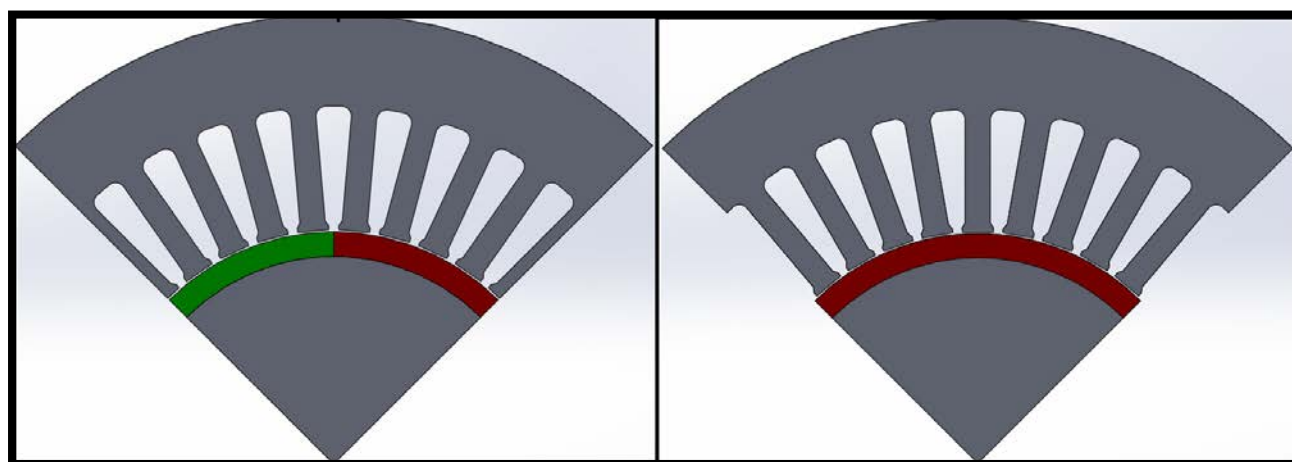


Figure 3.22: Solid Works Stator verification model a) Model A b) Model B.

To prove that the two separate models will provide nearly identical results, B_{sy} was calculated in both models. The air flux density for both models must be within 1% of each other to further ensure accurate results in the verification process.

To verify B_{sy} and B_{st} both the models will be used. The model set up in FEMM is as follows: The average peak value of the air gap flux density over the stator tooth must be 0.85 T and is situated on the d -axis of the rotor. Thus the PM properties were selected to ensure this value on the d -axis. Figure 3.23 is the simulation results containing the flux density distribution as well as the line used to calculate average values of B_{sy} and B_{st} (Black lines) and $B_{\delta avg}$. The flux density boundaries set for the plot are [0,1.8].

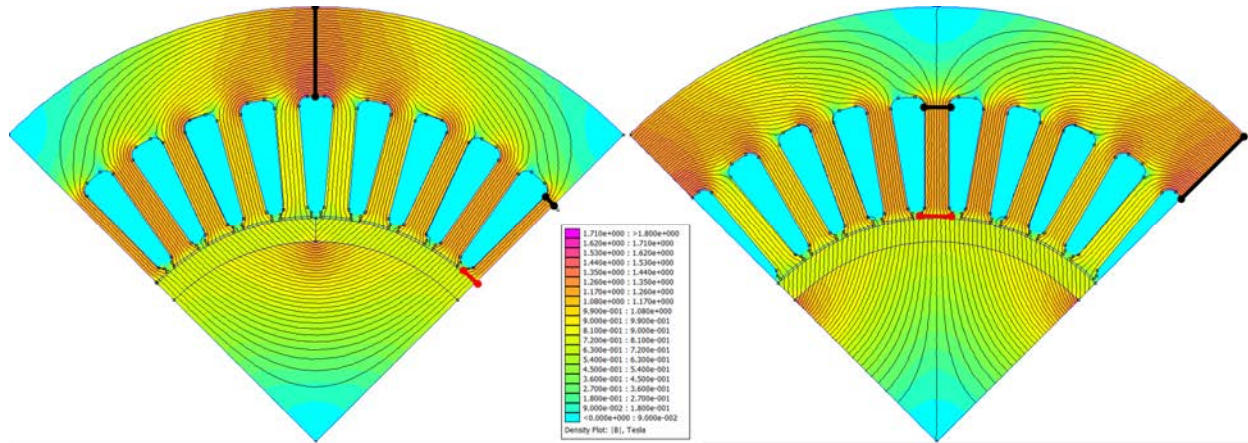


Figure 3.23: Flux density plot in FEMM of the Solid Works stator verification models.

Due to the calculated values of the slot width and height that have been reduced as stated earlier in the chapter, the flux density values in tooth and yoke need to be recalculated using Equation 3.19 and 3.20. The new values are indicated in Table 3.19 along with the data extracted from the simulation results. The new flux density values are indicated in bold. The added advantage of the reduced values is that both are still below the saturation point of M400-50A lamination material.

Table 3.19: Stator verification information

	$B_{\delta avg}$	B_{sy}			B_{st}		
		Results	Calculated	Difference in %	Results	Calculated	Difference in %
Model 1	0.85488T	1.337T	1.22 T	9.5	1.206 T	1.3 T	7.7
Model 2	0.851039 T	1.383T		13	1.233 T		5.4

From the table it can be seen that the selected flux density values both B_{sy} and B_{st} are within the range of the simulated values of both models. Thus the conclusion can be made that the theoretical stator design correlates with the simulation model.

3.4 Rotor design

As stated earlier in the chapter the rotor design is divided into two parts. The PMSM rotor is designed first to simplify the design process. Furthermore the braking torque component must be known in order to design the cage to generate enough torque to overcome it. The volume of the PMs and PMSM rotor also influence the space availability of the rotor. Once both rotors are designed they will be combined to form the LS PMSM rotor.

3.4.1 PMSM

This section will only focus on the design and design theory of a PMSM. As indicated in Figure 3.3, the starting point when designing a PMSM is selecting a topology and understanding PM theory to select the correct of PM type and grade. Appendix D contains the necessary background on PM materials and the application thereof.

3.4.1.1 Permanent magnet sizing.

Now that a better understanding of PM has been established, the next step is to determine the required magnet volume to produce the air gap flux density as selected in Section 3.3.2. It is assumed that there is no PM leakage flux inside the machine; thus the total flux produced by the magnet crosses the air gap, and a machine pole can be represented by Figure 3.24a. The magnetic circuit can however be reworked into Figure 3.24b containing only one air gap equivalent to that of the machine. This technique is derived from [2] and will be used to size the PM for the prototype, However the leakage flux must be taken into account once the initial sizing is done. This will be done with the aid of a FEM package.

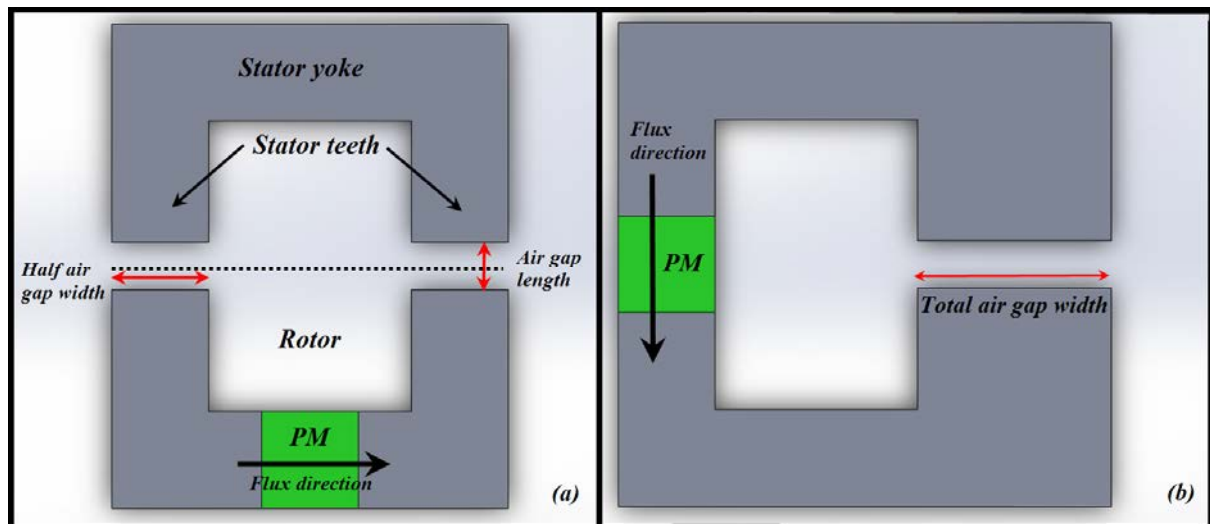


Figure 3.24: Representation of an ideal PMSM with only PM flux

As there is no leakage flux in the circuit the flux throughout the machine is equal thus

$$\begin{aligned}\Phi_{pm} &= \Phi_{\delta} \\ B_m A_m &= B_{\delta} A_{\delta} \\ B_{\delta} &= \frac{A_m}{A_{\delta}} B_m\end{aligned}\tag{3.29}$$

with B_m the operating point flux density of the PM, A_m the flux producing area of the magnet and A_{δ} the area of the rotor with the flux density value of B_{δ} . As the mmf acting on the magnetic circuit is equal to zero, then the mmf across the PM is equal to -1.

$$\frac{H_m l_m}{H_\delta \delta} = -1$$

$$H_\delta = \frac{-H_m l_m}{\delta} \quad (3.30)$$

with H_m and H_δ the magnetic field intensities in the PM and air gap. l_m represents the PM thickness. By multiplying Equation 3.30 by $B_\delta = \mu_0 H_\delta$ the following is found

$$B_\delta^2 = \mu_0 \left(\frac{l_m A_m}{l_\delta A_\delta} \right) (-H_m B_m) \quad (3.31a)$$

which can be rewritten as

$$Vol_m = \frac{Vol_\delta B_\delta^2}{\mu_0 (-H_m B_m)} \quad (3.31)$$

b)

If the magnet is operated at its maximum energy product then $H_m B_m$ can be replaced with the BH_{max} value of the selected grade magnet which in turn will give the smallest PM volume.

Equation 3.31b can be used in two ways, either by determining the required BH_{max} value if the magnet volume in the rotor is known by calculating the volume of the magnet if a certain grade magnet is selected. Before selecting the final magnet volume and dimensions, a percentage increase needs to be taken into account for the leakage flux. This increase is determined by the selected rotor topology. The PM volume also needs to be checked at both normal and operating temperature to ensure it is adequate.

3.4.1.2 Selecting a PMSM rotor topology.

In Chapter 3, PMSM rotor topologies applicable to LS PMSMs were discussed. It was decided to only inspect the four core topologies namely surface mount magnets (SMM), slotted surface mount magnets (SSMM), embedded radial flux magnets (IRFM) and embedded circumferential flux magnets (ICFM). As the focus of the design is to develop an optimised machine, the lowest number of PMs must be used in the design. Furthermore the PMs volume must also be as low as possible to reduce cost. From the literature it is clear that embedded combination topologies (ICT) tend to incorporate more magnets which often are not all the same size. These designs also tend to be time intensive. Due to the scope of the project and the time available ICT will not be considered as a topology option for the rotor.

Table 3.20 contains a summary of the four topologies. The table was constructed from information gathered during the literature survey done in Chapter 2. The four topologies are compared regarding the most influential aspects on the performance of an LS PMSM. From the table it is clear that there is a big difference in surface mount to embedded magnet rotors. This was also clear during the literature survey.

Table 3.20: Topology comparison

	<i>SMM</i>	<i>SSMM</i>	<i>IRFM</i>	<i>ICFM</i>
B_δ	$= B_m$	$= B_m$	$< B_m$	$> B_m$
<i>Starting Torque</i>	Low	Low	Moderate	Moderate
T_{syn}	High	High	Moderate	Moderate
<i>Cogging Torque</i>	High	High	Low	Low
<i>Back-EMF</i>	High	High	Moderate	Moderate
L_σ	-	-	Yes	Yes
L_q vs. L_d	$L_q > L_d$	$L_q > L_d$	$L_q > L_d$	$L_q < L_d$

Along with the information in Table 3.20 mechanical and physical aspects also needs to be taken into consideration. The volume of the rotor stack space to house both the PMs and the rotor cage is limited. This can also restrict additional components that may be needed to assemble the rotor-like locating pins or additional brackets.

The main focus of the rotor design is to use the lowest possible PM volume to produce the required air gap flux density. The other factors in Table 3.20 will also play an important role in the selection. During the literature study and survey it became clear that there is a lack of information regarding the influence of PM volume in conjunction with topologies on the air gap flux density, thus motivating further investigation. A study was done on the four core PMSM topologies as shown in Figure 3.25. The study was formulated into a publication and presented at a peer-reviewed international conference, (see Appendix E). The investigation was done using the stator design and machine dimensions for the prototype.

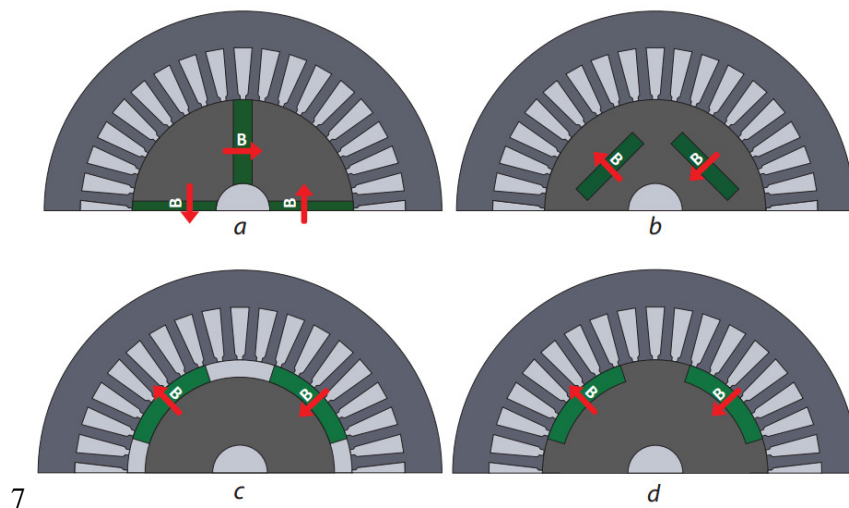


Figure 3.25: PM rotor topologies: a) ICFM; b) IRFM; c) SMM and d) SSMM [37]

Only an overview of the study will be given followed by a summary of the results. The goal of the study was to determine how a change in magnets' volume will affect the air gap flux density in both the average

peak value and the distribution over the pole without flux added by the stator coils. To investigate this, three coefficients were used, each influencing the previously mentioned aspects differently. Table 3.21 contains information regarding these coefficients. In Figure 3.25, the flux producing area of each PM is the same for each topology. This acts as the reference point for each coefficient study. ICFM was used to determine the base PM volume as this topology limits the magnet area due to the rotor radius.

Table 3.21: Influence coefficients and definition

<i>Coefficient</i>	<i>Pole arch (α_{pa})</i>	<i>Magnetic thickness (α_{mt})</i>	<i>Magnetic dept (α_{md})</i>
<i>Equation</i>	$\alpha_{pa} = \frac{\tau_s}{\tau_p}$	$\alpha_{mt} = \frac{D_{mi}}{D_{mo}}$	$\alpha_{md} = \frac{D_{mo}}{D_r}$
<i>Top Variable</i>	Pole pitch	Most inner point of the PM	Most outer point of the PM.
<i>Bottom Variable</i>	Pole arch width	Most outer point of the PM	Rotor diameter
<i>Comment</i>	Pole arch width is varied by increasing the tangential length of the PM	D_{mo} is held fixed value and D_{mi} is varied	PM magnet volume is kept constant and magnet position is varied

The pole arch coefficient was adapted from [6] while α_{mt} and α_{md} was formulated for the investigation. The magnetic depth investigation is only applicable on the embedded topologies as it looks at how the air gap flux density is influenced by the depth of the magnets inside the rotor. Table 3.22 contains the results of each topology investigation. B_δ refers to the air gap flux density and $B_{\delta D}$ to the air gap flux density distribution of the pole pitch

Table 3.22: Simulation results on air gap influence investigation

	As α_{pa} increase	As α_{mt} increase	As α_{md} increase
PM Volume	Decrease	Decrease	Constant
ICFM	B_{δ} ↓ $B_{\delta D}$ —	B_{δ} ↓ $B_{\delta D}$ —	B_{δ} ↓ $B_{\delta D}$ —
IRFM	B_{δ} ↓ $B_{\delta D}$ ↓	B_{δ} — $B_{\delta D}$ —	B_{δ} ↑ $B_{\delta D}$ —
SMM	B_{δ} ↓ $B_{\delta D}$ ↓	B_{δ} — $B_{\delta D}$ —	
SSMM	B_{δ} ↓ $B_{\delta D}$ ↓	B_{δ} — $B_{\delta D}$ —	

From the simulations it is clear that both surface mount topologies reacted in the same manner as expected. The peak B_{δ} value was not influenced by a change in α_{pa} however the average flux density value over the pole pitch increased as the PM volume increased. This in turn increases the pole flux density distribution over the pole pitch. At the point where the PM spans an entire pole pitch the surface mount topologies provides the best air gap flux properties. Furthermore at this point the flux produced by the magnets is at its highest as $\Phi_{pm} = B_m A_m$. When the effect of increasing the magnet volume in a radial direction was inspected, there was no change in the air gap flux and flux density as was expected as there was no increase in flux producing area.

When focus was shifted over to embedded magnets one of the main problems was the leakage flux inside both rotors. From the initial simulations and literature it was found that, reducing the leakage flux is one of the key design points of embedded magnet topologies. Leakage flux irrespective of the topology occurs at both magnet ends. For the IRFM topology the magnets were initially placed halfway in the rotor. During the simulations done for the α_{pa} it was found that both the air gap flux density and distribution was influenced by the change in coefficient. The change was similar to that of surface mount topologies, however the difference was that for the IRFM the peak B_{δ} value was also influenced by the change. As the PM volume decreased and the magnets became shorter, the value decreased; this is accorded to the increase in leakage flux inside the rotor. When the magnets become shorter, the material area between two magnets increases providing more space for the leakage. The interesting point of note was that unlike the surface mount topologies where the peak flux density value occurred in the middle of the pole and gradually decreased, with the IRFM the peak flux density value spanned the entire pole with respect to the α_{pa} . The results for α_{mt} were similar to that of the surface mount magnets, both the peak B_{δ} and the distribution are not influenced by a change in radial PM thickness. For the magnetic depth simulations it was found that as the magnets move closer to the surface, B_{δ} increased but the pole flux distribution stayed

the same. The increase in B_δ is accredited to a reduction in leakage flux. The IRFM had similar behaviour to that of the surface mount topologies; the only differences however are the lower peak B_δ and the distribution of the flux density over the pole.

For the ICFM topology, results are different. The first problem that occurred with this topology is the high amount of leakage flux through the shaft; this was eliminated by using a non-magnetic shaft. In doing so B_δ of the reference simulation increased from $\pm 1.15 \text{ T}$ to $\pm 1.4 \text{ T}$. This is a significant increase in air gap flux. What makes it even more impressive is that B_r value of the magnets used in the simulations is 1.25 T , this is due to the flux forcing introduced by the spoke-like placement of the magnets. From the simulation results the pole flux distribution spanned the entire pole regardless of the change in the PM volume and dimensions. For the ICFM topology to be effective, the PMs must span the entire radial length or flux barriers must be used when it is not the case. If this is not done the leakage flux is high, rendering this topology ineffective with regards to PM volume usage. Both the radial thickness and the depth of the magnets if the leakage flux is eliminated had no effect on the air gap flux density and distribution. However reducing α_{pa} increases the peak B_δ over the entire pole. What makes this even more unique is no additional flux is added by the magnets because the thickness of the magnets is increased and not the flux producing area. The increase in B_δ is because the laminated material between the two magnets is reduced thus B_δ has to rise.

Figure 3.26 and Figure 3.27 contains the comparison results of the four reference simulation. In Figure 3.27, the IRFM is by far the lowest; even if the magnets are brought closer to the surface, the air gap flux will still be the lowest.

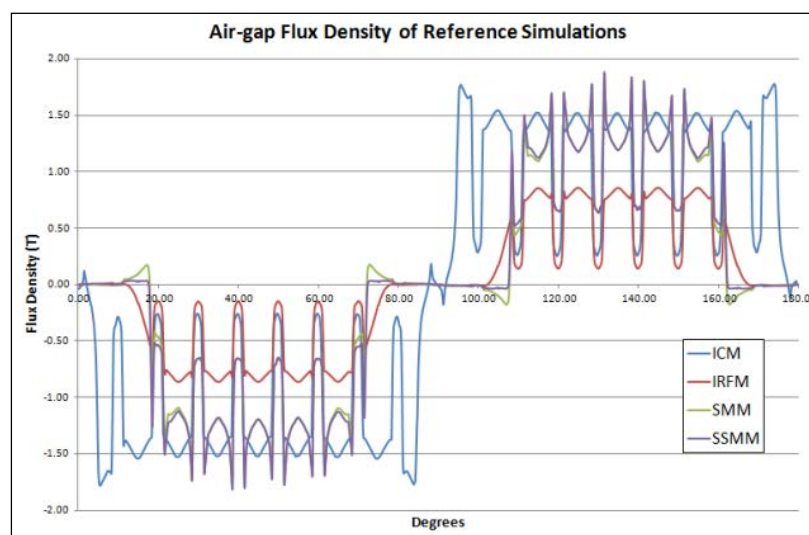


Figure 3.26: Air gap flux of all four topologies [37]

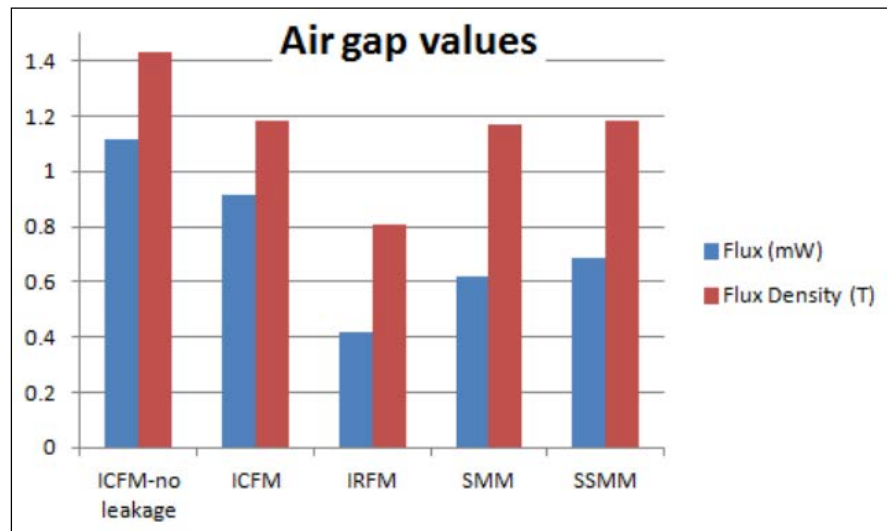


Figure 3.27: Comparison of reference simulation results [37]

The graph shows in Figure 3.27 that the ICFM topology provides the best air gap flux density and distribution performance with regards to PM volume.

It was decided to use the ICFM topology for the prototype rotor. Both surface mount topologies were eliminated on grounds of the poor transient operation capabilities although their steady state performance is slightly better than embedded topologies as listed in Table 3.20. Furthermore, from the literature it was clear the LW LS PMSM machines usually have embedded magnets rotors. The IRFM topology was eliminated on account of the PM volume requirement in comparison with ICFM. The IRFM will need more PM material to provide the same performance.

3.4.1.3 PMSM rotor design

To design the rotor the steps listed in Figure 3.3 in section 3.1.1 will be followed. To aid in following in the process Figure 3.28 is the same as Figure 3.3. In the figure the rotor topology has been selected and an understanding of PM material was gained.

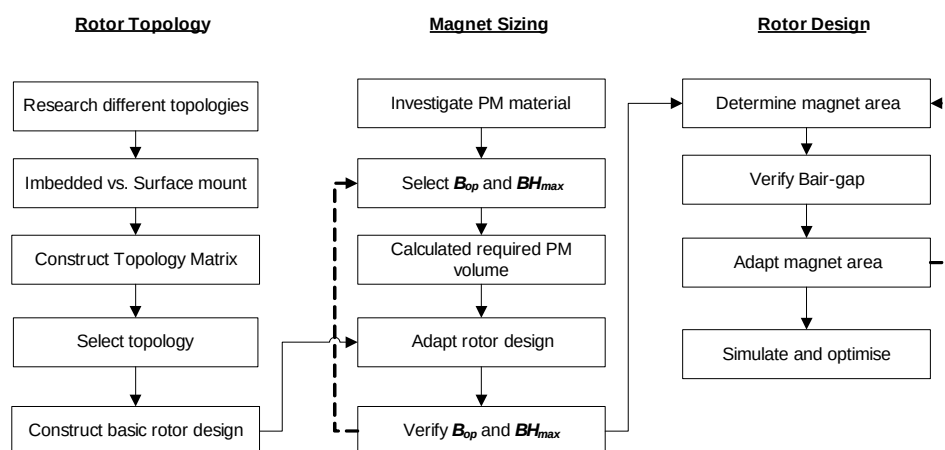


Figure 3.28: PMSM rotor design process

The first step in the rotor design is to define the dimensions of the rotor. As the outside diameter and stack length has already been determined, only the inner rotor diameter must to be specified. This diameter is critical as this is the torque transferring region of the rotor. As the scope of this project does not entail the design of the mechanical components it was decided to use a third party designer (pre-graduate mechanical engineering student) to design and verify the shaft design as this is a critical component in the machine. It was specified that the shaft must be manufactured out of a non-magnetic metal, such as stainless steel and it had to be able to transfer the maximum torque of the machine. The minimum shaft diameter had to be determined and used as the inside rotor diameter (D_{ri}). The shaft dimensions were taken from a 5.5kW IM rotor as all the dimensions regarding the length, bearing dimensions and section spacing is standard except for the D_{ri} section. The shaft manufacturing drawing is included in Appendix F.

To aid in the torque, transfer rotor keys are used; as the machine is symmetrical four keys are used. Information regarding the shaft is listed in Table 3.23. In the table the minimum value for D_{ri} is 38mm, however the inside diameter of the one bearing is 40mm; thus this will be used as the minimum value of D_{ri} . Once the PM and the squirrel cage dimensions have been determined, D_{ri} can be increased to form a flux barrier of saturation bridge between the shaft and PM.

Table 3.23: Rotor shaft information

<i>Material</i>	304 Stainless steel
<i>Torque transfer</i>	4 keys
<i>Min acceptable D_{ri}</i>	38 mm
<i>Bearings</i>	6308 ZZ 6207 ZZ

Although the goal of the design is to minimize the PM in the rotor, the properties of the ICFM topology must be taken into account as well. For the ICFM the PM has to span the entire radial length of the rotor or a non-conducting material needs to be placed between the magnet and air gap. To aid in the IM cage design, the decision was made to place a rotor bar between the magnets and the air gap. If necessary these four bars may be different from the other rotor bars to reduce the leakage flux. To aid in the sizing of the PM, an initial zone is reserved for the rotor bars. This is indicated in Figure 3.29. The orange zone is allocated to the squirrel cage, the dark grey to the shaft and the light grey is the remaining arc.

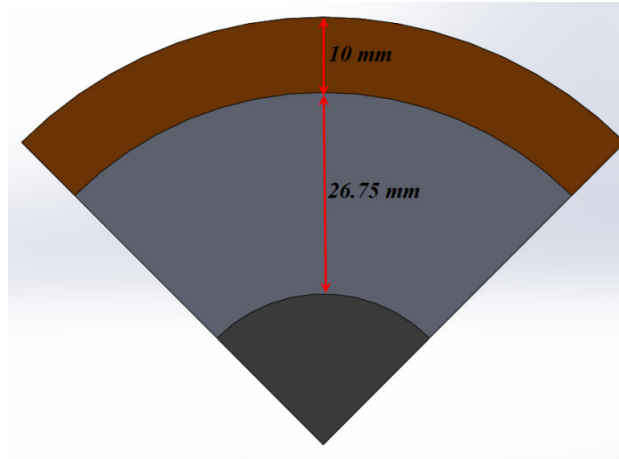


Figure 3.29: Rotor zoning for PMSM and IM

In Appendix D, it is clear from Figure D.3 that B_m must be selected as close as possible to $B_r = 2B_m$ for the PM to provide maximum magnetic energy. From the suppliers' datasheet the B_r values of PM is between 1.04 T and 1.4 T, thus the B_{rmax} value is between 0.52 T and 0.7 T. During the rotor topology study it was found that B_δ of the ICFM is capable of being ± 15 to 25 percent higher than the selected PM grades B_r value. As B_δ selected value is 0.85 T, the initial B_r is selected as 0.65 T. The corresponding PM grade N42 which has a BH_{max} value of 318 kA/m.

Since the length of the air gap and the magnets is the same and the magnet length is limited to 26.75mm, Equation 3.32 b can be further manipulated as

$$\begin{aligned}
 V_{pm} &= \frac{V_\delta B_\delta^2}{\mu_0 (-H_m B_m)} \\
 l_m h_m t_m &= \frac{V_\delta B_\delta^2}{\mu_0 (-BH_{max})} \\
 t_m &= \frac{V_\delta B_\delta^2}{\mu_0 (-BH_{max}) l_m h_m}
 \end{aligned} \tag{3.32}$$

c)

with l_m the length of the rotor, h_m the height of the magnets (26.75 mm) and t_m the thickness. This gave a magnet thickness of 3 mm. Upon further inspection it was found that the knee point of the N42 is in the second quadrant at 20°C and by operating the magnet at the selected point could be problematic as the temperature increased. It was decided to check other PM grades and the magnet thicknesses it provides. Table 3.34 contains this information.

Table 3.24: PM thickness sizing with different grades.

Grade	t_m (mm)
N42	3.1
N40	3.24
N38	3.78
N35	3.9
N33	4.2
N30	4.57

Theoretically all the magnets in Table 3.24 will provide the required air gap flux density at the magnet's BH_{max} value. To determine if it is true, three magnet grades are selected and used in a basic rotor along with the designed stator. Figure 3.30 contains a graph of the three selected magnet grades of the air gap flux.

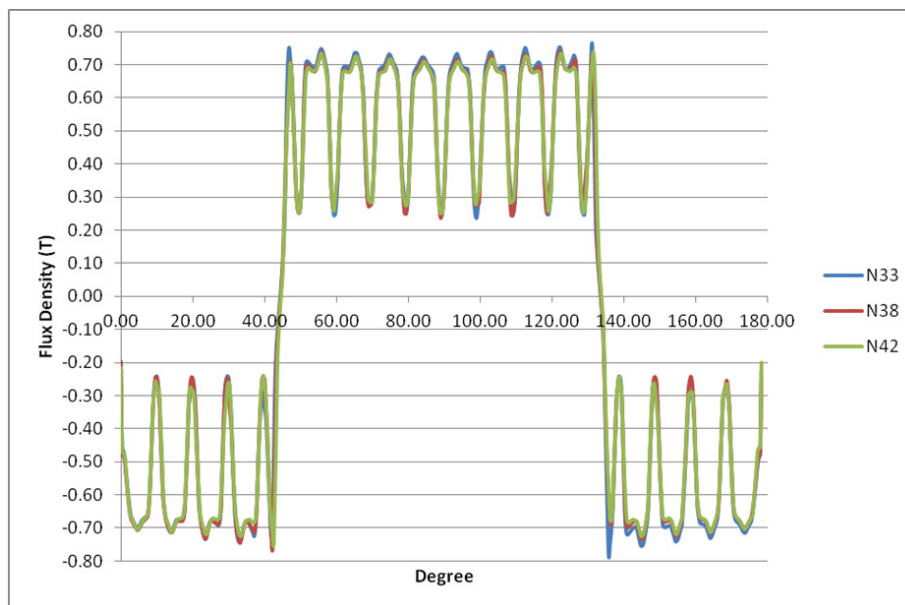


Figure 3.30: Air gap flux density of N42, N38 and N33

From the plot in Figure 3.30, it is clear that all three magnets provide the same air gap flux value at their respective volumes, however this value is 0.15 T below the required value. As the magnets of all grades magnets will provide the same air gap flux density value, it is decided to use the N33 grade PM.

To increase the air gap flux density, α_{pa} has to decrease. The decrease in α_{pa} will increase the thickness of the PM. The magnets' thickness is increases from 4.2mm to 6mm. The increase in thickness increased the air gap flux density value from 0.7 T to ± 0.85 T as indicated in Figure 3.31. The final magnet dimension is 6 x 26 x 115 mm. Note that the air gap flux density plot is in the absence of the stator flux. The air gap flux density component of the stator coils provides a more sinusoidal destitution in the air gap but provides the same average value.

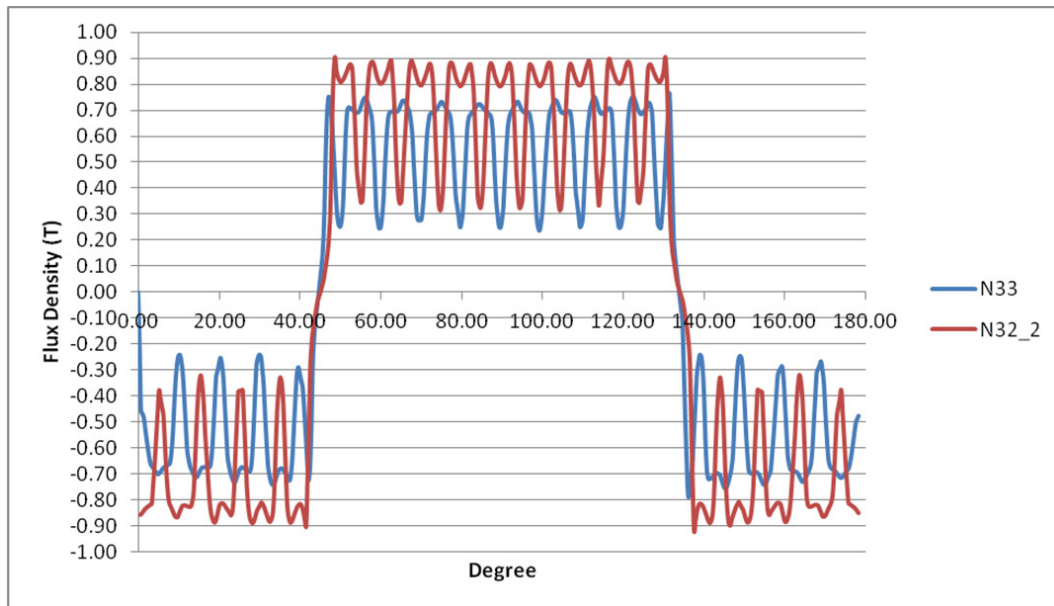


Figure 3.31: Air gap flux density comparison of theoretical value vs. actual value

The next step is to check if the selected grade PM can provide the required magnetic energy at the machines' operating temperature of $\pm 60^\circ\text{C}$. This can be done by adapting Equation 3.32, since the PM volume is known, the energy required to provide the needed air gap flux density can be calculated. The required energy is calculated at 191 kJ/m^3 . By using the magnet energy graph as in Figure 3.32 it is clear that at 20°C the magnet can provide the required energy.

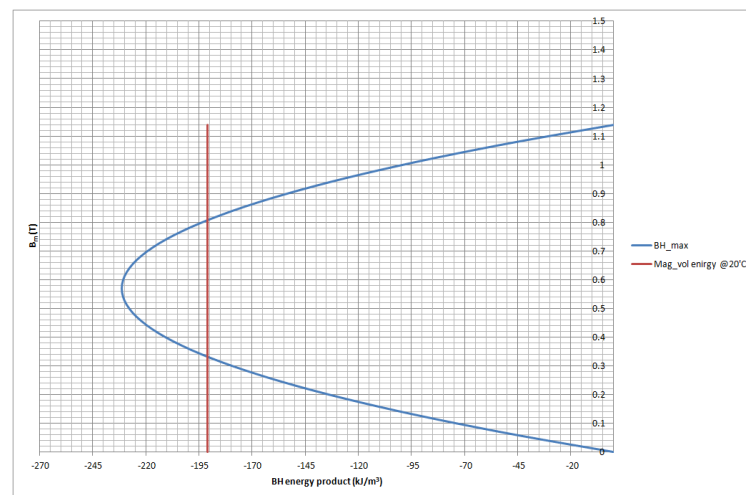


Figure 3.32: PM energy requirements at 20°C

By adapting the PM from 20°C to 60°C the BH_{max} value is decreased; the energy graph is also no longer a pure parabola as the knee point of the demagnetization curve is now in the second quadrant as indicated in Figure 3.33. The knee point is represented on the green line by the change in gradient at around -600 kA/m . The new H_c value of the magnets is -617 kA/m .

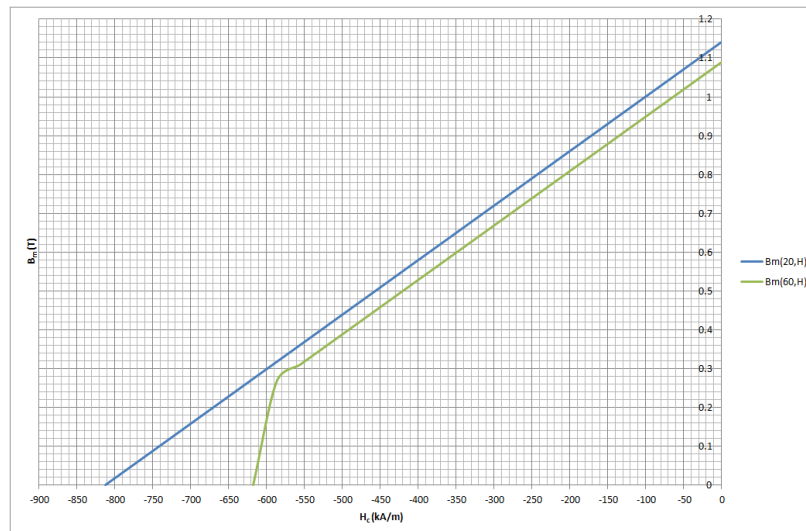


Figure 3.33: PM demagnetisation curve due to temperature increase

The magnetic energy graph for the PM at 60 °C is as in Figure 3.34. From the graph it is still clear that the PM is in its operating range. The PM operating point at this temperature is around 0.7 T which is slightly above the selected value of 0.65 T, however a higher B_m is better as it is further away from the knee point that is at 0.25 T.

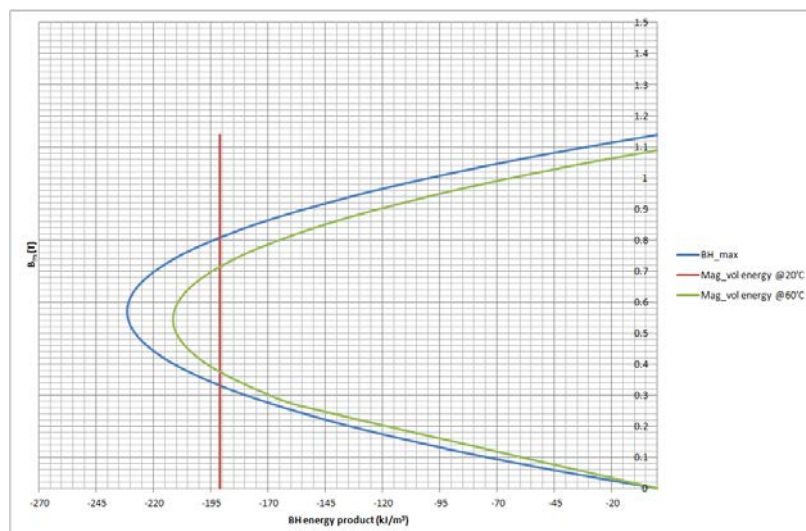


Figure 3.34: BH energy plot at 60 °C of selected PM grade

Now that the magnets have been seized, the next step is to determine the power vs. torque angle and torque vs. torque angle.

To calculate the torque of the PMSM, Equation 2.11 is used and the angled is varied. X_d and X_q is determined with the aid of a FEM package. Both X_d and X_q consist of the relevant magnetising inductance (X_{md} and X_{mq}) and the stator leakage reactance.

$$\begin{aligned} X_d &= X_{s\sigma} + X_{md} \\ X_q &= X_{s\sigma} + X_{mq} \end{aligned} \quad (3.33)$$

X_{md} and X_{mq} can also be calculated by using

$$X_{md} = \omega \mu_0 \frac{2m\tau_p}{p\pi^2 \delta_{def}} l (k_{ws1} N_s)^2 \quad (3.34)$$

with δ_d the effective air gap on the d axis. The same equation is used for X_{mq} but the effective q axis air gap is used [1]. The problem with using Equation 3.34 for embedded topologies is determining the effective axis air gap thus to it is easier to use a FEM package. Once X_d and X_q is known Equation 2.11 is used to plot the results shown in Figure 3.44. The data in Figure 3.35 can be used to plot the power vs. torque angle as in Figure 3.36. The value of X_d , X_q and E_0 is 29.329 Ω , 85.841 Ω and 186.362 V respectively.

$$T_{syn} = \frac{m}{\omega} \left[\frac{V_{ph} E_0}{X_d} \sin \delta + \frac{V_{ph}^2}{2} \left(\frac{1}{X_q} - \frac{1}{X_d} \right) \sin(2\delta) \right] \quad (2.11)$$

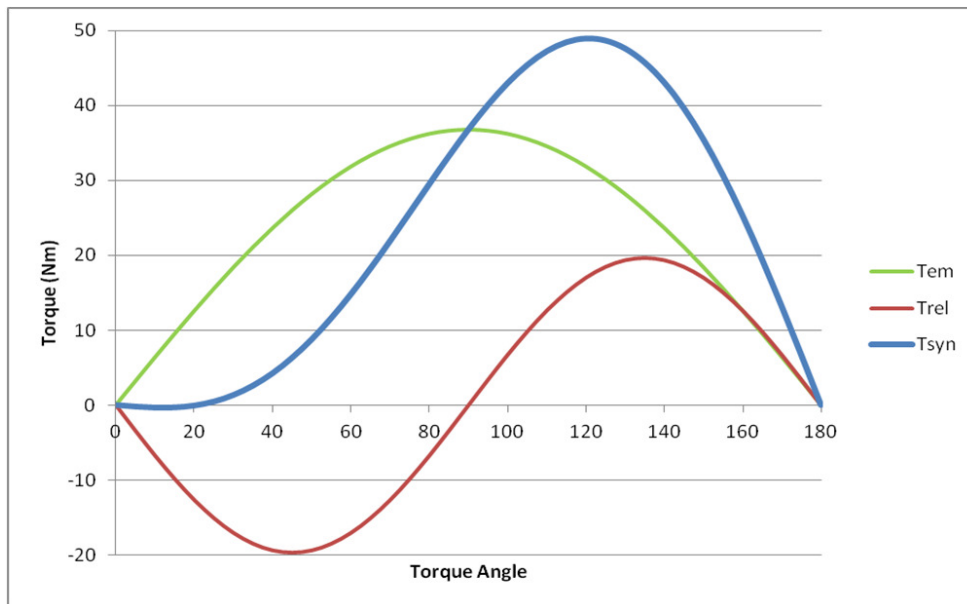


Figure 3.35: PMSM Torque vs. Load angle curve

In Figure 3.35 the two torque components are plotted individually; the green line represents the electromagnetic torque and the red the reluctance torque. The sum of the two torque components provides the synchronous torque line in blue. For a non-salient rotor the maximum torque is at $\delta = 90^\circ$ due to the absence of the reluctance torque (green line) where as an salient pole rotor provided maximum torque at a torque angle less than 90 [1]. However because the ICFM X_q is bigger than X_d the maximum torque is at an angle greater than 90° as seen in Figure 3.36.

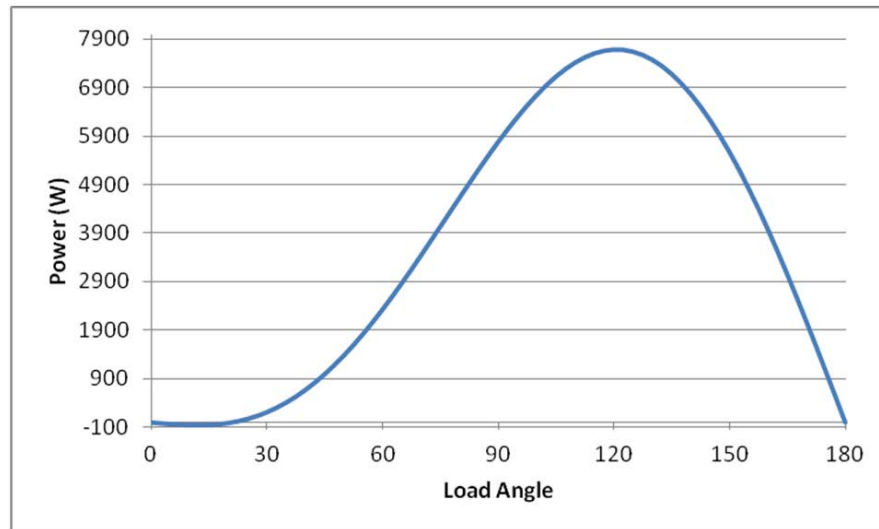


Figure 3.36: PMSM Power vs. Torque angle curve

By using the values of X_d , X_q and E_o in conjunction with the stator resistance value calculated in Section 3.3.3 the braking torque of the PMSM can be calculated. This is done by plotting Equation 2.2 as a function of slip. Figure 3.37 contains the braking torque plot as a function of p.u speed.

$$T_m = \frac{3p}{\omega(1-s)} \left[\frac{R_1^2 + X_q^2(1-s)^2}{R_1^2 + X_q X_d(1-s)^2} \right] \left[\frac{R_1 E_o^2(1-s)^2}{R_1^2 + X_q X_d(1-s)^2} \right] \quad (2.2)$$

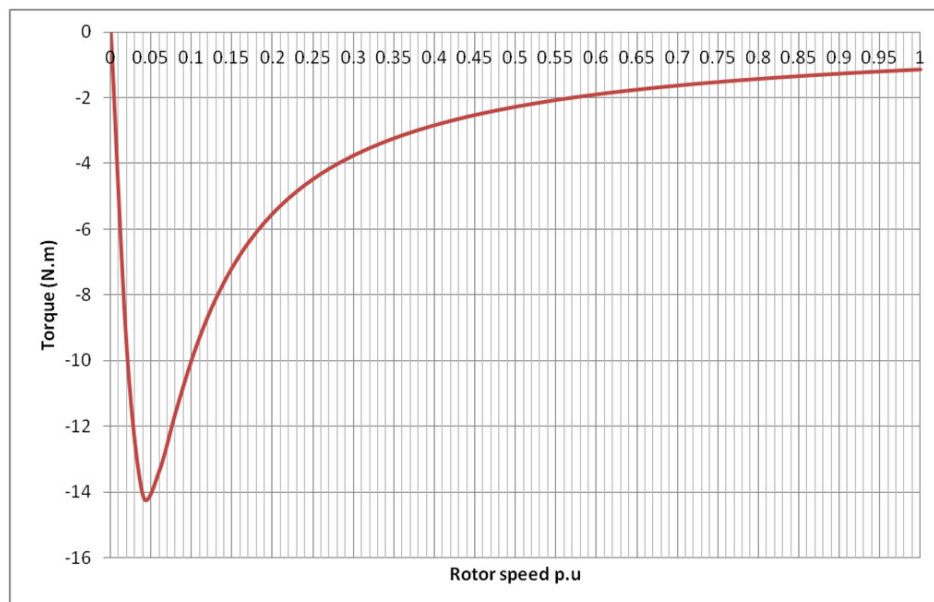


Figure 3.37: Braking torque component of the PMSM in the LS PMSM prototype

From Figure 3.37 the maximum value of T_m is ± -14 N.m and the shape of the plot also correlates with the literature. This information will be used during the design of the rotor cage.

Figure 3.38 is a sectional view of a quarter of the machine. At the air gap end of the magnets a non-magnetic material is placed that spans the area allocated to the cage bars as indicated by the red circle.

Between the shaft end of the magnet and the stainless steel shaft a 0.825mm gap was left to act as a saturation bridge as indicated by the black circle. During the FEMM simulations it was found that less than 5% of the flux passes through this gap. A saturation bridge will also be placed between the air gap end of the magnet and the rotor bar above it. This can only be done once the rotor slot has been designed. The two saturation bridges' goal is to make it possible to manufacture a single lamination instead of four sections. This will benefit assembly and manufacturing.

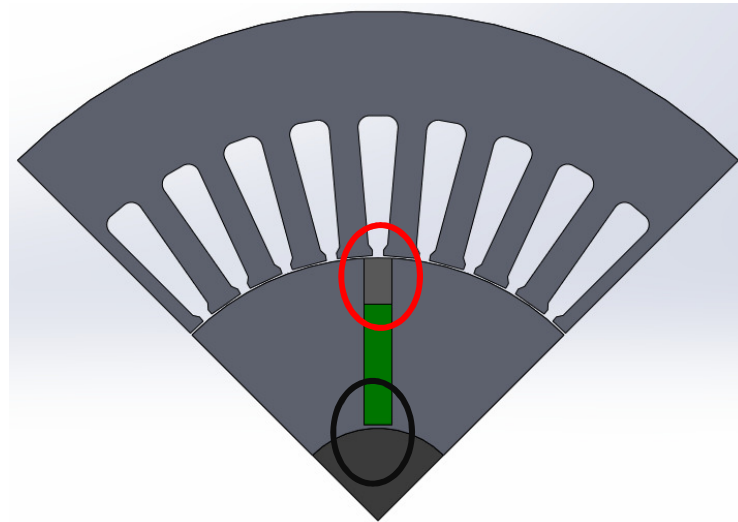


Figure 3.38: Quarter machine section of PMSM

The PMSM design will only be verified once the final rotor is done as the incorporation of the cage will influence the design.

3.4.2 Induction machine design

In Section 3.1 a design flow diagram is provided for the design of the IM cage. Figure 3.39 is a repeat of Figure 3.4 which is the proposed design method as listed in [1]

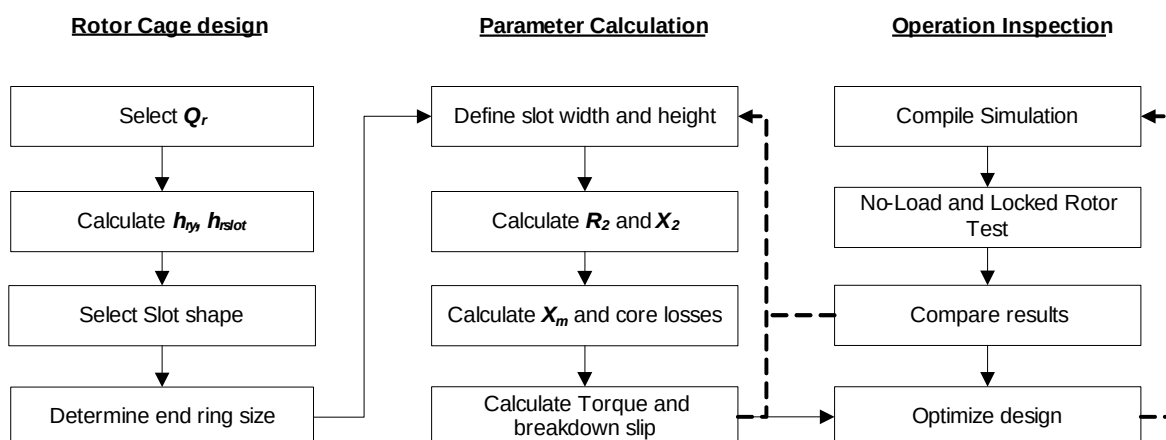


Figure 3.39: Design method of IM cage for LS PMSM

The design consists of three sections. The first step is to design the initial cage, the second step is to calculate the parameters of the equivalent circuit and lastly the design needs to be checked to see if it adheres to the required specifications.

3.4.2.1 Rotor cage design

Before the design of the cage can start, the manufacturing technique of the cage must be selected. Ideally the cage bar and end rings must be cast using a technique known as pressure die casting which is widely used to manufacture small IMs. However, as this is a once-off manufactured prototype, this is a problem as it will increase the manufacturing cost significantly. Thus for the LS PMSM prototype the bars will be manufactured separately and inserted into the rotor slots. The shape of the bar must be as simple as possible not to complicate manufacturing thus increasing manufacturing cost.

The first step in the cage design is to carefully select the number of slots (Q_r) and there have to be selected very carefully. The amount of slots is chosen according to the number of poles, slots per pole and Q_s to limit synchronous parasitic torque [3]. Whether the machine incorporates skewing also plays a role in selecting Q_r . When selecting the number of slots the value must be as low as possible. When selecting the Q_r , there are some guidelines that aid in this decision as in Table 3.25 [1, 3]. g represents any positive integer, p the pole pairs and q_s the number of slots per pole face for the stator which is equal to three for the stator designed in Section 3.2.

Table 3.25: Q_r selection criteria

Slot Harmonics		Mechanical Vibrations	Effects of skewing
$Q_r \neq Q_s$	$Q_r \neq Q_s \pm 2p$	$Q_r \neq 6pg \pm 1$	$Q_r = (6q_s + 4)p$
$Q_r \neq 2Q_s$	$Q_r \neq 2Q_s \pm 2p$	$Q_r \neq 6pg \pm 2p \pm 1$	$Q_r = Q_s + 4p$
$Q_r \neq 0.5Q_s$	$Q_r \neq 0.5Q_s \pm 2p$	$Q_r \neq 6pg \pm 2p \pm 1$	
	$Q_r \neq Q_s \pm p$		

As the rotor is skewed by one slot, the equation in Table 3.25 states that ideally $Q_r = 44$. However in both [1] and [3] tables are provided taking into account all the eliminating factors to aid the selection. The two sources' table is not an exact match but $Q_r = 24, 30$ and 44 overlaps. The LS PMSM prototype relays on symmetrical designs to ensure that a rotor bars is located above each magnets thus $Q_r = 30$ is eliminated. $Q_r = 44$ poses manufacturing problems as the bars will be very small and the number of bars that have to be placed in a pole span of 89 mm is very high, thus 24 rotors slots will be used. This also links with the original statement made in [3] that the least number of bars must be used.

The next step is to design the slot shape, but to do this the minimum rotor yoke height (h_{ry}) and maximum rotor slot height (h_{rt}) must be determined. This is done by selecting the flux density value of the rotor

yoke. In a Table provided in [1], the flux density value for the rotor yoke can be between 1 T to 1.6 T. However the B_r value of the PM for the prototype is 1.17 T and as it spans the entire radial length of h_{ry} , this value will be used [1,3]. h_{ry} is calculated by using the peak average B_δ value obtained from a FEMM simulation which incorporated the stator and PM flux. The peak B_δ is 1.1 T and has an average value of 0.83 T over the pole.

$$h_{ry} = \frac{\frac{B_{\delta peak} \tau_p}{\pi}}{B_{ry}}$$

$$h_{ry} = 26.63 \text{ mm}$$

This value is in line with the PM dimension that spans part of the radial length. The next step is to calculate the maximum slot height [3].

$$h_{rt} = \frac{D_{ro} - D_{shaft}}{2} - h_{ry}$$

$$h_{rt} = 10.11 \text{ mm}$$

The value of the slot height is close to the selected value in Section 3.4.1.4 of 10 mm. For the rotor slot shape there is one of two possibilities. The first option is a round slot and the second a square or rectangle slot. From the two options the round bar will be easiest to manufacture. Thus a round slot will be used to house the cage bars. The short-circuit rings can only be sized once the rotor bars have been sized.

Before the slot width can be calculated the rotor slot pitch τ_{ru} is needed. It is calculated in the same manner as the stator slot pitch in Table 3.8 but only the rotor value are used. τ_{ru} is calculated as 14.85 mm. By using Equation 3.35 the rotor tooth width is calculated

$$b_{rt} \approx \frac{B_{\delta peak}}{B_{rt}} \tau_{ur} \quad (3.35)$$

with B_{rt} the selected rotor tooth flux value [24]. B_{rt} is selected with the aid of information provided in [1] as 2.2 T. This is the maximum value of B_{rt} and occurs on the rotor d-axis. This will gradually decrease between two slots towards the rotor q-axis. B_{rt} is calculated as 7.425 mm. Now that the rotor tooth width is calculated the remaining slot dimensions can be determined for the slot in Figure 3.40.

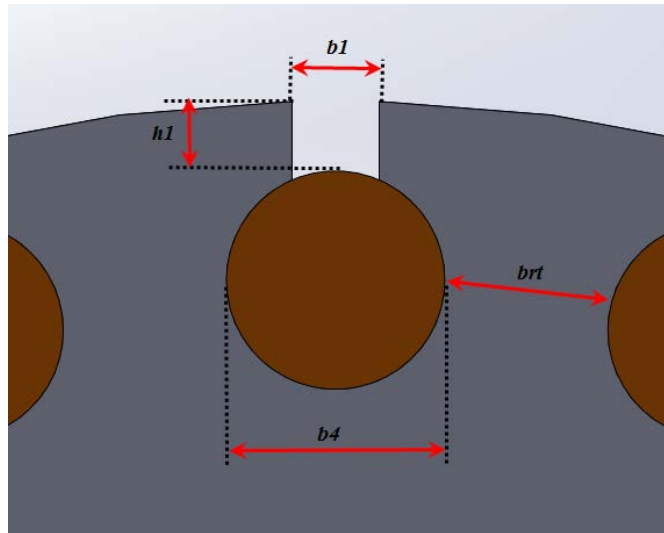


Figure 3.40: Rotor slot dimensions

To calculate the slot performance factor (λ_{ru}) for Figure 3.40 the following equation is used [1]:

$$\lambda_{ru} = 0.47 + 0.66 \frac{b_4}{b_1} + \frac{h_1}{b_1} \quad (3.36)$$

Along with λ_{ru} , the end ring performance factor (λ_{rev}) is used to calculate the rotor leakage reactance L_2 that needs to be referred to the stator reference frame (L_2'). Designing the slot to provide the lowest value for λ_{ru} will in turn improve the performance of the machine. If there is no slot opening λ_{ru} is highly dependent on the space between the slot and the air gap as this now forms a saturation bridge. This also adds additional complications when simulating the machine in a FEM package. b_4 can be calculated by subtracting the tooth width from the slot pitch. The most influential dimension is the slot opening (b_1). Increasing the slot opening area will reduce the leakage flux whereas a deep slot will increase it. The values for b_1 and h_1 are selected as 2 mm and 1 mm respectively which results in $\lambda_{ru} = 3.078$.

Next the end rings must be sized. This is done with the aid of the rotor bar currents and the end ring current after which the required area is calculated. Table 3.26 contains both the required equation and calculated value of the necessary steps up to the calculation of the cross-sectional area of the end ring. The current density in the end ring is selected to be 6.5 A/mm². The currents represented in Table 3.26 are viewed from the stator reference frame.

Table 3.26: Calculating required end ring area [23]

Parameter	Symbol	Equation	Value
Rotor bar current	I_{rb}	$I_{rb} = \frac{2mN_s K_{wl}}{Q_r} I_s \cos \theta$	456.98 A
End ring current	I_{rer}	$I_{rer} = \frac{I_{rb}}{2 \sin \left(\frac{\pi p}{Q_r} \right)}$	882.828 A
End ring area	A_{rer}	$A_{rer} = \frac{I_{rer}}{J_{rer}}$	135.818 mm ²

A general rule with regards to the end ring area is that it must be between 30-50% of the correctional area of the per pole rotor bars [1]. The above calculated area is in this range as well. The final dimensions of the end ring is selected as 12 mm wide and 11 mm high providing a slightly lower A_{rer} value and increasing J_{rer} to 6.68 A/mm². The equation to calculate λ_{rer} is listed in Table 3.27

Table 3.27: Calculating IM rotor parameters [1,3]

Component	Resistance		Inductance		Additional
	Equation	Value	Equation	Value	
End ring	$R_{rer} = \rho_{Al} \frac{l_{ring}}{A_{rer}}$	3.53 ($\mu\Omega$)	$\lambda_{rer} = \frac{l_{ring}}{l'} \ln \left(\frac{l_{ring} Q_r}{2 \sqrt{ab/\pi}} \right)$	0.37007	$l_{ring} \approx \frac{\pi(D_{ro} - b)}{Q_r}$ $l_{ring} = 0.0133m$
Bar	$R_{rb} = \rho_{Al} \frac{l'}{A_{rb}}$	99 ($\mu\Omega$)	$\lambda_{ru} = 0.47 + 0.66 \frac{b_4}{b_1} + \frac{h_1}{b_1}$	3.078	
Total: Rotor reference frame	$R_2 = R_{rb} + \frac{R_{rer}}{2 \sin^2 \left(\frac{2\pi p}{Q_r} \right)}$	106 ($\mu\Omega$)	$X_2 = \mu_0 \omega l' Q_r (\lambda_{rer} + \lambda_{ru})$	78.27 ($\mu\Omega$)	
Total: Stator reference frame	$R_2' = R_2 \frac{Q_r}{3k^2}$	2.187 (Ω)	$X_2' = X_2 \frac{Q_r}{3k^2}$	1.615 (Ω)	$k = \frac{Q_r k_{rskew}}{6N_s k_{wl}}$ $k = 0.01969$

Now that all the dimensions for the IM cage are known the rotor parameters can be calculated. Both the rotor bar and end ring have a resistive and inductive component that forms part of the rotor resistance (R_2) and rotor inductance (X_2). To calculate the different parameters an Excel sheet was constructed. Table 3.27 contains the equations used in the Excel sheet. The calculated value is also added in the table.

With the aid of the machine parameters the starting and breakdown torque can be calculated. Both of these torque values are a function of the starting current. To calculate the starting current the following equation is used [3]:

$$I_{start} = \frac{V_{ph}}{\sqrt{(R_1 + R_2')^2 + (X_1 + X_2')^2}} \quad (3.37)$$

The starting current is calculated as 55.93 A. Table 3.28 contains the equations and calculated values for both the starting torque (T_{start}) and the breakdown torque (T_{bd}) as well as the slip speed at which it occurs at.

Table 3.28: Starting and breakdown torque equations

Parameter	Symbol	Equation	Value
Starting torque	T_{start}	$T_{start} = \frac{3p}{\omega} I_{start}^2 R_2'$	130.67 N.m
Breakdown torque	T_{bd}	$T_{bd} = \frac{3V_{ph}^2}{\omega/p} \frac{R_2'/s_{bd}}{(R_1 + R_2'/s_{bd})^2 + (X_1 + X_2')^2}$	153.59 N.m
T_{bd} slip speed	s_{bd}	$s_{bd} = \frac{R_2'}{\sqrt{R_1^2 + (X_1 + X_2')^2}}$	0.511

By using the breakdown torque equation and performing a parameter sweep of the slip from zero to one the torque vs. slip speed can be plotted. This plot is shown in Figure 3.41

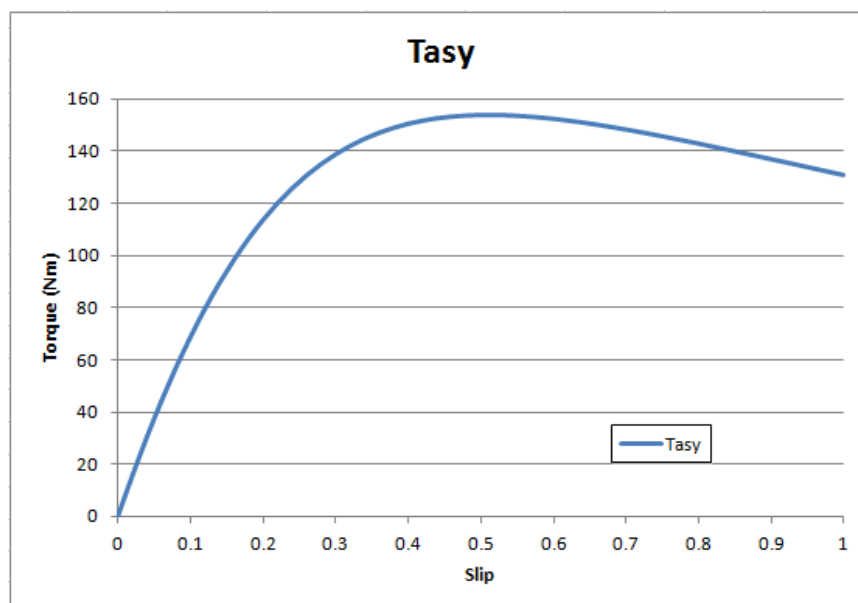


Figure 3.41: Torque vs. Slip speed plot of IM

3.4.2.2 Verifying the IM rotor design

Figure 3.42 is a front view quarter machine section of the IM rotor and the prototype machine's stator. The rotor key is also included in the IM design. This quarter machine section will be used to verify the IM rotor design by checking if the selected flux densities and the FEMM simulation results correlate. Both the rotor yoke and tooth flux densities will be inspected

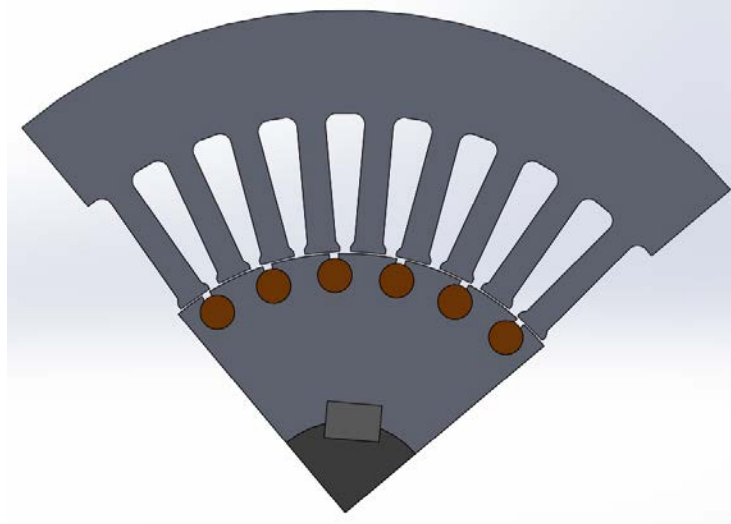


Figure 3.42: Quarter machine section of the IM

Figure 3.43 shows the FEMM simulation of the IM. It also indicates from where the flux density values are extracted. The flux density boundaries set for the plot is [0,1.8]. The air gap flux density value will also be checked to ensure that the flux density values are accurate. Due to the high flux density value in the rotor teeth, two areas are inspected.

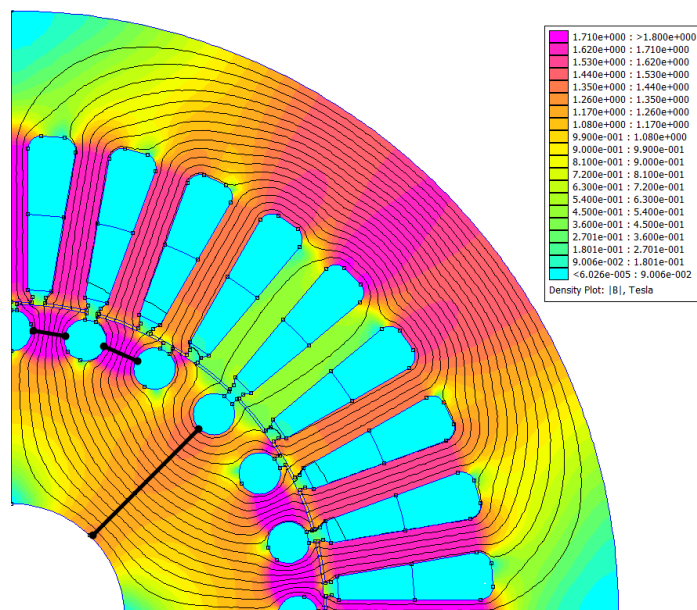


Figure 3.43: Flux density plot in FEMM of the IM.

Table 3.29 contains the selected value, the simulation results and the percentage difference of the rotor yoke and teeth. Both these values need to be within an acceptable limit of 10%. The average air gap flux density is 0.83 T.

Table 3.29: IM rotor verification information

	B_{rt}			B_{ry}		
	Results	Selected	Difference in %	Results	Selected	Difference in %
Tooth 1	2.46 T	2.2 T	11 %	1.306 T	1.17 T	11 %
Tooth 2	2.156 T		2 %			

The two values differ by about 10 %. When the PMs are introduced in the rotor, it will act as a flux density regulator in the back yoke reducing the value of B_{ry} . The PM slots also introduce very small air gaps between the lamination material and PM. Thus before any changes to the design are made the values will be rechecked once the PMSM and the IM are combined.

3.4.3 LS PMSM

Now that both the PMSM and IM rotor components have been designed the next step is to combine the two to form the LS PMSM prototype rotor.

3.4.3.1 Finalising the LS PMSM rotor

To combine the final rotor, the IM cage is placed in such a manner that a rotor bar is situated over each PM in the rotor as shown in Figure 3.44. The gap between the PM and shaft was kept at 0.825 mm; thus the distance between the rotor bar and PM is 1.425 mm.

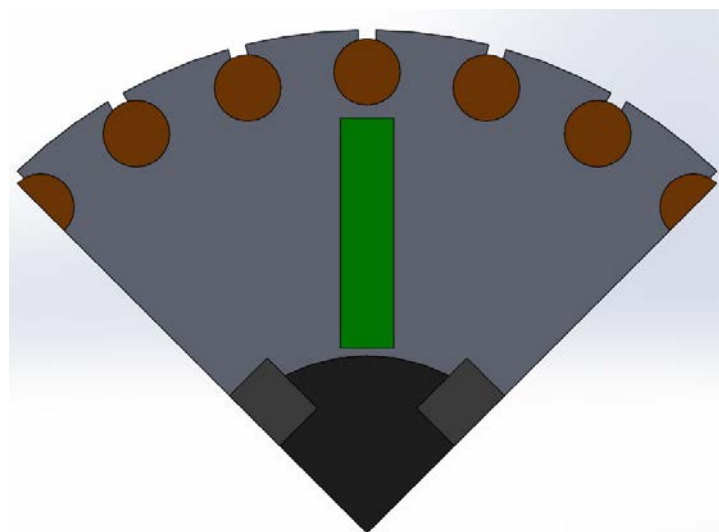


Figure 3.44: Quarter rotor section of LS PMSM

When the design in Figure 3.44 was imported into FEMM along with the designed stator the flux density simulation results are as in Figure 3.45. In (a), only the flux due to the PM are shown and (b) both the PM and stator coils flux are shown. In both simulations the flux density boundaries is set as [1.8,3] with flux line value set at 30.

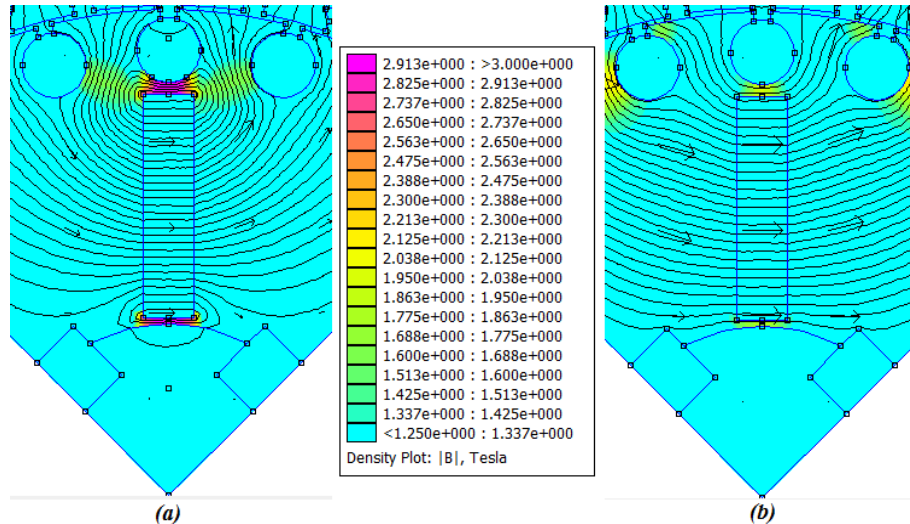


Figure 3.45: FEMM flux density simulation of the first iteration LS PMSM rotor with a) only PM flux b) PM and stator coil flux.

The leakage flux shown in a) between the PM and rotor bar is 18% of the total PM produced flux and in b) 10% of the flux. To reduce the leakage flux in the rotor the space between the PM and rotor bar must be reduced. This can be done by increasing the magnet slot length or by increasing the diameter of the shaft and keeping the distance between the shaft and the PM the same. From the two options the easier solution would be to increase the shaft diameter. By increasing the diameter from 40 mm to 41 mm the gap between the PM and bar is reduced to 0.925 mm. The same two simulations were run and the results are as in Figure 3.46.

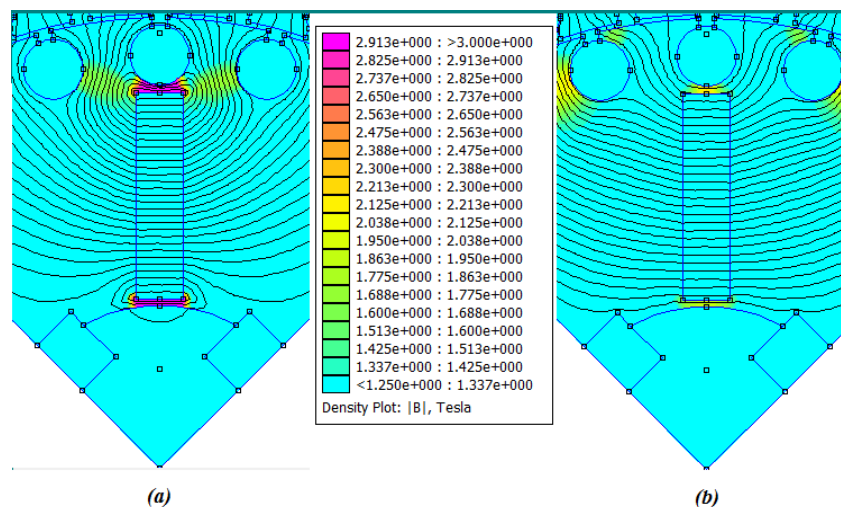


Figure 3.46: FEMM flux density simulation of the second iteration LS PMSM rotor with a) only PM flux b) PM and stator coil flux.

The leakage flux in a) was reduced to 11 % and in b) it was reduced to 5.5 %. By further decreasing the distance between the cage and PM the structure of the rotor will be weakened. The leakage flux when both the PM and stator coil flux is taken into account is relatively low. This will be the same when the rotor is synchronised with the stator field. Figure 3.47 is a CAD image of the final rotor showing all the components of the prototype rotor.

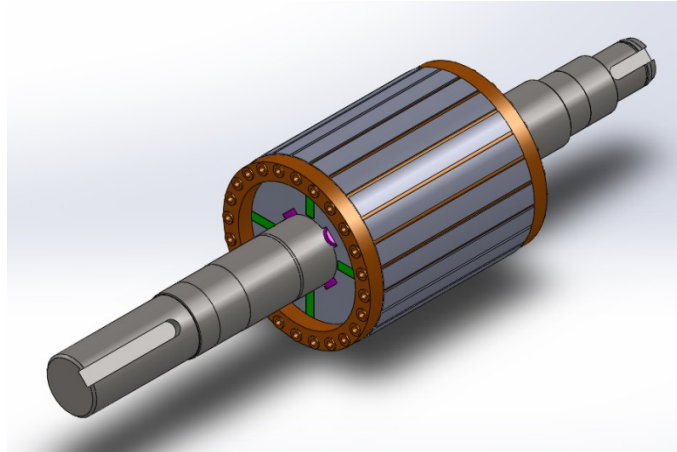


Figure 3.47: CAD representation of prototype LS PMSM rotor

In Figure 3.47 the copper colour represents the IM cage, the purple parts represent the four keys and the green components the four PMs.

3.4.3.2 Verification of LS PMSM rotor

To verify the LS PMSM rotor the air gap flux, rotor yoke and teeth flux density will be used. For the simulation the machine is driven a full load line current ($I_{line} = 10A$) at synchronous speed, Δf between the rotor and stator is zero.

Figure 3.48 shows the flux density plot of the machine with the boundaries set a $[0, 1.8]$. Furthermore the dq -axis of the stator was aligned with that of the rotor in such a manner that the flux direction of the rotor also aligns with the stators flux direction. To verify the LS PMSM rotor the same method is used as with the IM. Table 3.30 contains the verification results and all the values are within the 10 % margin.

Table 3.30: LS PMSM rotor verification information

	B_{rt}			B_{ry}		
	Results	Selected	Difference in %	Results	Selected	Difference in %
Tooth 1	2.29 T	2.2 T	4%	1.208 T	1.17 T	3.2 %
Tooth 2	2.082 T		6%			

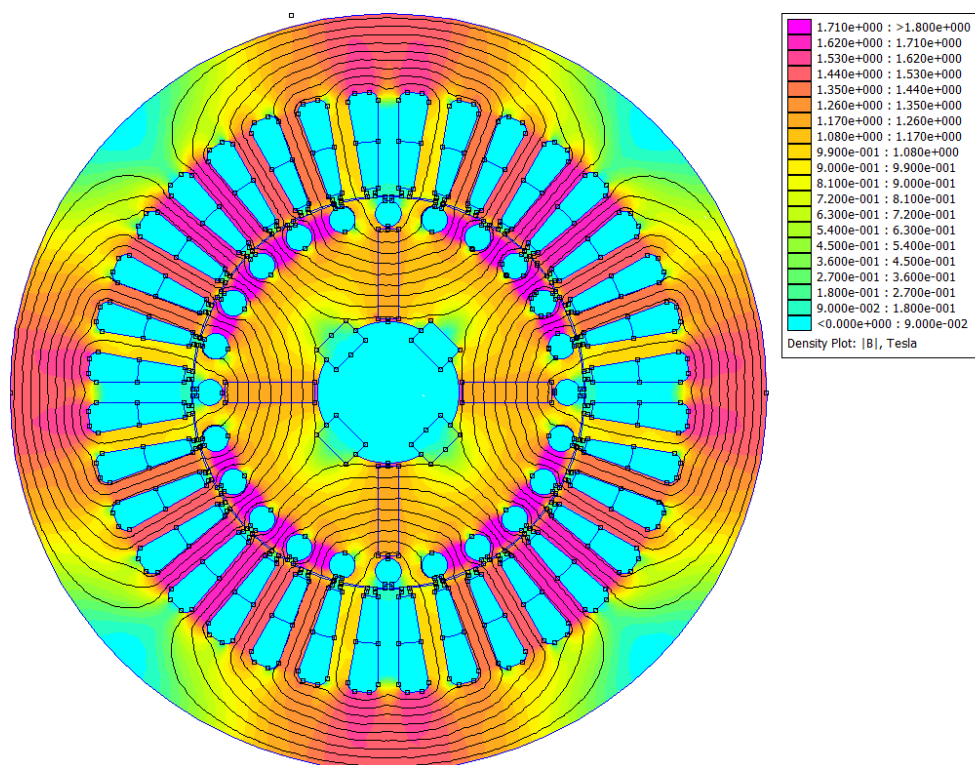


Figure 3.48: FEMM simulation plot of the flux density for the full LS PMSM prototype

In Figure 3.48 it is clear that saturation occurs in areas of the machine, predominately the stator. However it is acceptable for small electrical machines to have some level of saturation in the back yoke and teeth. The losses associated with the saturation are very low in comparison to the increase of machine size to limit the saturation. The saturation zones in the stator are not stationary and revolve with the stator field.

Next, the air gap flux density of the machine is verified. The air gap flux density plot is taken over the pole arch and not the pole pitch (pole arch was defined in Section 3.5). The difference between the two is 6mm. The air gap flux directly above the PM (or q-axis) is assumed to be zero for this simulation setup. The average flux density over the pole arch is calculated as 0.84 T which is still acceptable. Thus the conclusion can be made that the simulation model is an accurate representation model of the LS PMSM machine.

The torque performance and back-emf of the LS PMSM prototype will be inspected in Chapter 4.

In this chapter the detailed design of both the stator and rotor for the LS PMSM prototype was done. The process of design was followed as provided at the beginning of the chapter. This chapter also contained the verification of the stator design with regards to flux densities. The individual rotor components (PMSM & IM) were designed and combined to form the LS PMSM rotor which was also validated using FEMM to determine if the selected flux densities were achieved.