

References

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Appendix A: Orthogonal (dq) Axes.

The orthogonal axis consists of two vectors that are separated by an angle of 90° . The two quantities are an equal representation of $m = 3, 6, 9 \dots$ phase symmetrical windings. They are known as the direct and quadrature axis. Thus a multiple phase electrical machine can be reduced to a 2 base vector equivalent [2, 3].

The dq -transform is used in machine design to analyse the interaction of the stator and rotor in both salient and non-salient pole machines. With the aid of the transform, steady state stator flux and mmf-waves can be represented as equivalent quantities that rotate with the rotor at a constant spatial angle. This is similar to time-varying inductances of the different phases being transformed to rotate with the rotor so constant magnetic linkage are seen between the rotor and stator [2]. The dq -axis can also be used to model the transient response of a machine to estimate the electrical machine parameters [3].

The most commonly used electrical machines have a 3-phase stator so this discussion of the dq -transformation will only focus on this configuration.

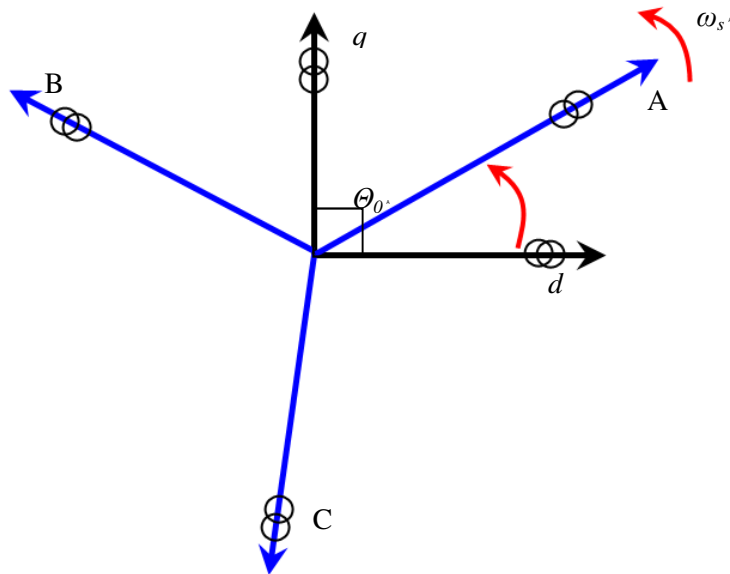


Figure A.1: 3-phase AC dq equivalence

In Figure A.1, if the quantity A, B and C represents the line current of a machine, i_A , i_B and i_C , then i_d and i_q represents the transformed current values. To transform the ABC to dq quantities the generalized Park transform is used:

$$\begin{bmatrix} d \\ q \\ \vec{0} \end{bmatrix} = [S_{park}] \begin{bmatrix} A \\ B \\ C \end{bmatrix} \quad (\text{A.1})$$



with

$$[S_{Park}] = \frac{2}{3} \begin{bmatrix} \cos(\theta_s) & \cos(\theta_s - 120^\circ) & \cos(\theta_s + 120^\circ) \\ -\sin(\theta_s) & -\sin(\theta_s - 120^\circ) & -\sin(\theta_s + 120^\circ) \\ 0.5 & 0.5 & 0.5 \end{bmatrix} \quad (A.2)$$

and

$$\theta_s = \omega_s t + \theta_0 \quad (A.3)$$

In Equation A.1, a zero order component is required as the transform is produced as a 3 by 3 matrix calculation. This is not a problem since the ABC quantities are equal because the electrical system is balanced. Thus if Equation 2.11 is calculated for any balanced system the zero order component is

$$\vec{0} = \frac{A+B+C}{3} \quad (A.4)$$

but

$$C = -B$$

resulting in

$$\vec{0} = \frac{A+B-C}{3} = 0$$

To prove that the dq transform is a representation of a balanced 3-phase machine the power of the machine is calculated by with the aid of the inverse Park matrix

$$[dq]^{-1} = \frac{3}{2} [dq]^t \quad (A.5)$$

The machine power is thus calculated by

$$\begin{aligned} P_{dq} &= [V_{dq}]^t [I_{dq}] \\ P_{dq} &= V_d I_d + V_q I_q + V_0 I_0 \\ P_{dq} &= V_d I_d + V_q I_q \end{aligned}$$

$$\frac{3}{2} P_{dq} = \frac{3}{2} [V_{ABC}]^t [S_{Park}]^t [S_{Park}] [I_{ABC}]$$

and

$$\begin{aligned} P_{dq} &= [V_{ABC}]^t [I_{ABC}] = P_{ABC} \\ P_{dq} &= P_{ABC} = V_A I_A + V_B I_B + V_C I_C \end{aligned}$$

thus

$$V_d I_d + V_q I_q = V_A I_A + V_B I_B + V_C I_C$$

To aid in the understanding of the dq transform, a physical machine model can also be used. Although each machine has its own mathematical and physical dq model the basic principal and indication of the

dq -axis inside the machine is the same. These axes can be identified with the aid of flux density lines or machine poles.

The number of dq -axes will equal the number of poles. Figure A.2 below is several machine sectional views. The magnetic flux lines aid in indicating the position of the dq -axis in the machine.

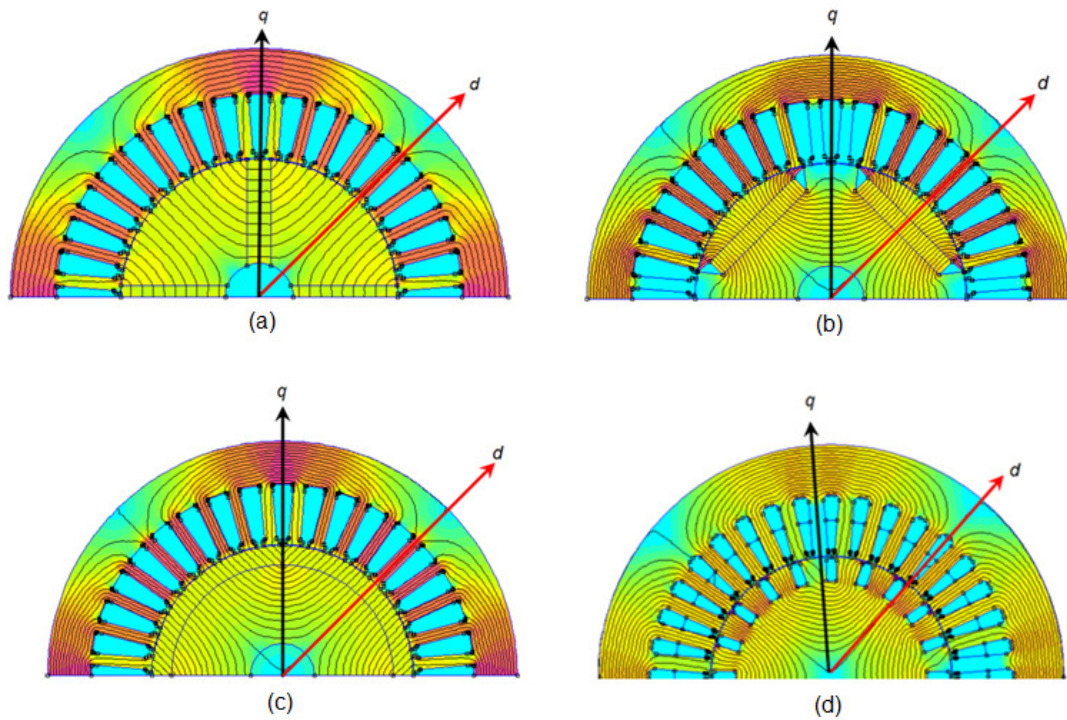


Figure A.2: dq axes in a 4 pole a) tangential flux PMSM, b) radial flux PMSM, c) surface mount PMSM, d) IM

The dq -transform can be performed with the analytical and FEM results to construct a dynamic model that is an accurate representation of the machine behaviour in transient and steady state operation. This technique can be used to model a wide variety of electrical machines, from DC brush PM motors to IM and incorporates both magnetic saturation and skin effect that occurs in the machine [3].

Appendix B: IEC machine frame sizes.

The IEC developed international standards regarding the dimensions and the way in which motors can be mounted. These standards are applicable to all industrial manufactured machines. A frame size is represented by a code, for example 90L –IP 55. Each component provides information regarding a specific sizing parameter. The 90 represents the height of the motor from the middle point to the mounting plate/bracket. The L represents the axial length of the machine. With regards to the machine length there are three options, short (S), medium (M) and long (L). IP 55 represents the ingress protection of the motor. A machine's ingress protection (IP) rating indicates the degree of protection against the capability of foreign objects entering the machine. Most LV machines have an IP rating of IP 55. The first integer states that the machine is protected against the ingress of dust and the second integer states that the machine will not let water from a water jet with a 6.3mm nozzle enter from any direction.

Industry manufactured machines are manufactured according to fixed frame sizes. All induction machines of the same kW rating use the same frame size regardless of the manufacturer; the same idea is applied to PMSMs and other machines with different frame sizes as the size differs due to the operating principal. Machines can be mounted in two different ways. Figure B.1 indicates the two mounting types.

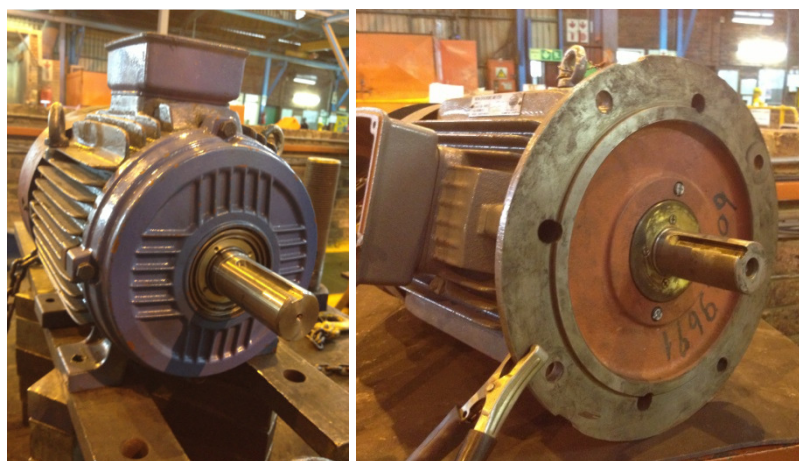


Figure B.1: a) Foot or base mounted. b) Flange mounted

Figure B.2 represents a standard IEC frame with the terminal box mounted on the right handed side. The terminal box can also be on the top or the left hand side depending on the installation space. This particulate machine is foot mounted machine. For a flange mounted machine the dimensions are the same.

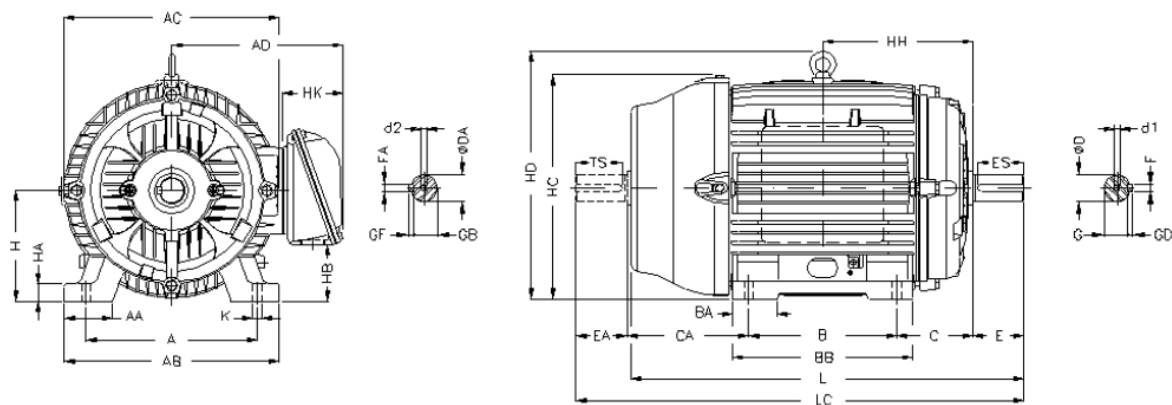


Figure B.2: IEC frame dimensions

Frame size	A	B	C	D	E	H
71	112	90	45	14	30	71
80	125	100	50	19	40	80
90S	140	100	56	24	50	90
90L	140	125	56	24	50	90
100L	160	140	63	28	60	100
112M	190	140	70	28	60	112
132S	216	140	89	38	80	132
132M	216	178	89	38	80	132
160M	254	210	108	42	110	160
160L	254	254	108	42	110	160
180M	279	241	121	48	110	180
180L	279	279	121	48	110	180
200M	318	267	133	55	110	200
225M	356	286	149	55	110	225
250	406	311	168	60	140	250
280	457	457	368	190	140	280

Appendix C: Stator winding factors

The mmf distribution in the air gap due to the winding is not calculated directly with the number of coil turns as the winding configuration interacts with a certain phase shift to the stator surface. Depending on the configuration the flux in the machine does not interact with all the windings in the same manner at a specific time. Thus to calculate the mmf the winding factor is incorporated in the calculation thereof. To calculate the k_{wv} (winding factor) of the stator, the product of k_{dv} (distribution factor), k_{pv} (pitch factor) and k_{sqv} (skewing factor) must be calculated, thus

$$k_{wv} = k_{dv} k_{pv} k_{sqv} \quad (C.1)$$

respectively. Each of the three factors will be discussed individually before the actual design is done. The effect of incorporating skewing in the machine is added in the winding factor as it has a weakening effect on the induced voltages in the rotor bars.

C.1 Distribution factor - k_d

The distribution factor of a stator can be defined as the factor by which the phase voltage is reduced due to the increase of slots per phase zone (q).

If there is no coil distribution ($q = 1$) then k_{dv} will be 1 and will have no influence on k_{wv} as all the coil turns are installed in one slot. This however is not recommended as the needed slot area will be very big as the slot fill factor will also need to increase. This will in turn add other harmonic component problems to the machine at a later stage [3]. Figure C.1 below is a visual representation of the **physical** influence q has on the stator when the value of q is increase.

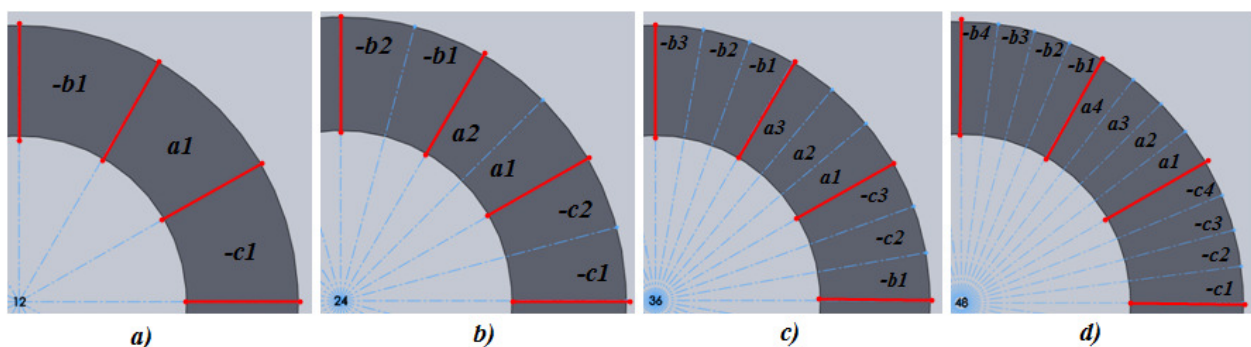


Figure C.1: Stator variations when a) $q=1$ then $Q = 12$, b) $q=2$ then $Q = 24$, c) $q=3$ then $Q = 36$, d) $q=4$ then $Q = 48$

Figure C.1, represents a single pole pitch region of the stator; the area between two red lines represents a phase zone area. The area between any two lines represents the area allocated to a single slot. The blue lines are used to indicate the increase in slot per phase zones as q increases.

$$k_{d1} = \frac{\text{geometric sum}}{\text{sum of absolute values}} = \frac{U_1}{qU_{coil}} \quad (\text{C.2})$$
$$\begin{aligned}\sin \frac{q\tau_u}{2} &= \frac{U_1/2}{r} \\ U_1 &= 2r \sin \frac{q\tau_u}{2}\end{aligned}\tag{C.3}$$
$$\sin \frac{\tau_u}{2} \frac{U_{v1}}{r} \quad (C.4)$$

$$U_1 = 2r \sin \frac{\tau_u}{2}$$
$$k_{d1} = \frac{U_1}{qU_{coil}} = \frac{2r \sin \frac{q\tau_u}{2}}{q2r \sin \frac{\tau_u}{2}} = \frac{\sin \frac{q\tau_u}{2}}{q \sin \frac{\tau_u}{2}}$$
$$k_{d1} = \frac{2 \sin \frac{\pi}{2m}}{\frac{Q_s}{mp} \sin \frac{\pi p}{Q_s}} \quad (\text{C.5})$$

since $Q_s = 2p q m$ and $\tau_u = 2p\pi / Q_s$. To determine the v^{th} harmonic component influence of the distribution factor on the air gap flux density the $v\tau_u$ angle is applied thus

$$k_{dv} = \frac{2 \sin\left(v \frac{\pi}{2m}\right)}{\frac{Q_s}{mp} \sin\left(v \frac{\pi p}{Q_s}\right)} \quad (C.6)$$

The value of k_{dv} is now dependent on the electrical phases, the amount of pole pairs and the slot count of the stator. As both the phase and pole pair is fixed ($m = 3, p = 2$) k_{dv} for the LS PMSM prototype is only dependent on the slot count. Figure C.3 below contains the harmonic plot for the three selected values of q

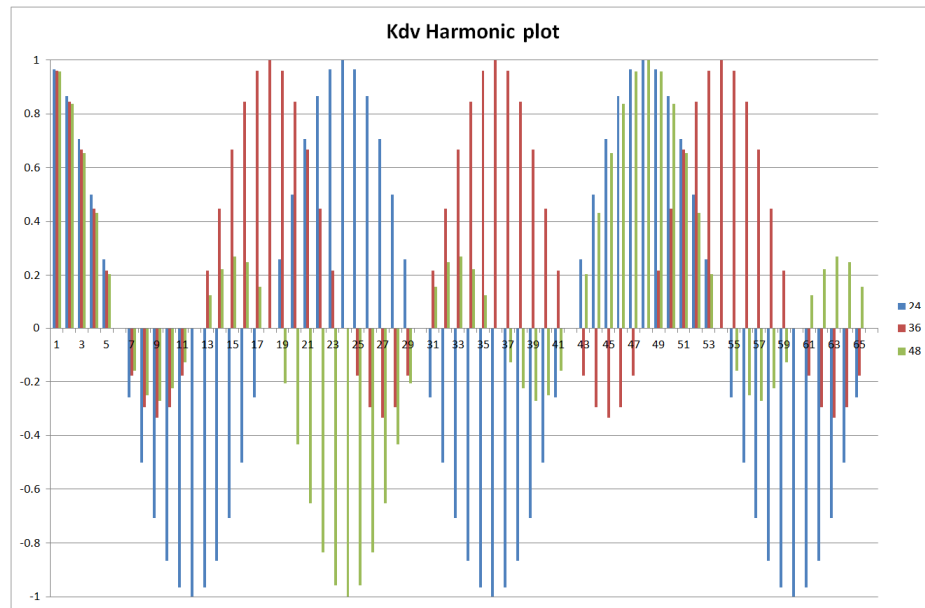


Figure C.3: Harmonic plots of k_{dv} for various values of q

Table C.1 contains the fundamental values for the various values of q .

Table C.1: k_{d1} for a 24, 36 and 48 slot stator

Q_s	q	k_{d1}
24	2	0.9659
36	3	0.9597
48	4	0.9610

C.2 Pitch factor - k_p

The pitch factor of a stator can be defined as the factor by which the phase voltage is reduced due to short pitching the coil. If there is no short pitching ($y = y_{Qs}$) then k_{pv} will be 1 and will have no influence on k_{wv} as the coil spans the entire pole pitch.

A phase coil is short pitch when it does not span the entire pole pitch. Short pitching is done in multiples of slot pitches (τ_u). The actual coil side (starting point of the coil) is placed in the inner part of the slot and the short pitched winding from a different coil side (end point of the coil) is placed in the outer part of the slot. Figure C.4 indicates the differences in a double layer configuration vs a double layer that is short pitched by one coil slot.

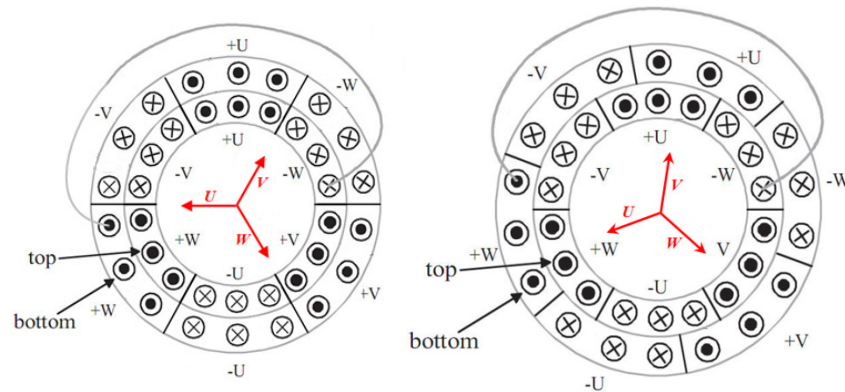


Figure C.4: Stator pitching configuration a) no short pitching b) short pitching by one slot [1]

The stator in Figure C.4 is that of a three phase, two pole, 18 slot machine. The configuration for a four-pole machine is different but the concept behind short pitching is the same. The main advantage short pitching has is the reduction in copper wire; because of the shorter coil span, less copper is used to wind the machine as the end windings are shorter. This will lead to lower stator copper losses. Further, when short pitching is used correctly a more sinusoidal current linkage waveform is produced than that of a full pitched winding. However short pitching has a disadvantage as well: the smaller coil area reduces the flux linkage, but this can be overcome by increasing the number of turns in the coil. The material saving due to short pitching is higher than that of the winding increase, thus the end result is lower copper losses.

Figure C.5 represents a two pole stator that is short pitched by one slot. The dashed line indicates the calculated coil pitch (y) from slot 1 and the black arrow the actual coil span (y_{Qs}).

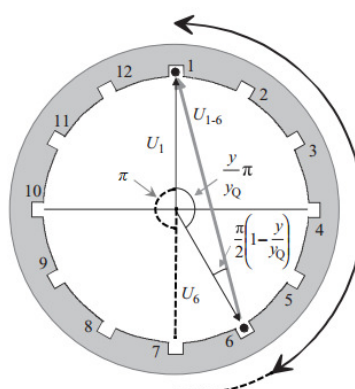


Figure C.5: Pitch factor determination [1]

The calculated coil span can be expressed in radians as π and the actual coil span as $(y/y_{Qs})\pi$ which takes into account the reduction ratio of the coil. From Figure C.5 it is clear that $(y/y_{Qs})\pi$ is complimentary to π

thus the difference between the calculated and actual coil spans angles is the sum of the remaining angles of an triangle. As the triangle has two equal sides the angles are equal as well.

The pitch factor is the ratio between the geometric sum of the phasors and an absolute voltage phasor [1].

$$k_p = \frac{U_{total}}{2U_{slot}} \quad (C.7)$$

Thus

$$\cos\left(\frac{\pi}{2}\left(1 - \frac{y}{y_{Q_s}}\right)\right) = \frac{U_{total}/2}{U_{slot}} = \frac{U_{total}}{2U_{slot}} = k_p$$

and can be transformed to

$$k_{p1} = \sin\left(\frac{y}{y_{Q_s}} \frac{\pi}{2}\right) \quad (C.8)$$

To determine the v^{th} harmonic component of Equation C.7 it is rewritten as

$$k_{pv} = \sin\left(v \frac{y}{y_{Q_s}} \frac{\pi}{2}\right) \quad (C.9)$$

y/y_{Q_s} in Equation C.9 represents the ration of actual coil span of the calculated coil span in amount of slots. For example the values of y and y_{Q_s} of a four-pole, 36 slot stator that is short pitched by one slot, is 8 and 9 respectively. Figure C.6 represents the harmonic plots for a single short pitched winding for a 24, 36 and 48 slot stator.

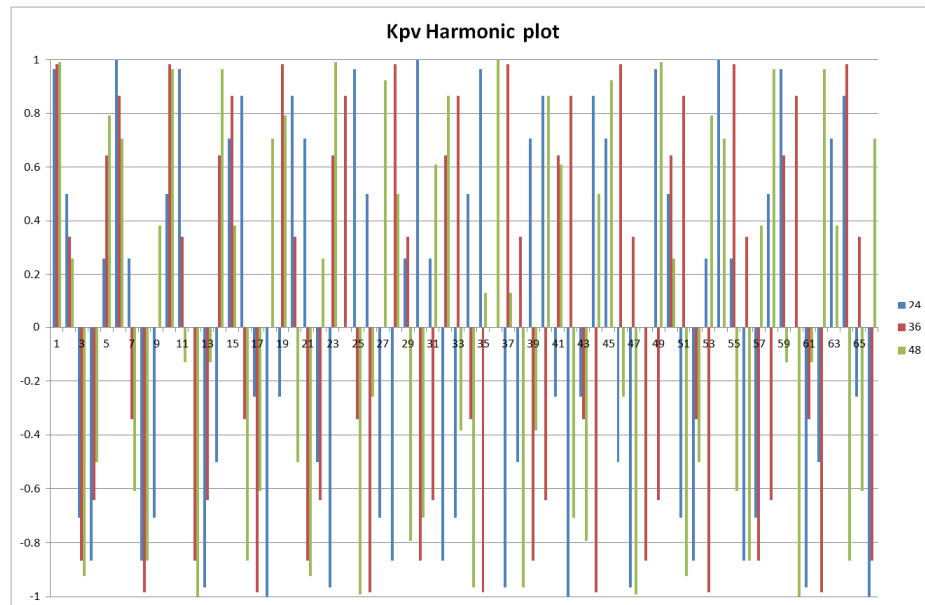


Figure C.6: Harmonic plots of k_{pv} for various values of q

The main difference between the 3 plots is the periodical increase in the harmonic component as the v^{th} increase. Table C.2 contains the fundamental pitch factor values for the different stator designs in the figure.

Table C.2: k_{p1} for a 24, 36 and 48 slot stator

Q_s	y	y_{Qs}	K_{p1}
24	5	6	0.9569
36	8	9	0.9848
48	11	12	0.9914

C.3 Skewing factor - k_{sq}

Slot skewing is measured by the angle between the rotor and the stator slots, the degree to which the rotor and stator slots are skew affects and reduces the harmful harmonics in the machine. The skewing factor is derived with the aid of the distribution factor. In Figure C.6 a skewed bar is divided into several shorter bars with the angle between two bars defined as $\Delta\alpha$.

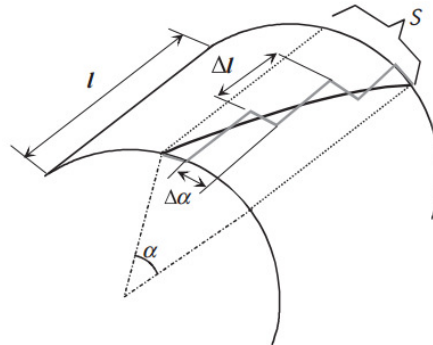


Figure C.7: Determination of the skew of a rotor bar [1]

By dividing a bar into $\Delta\alpha$ sections the emf of a bar is calculated in the same manner as a coil, thus q can be replaced by $\alpha/\Delta\alpha$ and τ_u with $\Delta\alpha$. If the limit of $\Delta\alpha$ approaching zero is calculated the skewing factor is defined as:

$$k_{sqv} = \lim_{\Delta\alpha \rightarrow 0} \frac{\sin\left(v \frac{\alpha}{\Delta\alpha} \frac{\Delta\alpha}{2}\right)}{\frac{\alpha}{\Delta\alpha} \sin\left(v \frac{\Delta\alpha}{\alpha}\right)} \quad (C.10)$$

$$k_{sqv} = \frac{\sin\left(v \frac{\alpha}{2}\right)}{\frac{\alpha}{\Delta\alpha} v \frac{\Delta\alpha}{2}} = \frac{\sin\left(v \frac{\alpha}{2}\right)}{v \frac{\alpha}{2}}$$

with $\alpha = s\pi/\tau_p$ [1]. By substituting α in Equation C.9

$$k_{sqv} = \frac{\sin\left(v \frac{s}{\tau_p} \frac{\pi}{2}\right)}{v \frac{s}{\tau_p} \frac{\pi}{2}} \quad (C.11)$$

In [1] it is proven that by skewing a rotor/stator slot by one slot pitch the harmonics are effectively cancelled. This entitles that s/τ_p can be replaced with $1/mq$ [1].

$$k_{sqv} = \frac{\sin\left(v \frac{\pi}{2} \frac{1}{mq}\right)}{v \frac{\pi}{2} \frac{1}{mq}} \quad (C.12)$$

Figure C.8 contains the harmonic winding plot of a single layer 36 slot stators with one slot skewing (blue bars) vs no skewing (red bars). This is a clear indication that incorporating skewing in the machine design reduces v^{th} harmonics components.

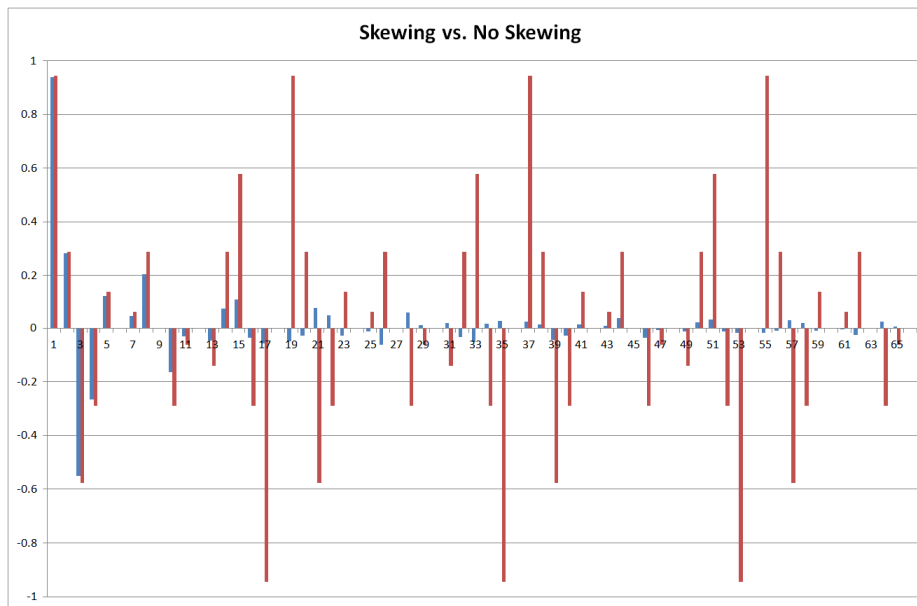


Figure C.8: Skewed vs. un-skewed machine

By using Equation C.12 for the different values of q (2, 3, 4) the harmonic skewing factor plot was generated and is depicted in Figure C.15

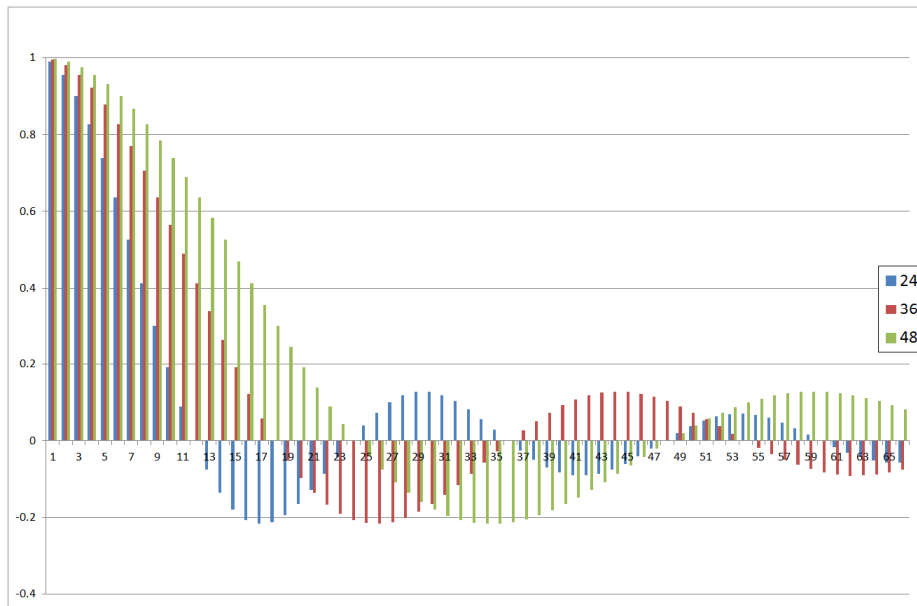


Figure C.9: Harmonic plots of k_{sqv} for various values of q

Table C.3 contains the fundamental components that will be used in calculating the various winding factors.

Table C.3: k_{sq1} for a 24, 36 and 48 slot stator

Q_s	q	k_{sq1}
24	2	0.9886
36	3	0.9949
48	4	0.9971

Appendix D: Permanent Magnet Background

In Section 3.4.1.1 the PMs for the LS PMSM is sized. Good understandings of PMs are needed to do this. Over the years PM manufactured from different material was developed or discovered all with different application possibilities. PM materials that are used in electrical machines will be discussed in this appendix. The most commonly used PM material in electrical machines are Alnico, ceramics and rare-earth materials [27].

D.1 Permanent magnet material

All PMs are specified by manufacturers according to their remanance magnetization (B_r), coercicity (H_c) and maximum magnetic energy product (BH_{max}). Other factors also include temperature coefficients effecting B_r and H_c , Curie temperature, material resistivity (ρ) and the second quarter hysteresis curve form [1, 2, 5]. Table D.1 contains information regarding the relevant magnetic materials.

Table D.1: Different PM material properties [1, 5, 27, 28]

	Alnico	Ceramic	Rare-earth
Discovered	1931	1960's	1983
Composition	Al, Ni, Co, Fe	Sm, Co, Fe, Cu	Nd, Fe, B
BH_{max}	30 to 110 kJ/m ³	130 to 255 kJ/m ³	200 – 310 kJ/m ³
B_r	0.7 - 1.2 T	0.82 – 1.16 T	1.03 – 1.3 T
H_c	50 kA/m to 150 kA/m	493 kA/m to 1.59 MA/m	875 kA/m to 1.99 MA/m
B_r reverse temperature coefficient	-0.02 %/K	-0.3 to -0.4 %/K	-0.11 to -0.13 %/K
H_c reverse temperature coefficient	-	-0.15 to -0.30 %/K	-0.55 to -0.65 %/K
Curie temperature.	520 °C	800 °C	320 °C
Resistivity	450 nΩm	860 nΩm	1100-1700 nΩm

The following can be derived from Table D.1: Alnico magnets have a high B_r value that is not as sensitive to temperature changes. However its H_c is very low, thus resulting in a low BH_{max} value. Alnico magnets have a poor non-linear demagnetization curve thus the magnets are easily magnetized but just as easily demagnetized.

Ceramic magnets have a much higher H_c value than Alnico but a slightly lower B_r . This difference leads to ceramic magnets having a much larger BH_{max} . However ceramic magnets have much higher reverse temperature coefficients. The biggest advantage of ceramic magnets is their high electric resistance value

which reduces eddy-current losses to such a point that it can be neglected. The other advantage of ceramic over Alnico is that its demagnetization curve is linear resulting in a stable operation point.

The newest and currently the most widely used PM in electrical machines is rare-earth magnets made primarily from Neodymium-Iron-Boron. These magnets produce the highest energy product of all magnets. First generation rare-earth magnets were manufactured from SmCo_5 which has a slightly lower energy product than NdFeB, but still much higher than Alnico and ceramic magnets. NdFeB is favoured over SmCo_5 due to its lower cost and possible future improvements. Both NdFeB and SmCo_5 have linear demagnetization curves with a gradient similar to that of ceramic magnets. One of the disadvantages of rare-earth magnets is that H_c is very dependent on the operating temperature of the magnet which in turn negatively affects the demagnetization curve of the magnet. Another drawback is that NdFeB is susceptible to corrosion; this can be eliminated by coating the magnet with a non-corrosive substance. To conclude the comparison, rare-earth magnets produce the highest energy product per unit volume of all the magnetic materials. This in turn means that smaller machines can be manufactured to produce the same output. Thus for the LS PMSM prototype NdFeB magnets will be used.

D.2 Permanent magnet operation

Permanent Magnet's performance with regards to the previously mentioned variables can be plotted on a four quadrant axis graph, however the second quadrant is the most important as this contains the information regarding the performance of magnets. The B_r value of the magnet is located on the y-axis and is the maximum flux density value that will remain in a closed magnetic structure if the applied mmf is zero. The H_c value of the magnet is located on the x-axis and indicates the magnetic field intensity needed to force the flux density of the magnets to zero, thus the H_c value of a PM is found on the negative x-axis [1, 2]. By joining the two points, the demagnetization curve is formed and looks similar to that of the blue line in Figure D.1. The red line also represents a PM demagnetization curve, however for this magnet, the knee point is situated in the second quadrant. If the magnet is forced below this point, the magnet can partially demagnetized resulting in a lower B_r value [5]. To prevent this from occurring, the magnetic circuit needs to be designed that the load line or magnet operating point is well above the knee point to prevent the load point moving below this point.

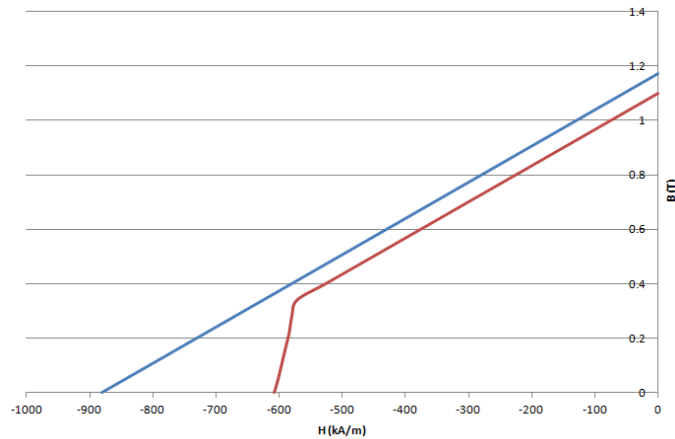


Figure D.1: BH curves of typical NdFeB magnets

From the graph in Figure D.1, the magnets energy production graph can be constructed. This is done by multiplying the B and H values of each point along the demagnetization line. This value can then be plotted against B or H , for machine design purposes BH is plotted against B . The graph in Figure D.2 is the magnetic energy product line of the blue graph in Figure D.1. The maximum energy produced by the magnet occurs when the magnet is operated at $\pm 0.6\text{T}$ for this specific grade of magnet.

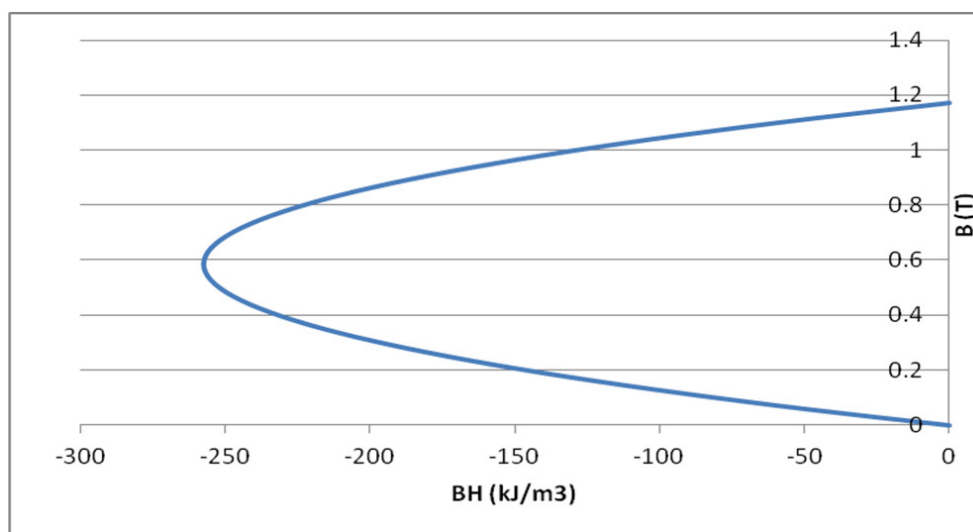


Figure D.2: BH_{max} curve of an NdFeB magnet

If the demagnetization line of the magnet is linear, the optimal energy point will be at half the remanent value of the magnet. This however is not the case if the magnet's knee point is situated in the second quadrant. This is usually indicated on the supplier's data sheet.

As PM performance is affected by the operating temperature the demagnetizing line needs to be adapted to the specific operating temperature as this reduces the performance. The information supplied by the manufacturers regarding the B_r , H_c and BH_{max} is at a specific temperature (usually $20\text{ }^{\circ}\text{C}$); coinciding with these values is their respective temperature influence coefficient expressed in a percentage decrease per

°C. Table D.2 below contains a practical depiction of the temperature effects on the magnet when the operating temperature is 50 °C. The B_r temperature coefficient is -0.11 %/°C and H_c 's is -0.6 %/°C.

Table D.2: Temperature influence on PM material

	B_r	H_c	BH_{max}
Value at 20 °C	1.22 T	900 kA/m	279 kJ/m ³
% decrease at 50 °C	3.3%	18%	9%
Value at 50 °C	1.18 T	739 kA/m	254 kJ/m ³

Along with the relevant electromagnetic data on the manufacturer's sheet, different grades of magnets of the same material type are listed. The grade of the magnet is directly linked to the BH_{max} value of the magnets. The higher grade is associated with a higher energy product. This in turn also increases the price of the magnets which is linked to the grade and volume of the magnet. Thus when sizing the PMs, the decision must be made regarding the energy per volume vs. cost. A low magnetic energy magnet can be just as effective as a small high energy magnet as long as both magnets can supply the required magnetic energy demand. A practical example of this would be if three equal size PM of different grads are used in a magnetic circuit which has the energy demand of 270 kJ/m³, as indicated in Figure D.3. If the magnetic operating point is selected between 0.6 to 0.8 T, the N28 grade magnet will only be able to supply ± 200 kJ/m³. Although the N48 grade magnet would be able to supply the demand, using this magnet would be unwise as it will increase the flux density value in the magnetic circuit. However N38 grade would be the best fit. By reducing the volume of the N48 grade magnet the per volume energy product will also be reduce resulting in a smaller magnets in the magnetic circuit. The N28 grade magnet volume needs to be increased to produce the required energy.

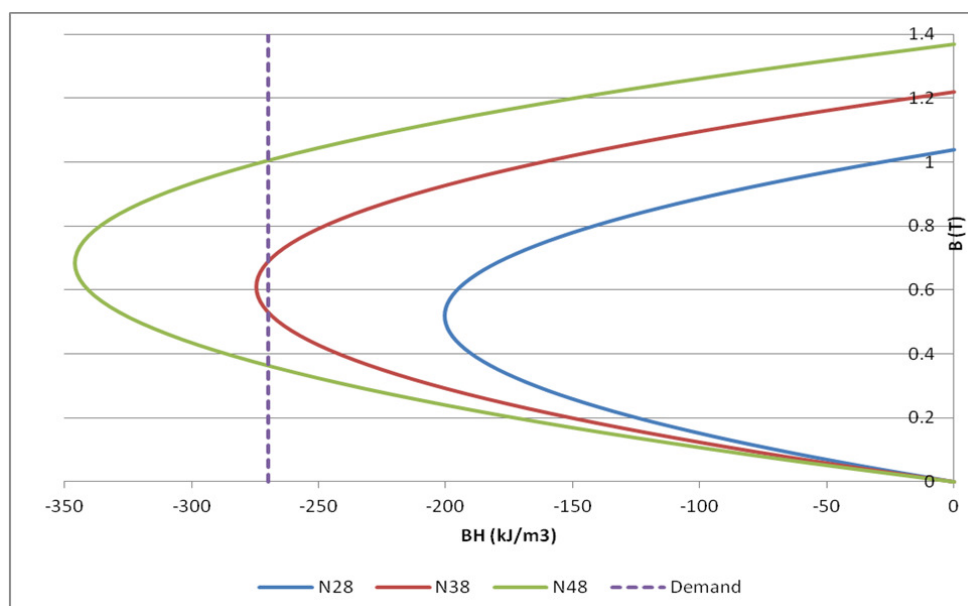


Figure D.3: Magnetic energy demand example

Appendix E: Presented Article

Influence of Magnet Size and Rotor Topology on the Air-gap Flux Density of a Radial Flux PMSM

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Abstract – When designing the rotor of a radial flux permanent magnet synchronous machine (PMSM), one key part is the sizing of the permanent magnets (PM) in the rotor to produce the required air-gap flux density. This paper focuses on the effect that different coefficients have on the air-gap flux density of four radial flux PMSM rotor topologies. A direct connection is shown between magnet volume and flux producing magnet area with the aid of static finite element model simulations of the four rotor topologies. With this knowledge, the calculation of the flux producing magnet area can be done with ease once the minimum magnet volume has been determined. This technique can also be applied in the design of line-start PMSM rotors where the rotor area is limited.

Keywords— PMSM, LS PMSM, Interior magnet rotor, Surface mount magnet rotor, Permanent magnet sizing.

I. INTRODUCTION

When designing an electrical machine several factors influence the performance of the machine. One such factor in permanent magnet synchronous machines (PMSM) is the permanent magnets (PM) used in the rotor to aid in air-gap flux density [1]. The same is also true for line-start PMSMs as they are closely related to PMSM's and operates in the same manner at steady state operation [2, 3].

The air-gap flux density of a machine is one of the main parameters in the design process. This is greatly influenced by the specific sizing dimensions of rotor PMs. The magnet volume and flux producing area of the magnets influences the machines performance in both steady and transient state thus, the required magnet volume needs to be determined first. The magnet volume is directly linked to the potential maximum energy production (BH_{max}) of the PMs which intern affects the breaking torque and the steady state inductances of PMSM machines [4, 5, 6]. Once the magnet volume has been determined, the PM's specific sizing dimensions needs to be established for the specific rotor topology. This article focuses on the magnet volume and the flux providing area of PMs with respect to the air-gap flux density. This is done to inspect both magnet energy and flux production of the magnet at its operating point in comparison with the rotor topology and the correlating air-gap values.

II. PMSM ROTOR TOPOLOGY

A PMSM air-gap flux density is greatly influenced by the placement of the PM in the rotor [7, 8, 9]. All PMSM rotor topologies can be divided into two categories: surface mount (SM) and interior/imbedded magnets (IM) [2, 7, 9,10]. The focus of this article is the effect that the different rotor topologies have on the air-gap flux density, flux and pole flux distribution. The air-gap is used as the focus point and as such the PM placement is classified as seen from the air-gap into the rotor. The four most common PM placement are: surface mount magnets (SMM), slotted surface mount magnets (SSMM), interior radial flux magnets (IRFM) and interior circumferential flux magnets (ICFM). The application of the motor determines which topology is the best suited as none of the topologies are universally superior. Figure 1 illustrates the four topologies.

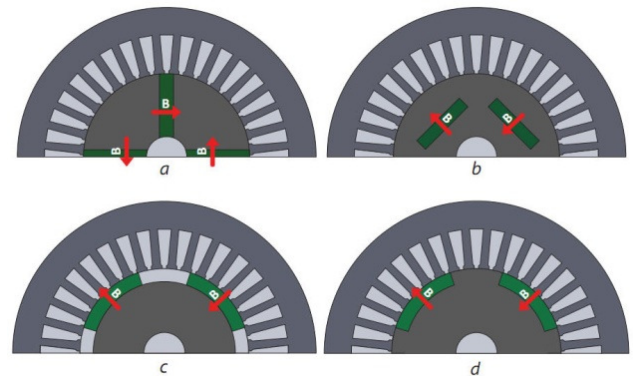


Fig. 1: PM rotor topologies : a) ICFM; b) IRFM; c) SMM and d) SSMM

Figure. 1 a-d represents a four pole PMSM of each topology, each with similar cross sectional PM area and axial length. The stator is that of a 7.5 kW double layer 36 slot PMSM. The saturation effects of the stator teeth and yoke can be neglected as the actual air gap flux value is not as important as the influence of the change in PM volume on the air gap flux. The magnet used in the simulations has a B_r (PM remanence flux density value) of 1.25 T. This paper investigates the influences that effects the change in magnet volume has on the air-gap flux density and flux. For a better understanding of the magnetic pole forming in the different topologies, adequate information is generally available in electrical machine literature [3, 11, 12].

III. ANALYSIS AND CALCULATIONS OF AIR-GAP FLUX DENSITY

Four studies were done on each of the four rotor topologies. The four studies are: reference simulation, pole arch coefficient, magnet thickness coefficient and magnet depth coefficient. The method for each is as follows:

A. Reference Simulation

The first simulation was done to use as a reference point. The reference simulations are as in Fig. 1. All the PM has equal volume and face area. This simulation was also used to compare the four topologies with each other and to calculate the initial values for all the relevant coefficients.

B. Pole Arch Coefficient

The Pole Arch Coefficient (α_{pa}) is determined as

$$\alpha_{pa} = \frac{\tau_s}{\tau_p} \quad (1)$$

with τ_p the pole pitch and τ_s the poles arch width (the length that the flux density spans in the pole pitch, $\tau_p \geq \tau_s$), both in degrees [7]. τ_s is determined as follow: start from the middle point of the rotor and draw a radial line to the inner point of a magnet closes to the pole pitch end. Do the same for the other magnet that also aid in forming the pole arch. The angle between the two radial lines is the poles arch width. As the magnet volume increases in a tangential direction, α_{pa} will decrease.

C. Magnetic Thickness Coefficient

The Magnetic Thickness coefficient (α_{mt}) formula

$$\alpha_{mt} = \frac{D_{mi}}{D_{mo}} \quad (2)$$

was derived from [1]. D_{mi} is the inner most point of the magnet diameter and D_{mo} the outer most point of the magnet. D_{mo} is held at a fixed position and D_{mi} is varied. As the magnet's volume is increased in an inward radial direction, α_{mt} decreases. The coefficient is not dependant on the flux direction, but rather on the inner and outer radius of the magnet.

D. Magnet Depth Coefficient

Magnet Depth Coefficient (α_{md}) is only applicable in IMPM rotor topology. To calculate α_{md} , the volume is kept the same, but the magnet dept from the rotor surface is varied. α_{md} is thus the relation between

$$\alpha_{md} = \frac{D_{mo}}{D_r} \quad (3)$$

with D_{mo} is as in C and D_r is the rotor's diameter. This simulation is done in conjunction with C to determine the optimum dept and magnet volume.

For all four studies, the magnetic flux density graphs will be drawn on the stator teeth over two poles. The total pole flux will also be calculated from this data.

IV. MAGNET TOPOLOGY ANALYSIS OF AIR-GAP FLUX DENSITY AND FLUX

The ICMF topology was used as the base reference for the simulations. The reason is the internal magnets volume is greatly limited by the radius of the rotor, thus this topology was used to determine the reference volume. The exact volume and dimensions of the PMs used for this article is not relevant, but rather that the same reference volume PMs are used in all cases.. The numbering of this section corresponds with that of Fig 1, ICRF being A.

A. Internal Circumferential Flux Magnets:

1) One of the major drawbacks of ICFM topology is the leakage flux through the shaft [2, 7, 8, 9]. Upon inspection from the simulating results Fig. 2 was produced.

Figure. 2 shows that an increase in air-gap flux density from $\pm 0.25T$ (25% difference) can be achieved if the shaft leakage flux is eliminated. The calculated air-gap flux difference between the two static simulations is ± 2 mW. This increases as the machine load increase, thus for optimum energy usage of the magnet, the leakage flux must be reduced or eliminated completely. This can be done by using a non-magnetic shaft or shielding material around the shaft [2, 7]. The leakage flux was not eliminated for in remainder of the simulations.

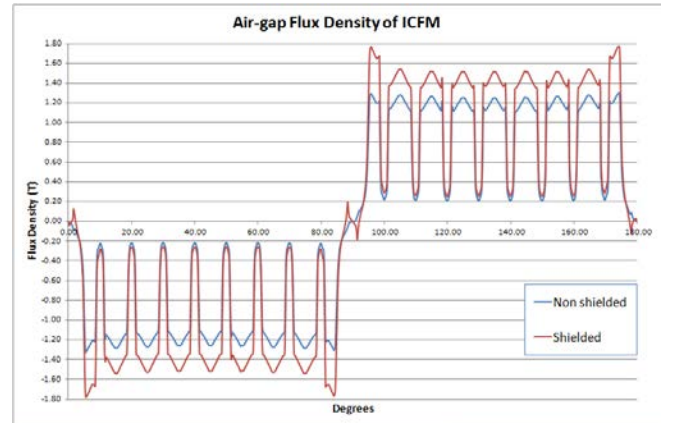


Fig. 2: ICFM – air-gap flux density: leakage flux vs. no-leakage flux

2) The reference design pole arch coefficient is 0.53. When increasing the value of α_{pa} in intervals up to 0.9, the flux density of the air-gap as well as the air-gap flux decreased. This is due to the decrease in the area of the magnetic rotor material facing the air-gap and in turn an increase in PM material on the rotor surface. Figure 3 provides the results from the simulations. From Fig 3, it can also be seen that the flux density distribution per pole does not vary as α_{pa} vary. This is because this topology is dependent on flux forcing to form the machine poles [7, 9, 10]. The flux density distribution is thus independent of the magnet thickness and volume.

Both analysis methods 3 and 4 were not applicable to this topology as the amount of leakage flux is so high that it is not advisable to reduce the magnet length (in a radial direction from either end). If the leakage flux is reduced greatly or eliminated the flux density results will be the same as in 2. As the magnet becomes shorter from either the shaft end or the air-gap end, the air-gap flux density and flux will be reduced. The flux density distribution will however still span the whole pole arch.

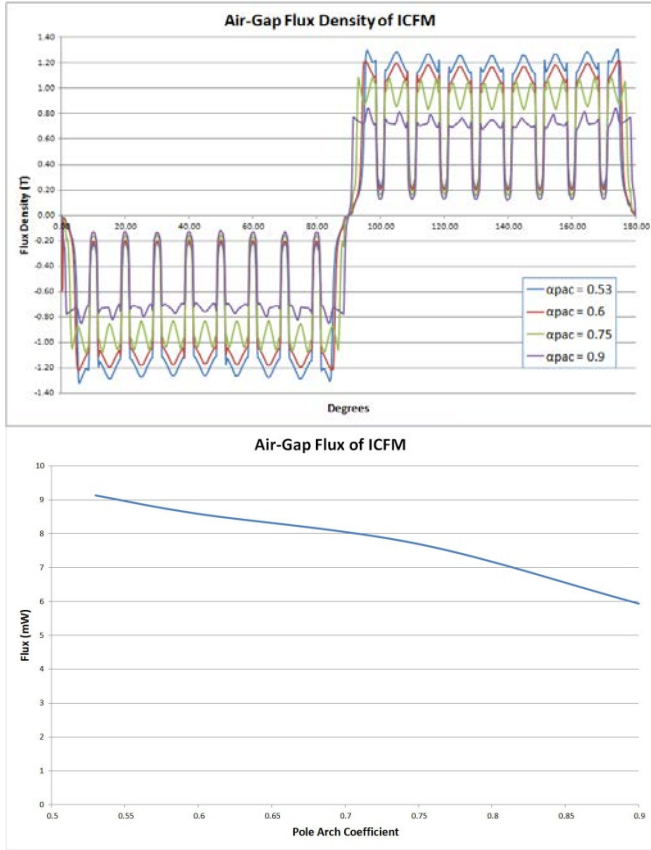


Fig. 3: α_{pa} graph of ICFM a) air-gap flux b) air-gap flux

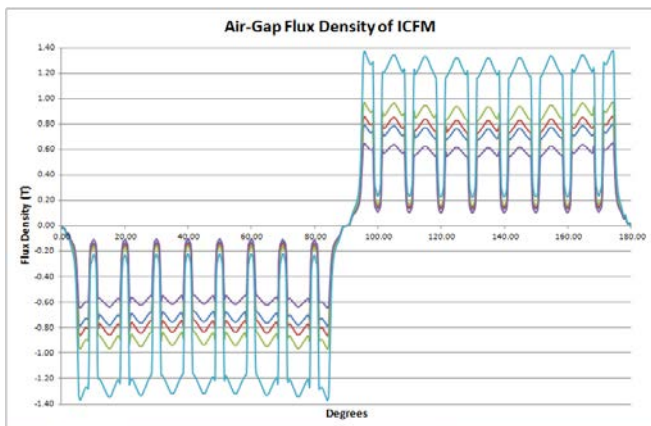


Fig. 4: Air-gap flux density of ICFM as magnet volume decreases in radial direction

Figure 4 was constructed by reducing the reference simulation magnet length by $\pm 25\%$ at the air-gap end (Note, the shaft leakage flux is still present). Non magnetic material was then added in intervals up to the air-gap. The per pole flux increased from 4.5mW to 6.8mW, when the shaft leakage flux was eliminated the flux increased to 9.5mW.

B. Internal Radial Flux Magnets

1) For the reference simulation the same magnet size and shape was used as in A. The magnet was placed one quarter of the radius from the air-gap end. From the simulation results the blue graph was produced in Fig. 5 a).

2) The reference designs pole arch coefficient is 0.27. When increasing the value of α_{pa} in intervals up to 0.75, the flux density of the air-gap as well as the air-gap flux decreased radically. Fig. 5 provides the results from the simulations. When $\alpha_{pa} = 0.75$ the air-gap flux density was near 0T, this is because the rotor leakage flux is very high. The flux flows through the rotor material rather than the air-gap.

As seen in Fig 5, α_{pa} influence the air-gap flux density, air-gap flux as well as the pole flux density distribution. As α_{pa} increases all three air-gap values decreases.

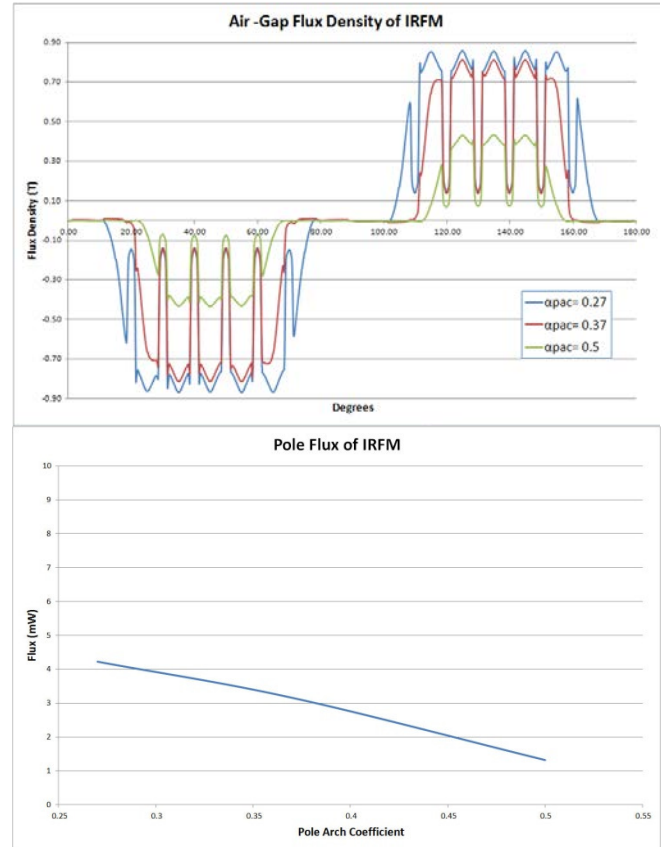


Fig. 5: α_{pa} graph of IRFM a) air-gap flux b) air-gap flux

3) The reference designs magnet thickness coefficient is 0.83. For the analysis the magnets outer diameter was kept constant. α_{mt} was varied between 0.65 and 0.9. Fig. 6 contains the results of the simulations.

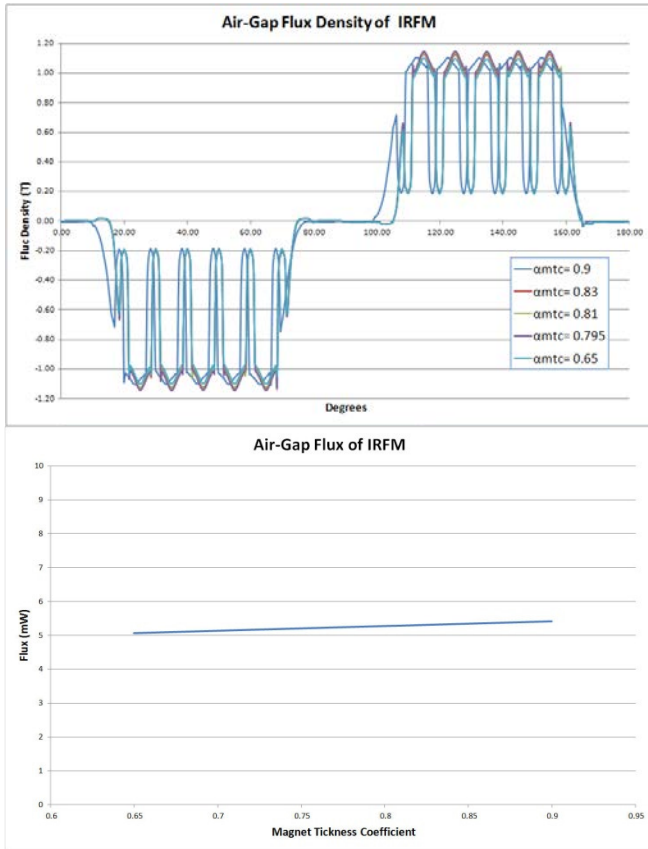


Fig. 6: α_{mt} graph of IRFM a) air-gap flux, b) air-gap flux

As seen in Fig. 6 the air-gap flux density is not affected by α_{mt} , but the air-gap flux is slightly affected by α_{mt} . The flux decreases as the volume of the magnet increases.

4) The reference designs magnet depth coefficient is 0.83. For this analysis, α_{md} is varied between 0.9 and 0.6. At $\alpha_{md} = 0.6$ the magnets is at its maximum depth from the rotor surface as the magnet sides are now touching each other. The minimum value of α_{md} is determined by the length of the magnet. Fig 7 displays the results of this analysis.

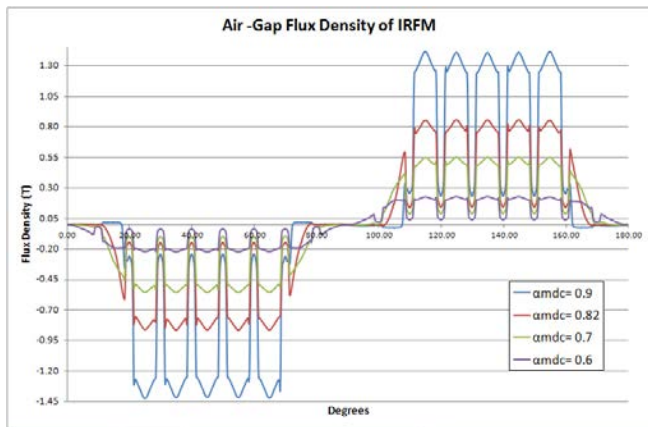


Fig. 7 a): α_{md} graph of IRFM air-gap flux

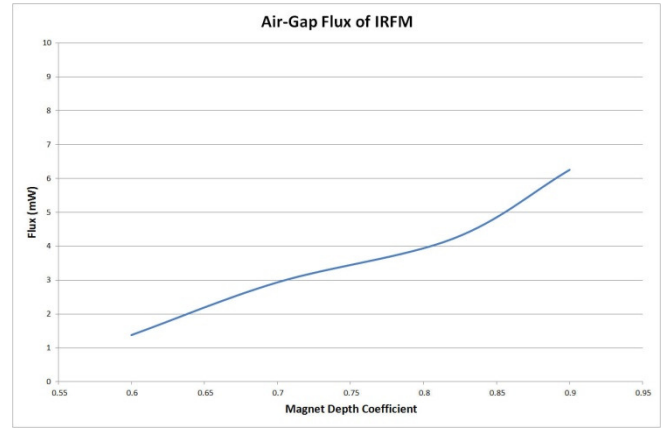


Fig. 7 b): α_{md} graph of IRFM air-gap flux

As seen in Figure 7, as the magnet moves closer to the rotor surface, both the air-gap flux density and air-gap flux increases. The deeper the magnets are placed in the rotor the higher the leakage flux is through the rotor material.

C. Surface Mount Magnets

1) For the reference simulation the same magnet size was used as in A. The magnets thickness was kept the same, but it was rounded to fit on to the rotor surface. From the simulating results the blue graph in Fig. 8 a) was produced.

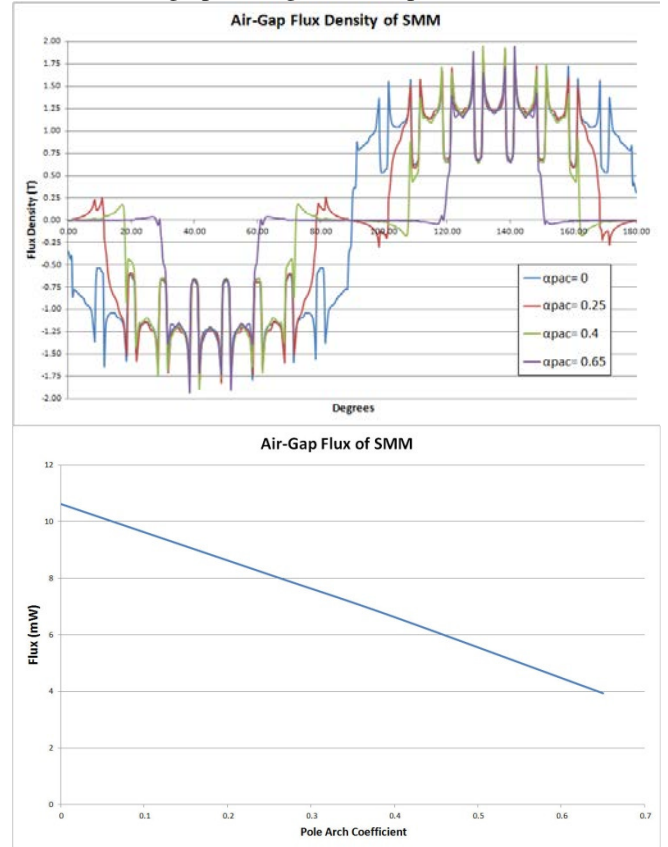


Fig. 8: α_{pa} graph of SMM a) air-gap flux, b) air-gap flux

2) The reference designs pole arch coefficient is 0.4. When varying the value of α_{pa} between 0 to 0.65. The simulation results are indicated in Fig. 8.

As depicted in Fig. 8 the maximum air-gap flux density is not influenced by α_{pa} , however the pole flux density distribution is influenced greatly. As α_{pa} decreases, an increase in pole flux and pole flux density distribution occurs.

3) The reference designs magnet thickness coefficient is 0.85. α_{mt} was varied between 0.9 and 0.55. Figure 9 displays the results. As seen in Fig. 9, the air-gap flux density and pole flux density distribution is not influenced by α_{mt} , the flux is however only influenced minimally as α_{mt} varied. Analysis 4 is not applicable to SMM as the magnet stays on the surface.

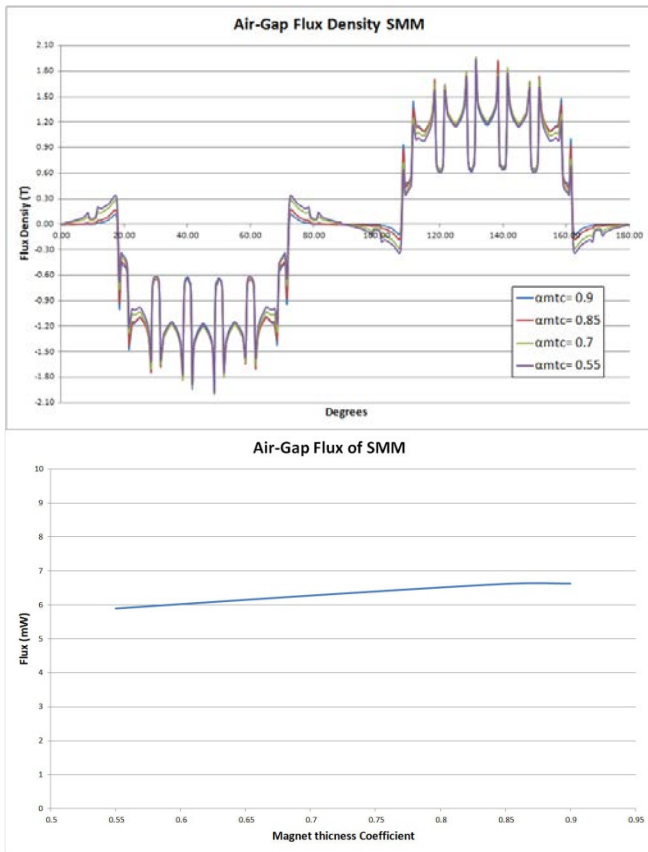


Fig. 9: α_{mt} graph of SMM a) air-gap flux, b) air-gap flux

D. Slotted Surface Mount Magnets

1) For the reference simulation the same magnet size and shape was used as in C. The magnet however is now placed in slots on the surface. From the simulating results the blue graph in Fig. 10 a) was produced.

2) The reference designs pole arch coefficient is 0.4. When varying the value of α_{pa} between 0 to 0.65. The simulation results are indicated in Fig. 10. As seen in Fig. 12, α_{pa} does not affect the maximum air-gap flux density but it effects the Air-Gap flux and pole flux density distribution

greatly. The higher α_{pa} , the lower the pole flux. The simulated results were similar to that of SMM analyse.

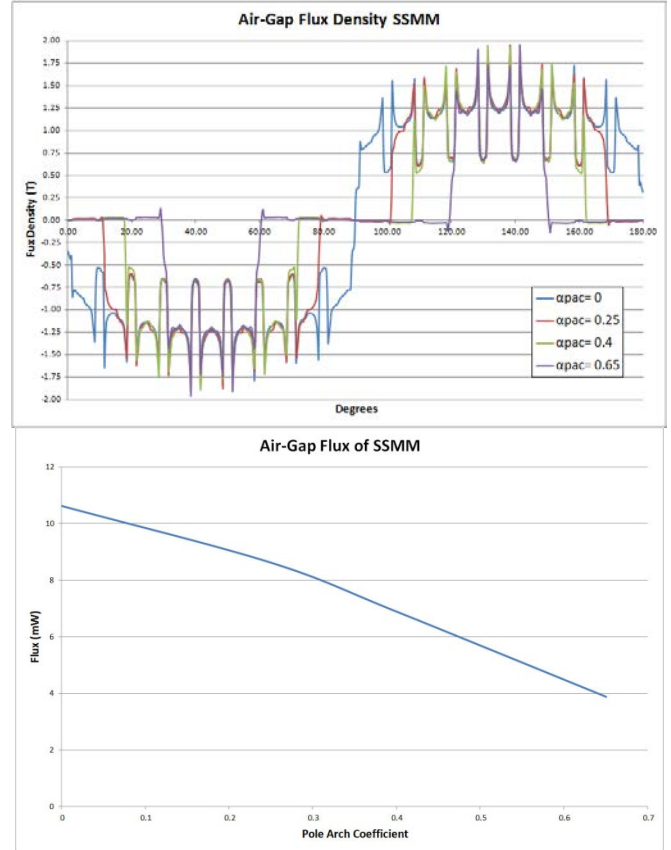


Fig. 10: α_{pa} graph of SSMM a) air-gap flux, b) air-gap flux

3) The reference designs magnet thickness coefficient is 0.85. α_{mt} was varied between 0.9 and 0.55. Fig. 11 displays the results of the simulation. As seen in Fig. 11 the air-gap flux density and pole flux density distribution is not influenced by α_{mt} , the flux is however only influenced minimally as α_{mt} varied.

Analysis 4 is not applicable to SSMM as the magnet stays on the surface.

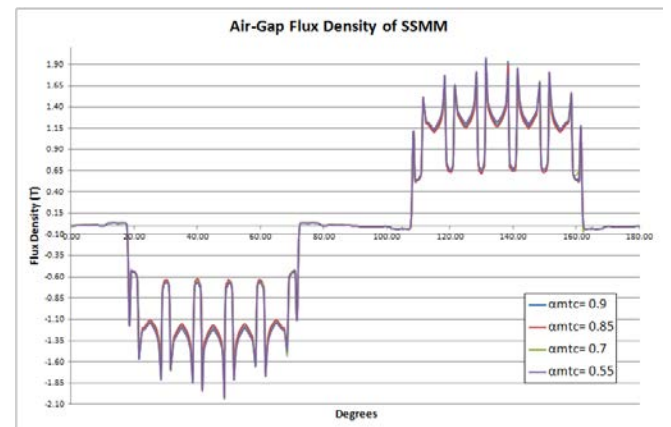


Fig 11 a): α_{mt} graph of SSMM air-gap flux

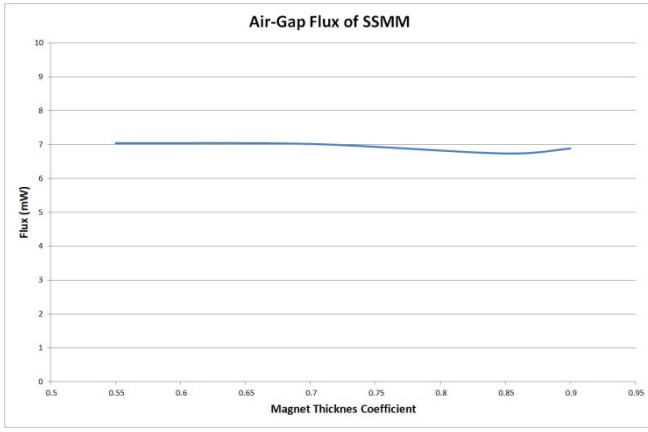


Fig. 11 b): α_{mt} graph of SSMM air-gap flux

V. RESULTS AND CONCLUSION

Each of the four rotor topologies' simulation results will be discussed. The second part of this section the focus will be placed on the reference simulations results where all four topologies magnet volume is the same.

A. ICFM: The biggest drawback of this design is the high flux leakage through the rotor shaft and air-gap end of the magnet. This occurs when the magnets do not span the entire radius of the rotor or the shaft of the rotor is not shielded. If the leakage flux can be eliminated or reduced by using shielding material or incorporating flux limiting material saturation zones, this design can provide numerous advantages [2, 3, 7, 8, 11]. From the simulation results it can be concluded that, a high air-gap flux density value can still be gained with a relatively thin magnet as depicted in the pole arch coefficient analysis. The corresponding flux value is still high even though the flux leakage through the shaft was still present. A 15% to 25% increase in both flux density and flux was noted when the leakage flux was eliminated. The big advantage of this topology is high air-gap flux density and flux with relatively small volume magnets. A further advantage is the very high pole flux density distribution and the nearly rectangular air-gap flux density shape across the whole pole. When the magnet's thickness and length is reduced, a high flux density and flux value can be expected. This is because the design incorporates flux forcing to form the machine poles, thus an air-gap flux density value higher than the B_r (1.25T for the magnets used) can be gained.

B. IRFM: The effect of the leakage flux in the rotor is a major drawback of all imbedded/internal magnet placement designs [3, 8]. One of the factors that influence the amount of leakage flux is influenced by α_{md} , the lower this coefficient, the higher the leakage flux. This design is not prone to leakage flux and can easily be eliminated by incorporating flux barriers in the rotor. From the simulation results it can clearly be seen that with this design the air-gap flux density will very rarely rise above the B_r value of the magnet. This design topology also promotes very low air-gap flux. If the leakage flux is eliminated in the rotor, the fundamental formula of $\phi_{ag} = B_r A_{ag}$ is applicable on this design topology. This can be seen in

the α_{ap} analyses, thus the values of both the air-gap flux density and flux can be manipulated. α_{mt} had no effect on the flux density and very small effect on the air-gap flux. If the leakage flux can't be eliminated completely in the rotor designs, α_{md} clearly indicated the effects it have on both the air-gap flux density and flux. The deeper the magnets is placed in the rotor, the harder it becomes to eliminate the leakage flux, thus it is a good idea to place the magnets at least half way between the rotor surface and shaft.

C. SMM: Similar to the interior magnet type rotor topologies, this design topology has no leakage flux in the rotor itself. If the air-gap distance increase beyond a certain point, the leakage flux can no longer be neglected [3]. The average air-gap flux density on this design will be the same as the B_r of the magnets used in static simulations. Referring to the simulations results it is noted that the peak air-gap flux density is not influenced by α_{pa} but only the air-gap flux. This is because as the magnet becomes shorter, the flux density distribution is reduced, this is clearly noted in the equation $\phi_{ag} = B_r A_{ag}$. However α_{mt} had minimal influence on both the air-gap flux density and the flux. Thus to optimize the magnet's energy product, a thin magnet that is the length of the pole span ($\alpha_{pa} = 0$) will produce the best pole flux density distribution.

D. SSMM: As with SMM, this design topology's flux leakage is dependant in the air-gap. This design topology's simulation result is very similar to that of SMM. The best usage of magnet volume would be if the magnet spans the entire pole arch, thus making this design an inefficient one with respect to optimal magnet volume.

E. Topology Magnet volume comparison: As stated, the reference simulation of all four topologies had the same magnet volume with the same thickness and centre length. With the information gained from the different coefficient analyses, a relevant conclusion can be made to use the installed magnets at its optimum energy point (BH_{max}). Fig. 12 contains the four rotor topologies air-gap Flux density graphs ((a) ICFM with shaft flux leakage, (b) ICFM with no flux leakage.)

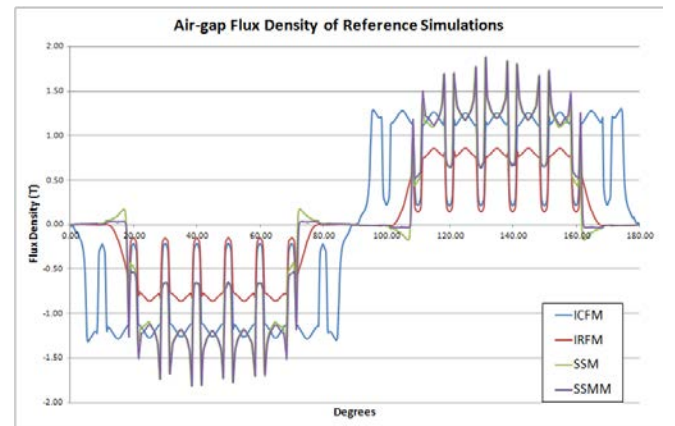


Fig 12 a): Air-gap flux of all four topologies ICRF with leakage flux

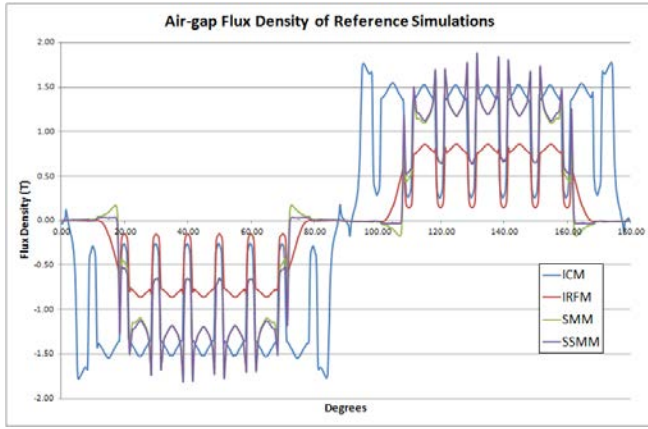


Fig. 12 b): Air-gap flux of all four topologies with no leakage flux

As indicated in Fig. 12, ICFM provides a better pole flux density distribution than the other three topologies. However, both SSM and SSMM can also provide the same distribution when the magnets span the entire pole. This would lead to a change in magnet volume.

Figure 13 contains a comparison between the different topologies with respect to the average flux density and flux of the air-gap. The flux in the bar graph was down scaled by one decimal for better visualisation. This bar graph was constructed with the simulation results in Fig. 14.

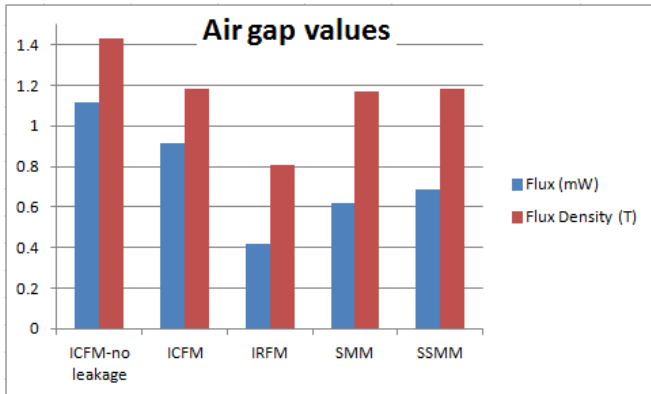


Fig. 13: Comparison of reference simulation results

As indicated in Fig. 13, for this magnet shape, the ICFM topology produces the highest air-gap values. However, from the coefficient analysis it became clear that the coefficient with the biggest influence on the air-gap values is α_{pa} and that the thickness (α_{mt}) of the magnet hardly had any influence. Thus the same air-gap values can be reached with the same magnet volume but with a different shape. The IRFM topology can become challenging as the air-gap values are also influenced by the depth of the magnet (α_{md}); this coefficient limits the length of the magnet. Interior magnet rotors electromagnetic design and mechanical design can become complicated and both design aspects need to be kept in mind during the design phase with great emphasis placed on if the machine design's assembly is possible.

VI. CONCLUSION

In PM motor design, one of the biggest cost determining factors is the magnets used in the machine, thus to design the machine not to operate at or near the magnet maximum energy (BH_{max}) point when at full load is wasting magnetic material and adding unnecessary cost. This will also influence the machine's operation at both steady state and transient operation. The static air-gap value of the machine is usually higher than at full load.

When designing a radial flux electric machine both the magnet volume and shape play a big role in the air-gap values and machine performance. To optimize the PM material used in a machine one first has to determine the needed magnet volume at a certain energy product point (BH) and air-gap flux density value, preferably the maximum energy point, to limit the needed PM material in the rotor [12]. Once the volume is known, α_{pa} can be used to manipulate the needed flux density and flux values [7]. As this coefficient focuses on the flux production of the magnet, $\phi_m = B_r A_m$.

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DESIGN PROCEDURE OF A LINE-START PERMANENT MAGNET SYNCHRONOUS MACHINE

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Abstract: This paper presents a method for designing an LS-PMSM by dividing the design into sub-machine components. Each of the individual sub-machine components is then designed using several classical machine design principles. Once all the components have been designed they can be combined to form an LS-PMSM. The machine design is verified in two manners. Firstly the selected flux density values are compared with the FEM values to ensure they are in range of each other. Secondly the calculated asynchronous and synchronous performance is compared with the simulation results from a commercial design software package. Finally the proposed method is applied to the design of a prototype machine. The results show a good agreement with that obtained from commercial design package.

Key Words. LS-PSMS; permanent magnet synchronous motor; line-start; machine design.

1. INTRODUCTION

Induction machines (IM) are the backbone of the industry machine installation as they are robust, reliable and have relatively high efficiency. In a typical South African chemical plant the majority of electrical machines are IMs ranging between 2.2 kW and 22 kW. The majority of the machines in this power range are connected to pump and fan loads. Both of these load types are operated at a constant speed. As the price of electrical energy increases and stricter efficiency regulations are implemented in place, there is a need for more efficient electrical machines.

In 2008 the International Electro-technical Committee (IEC) standardized the efficiency classes for three-phase, line-fed general purpose machines to promote a market transformation [1]. The IE4 (Super-premium efficiency) is the highest of the four standards and was initially only intended to be informative. Many machine manufacturers saw no possibility of reaching these efficiency levels with IMs within the respective IEC frames [1]. This forced the manufacturers to investigate other technologies like permanent magnet synchronous machines (PMSM). The problem with a PMSM is that it is not self-starting thus limiting the possibility of acting as a direct replacement of an IM.

In 1971, Binns et. al. proposed a new type of self-starting synchronous machine that utilized both permanent magnets (PM) and a die-cast cage within a single stack of laminations [2]. The cage generates asynchronous torque during transient operation, which makes it possible for the machine to be started directly from the ac supply. This machine type is now known as a line-start permanent magnet synchronous machine (LS-PMSM).

In theory an LS-PMSM is a hybrid of a PMSM and an IM in a single rotor. During the transient period the asynchronous torque (T_{asy}) is the result of the interaction between the cage torque (T_c) and breaking torque (T_m) [3]. Both the breaking torque and the cage torque are dependent on different component in the

machine. T_m is generated by the PM and has a negative effect on the machine's start-up performance. The magnitude of this negative torque component is dependent on the PM volume [4, 5]. At steady state, the synchronous torque (T_{syn}) is mainly produced by the PM alignment torque. Depending on the rotor topology there can also be a reluctance torque component. The performance of the machine is greatly influenced by the interaction between the IM and PMSM topology [3,5]. Once the rotor is synchronized with the stator's rotating MMF, the rotor cage has no effect on torque production. This eliminates the cage rotor losses of the machine at steady state because there are no induced currents in the bars. Thus the efficiency of an LS-PMSM can be higher than that of an IM [3]. Figure 1 illustrates the torque components of a generic LS PMSM.

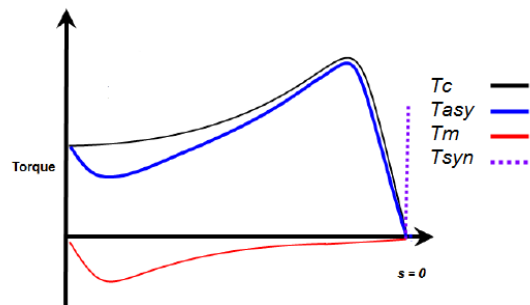


Figure 1: LS PMSM theoretical torque curves [3, 6]

Several comparison studies between IMs and LS PMSMs have been done during the last couple of years. In [7], it is concluded that an LS-PMSM has superior performance when compared to an IM with regards to pump, fan and compressor loads (constant speed with long operating cycle loads). Typically an LS-PMSM has higher efficiency, power factor and torque density and its operating temperature is lower than that of an IM due to the absence of rotor bar currents [1,7-9]. Although these machines' initial cost is higher than that of an IM, the cost of ownership is much less, making it a very suitable machine for certain applications. The drawback of this machine is however its starting torque and synchronisation capabilities, which place limits on its application capabilities.

This paper presents a design methodology for LS-PMSM, which utilizes classical machine design theory in combination with finite element method (FEM) analysis. The proposed method is applied to the design of a four-pole, low voltage LS-PMSM driving a fan-type load.

2. PROPOSED DESIGN METHOD

When designing an LS-PMSM there are two options available. The most popular one found in literature is a retro-fit design or IM rotor swop-out [6]. Since an LS-PMSM can operate with the same stator as an IM the IM rotor can be replaced with an LS-PMSM rotor [3, 4, 9, 10]. This option eliminates the need of machine sizing. The second option is to do a complete machine design. The method proposed in this paper can be used for both options. For the retro-fit option the stator must be characterised instead of designed but the same steps must be followed as some of the stator parameters are needed during the rotor design.

Figure 2 shows the low-level design flow diagram. The black dashed lines represents the need for changes in the design should the relevant parameters and flux density's not be met during the FEM verifications. Each of the four blocks is an independent sub-design for a component of the machine. As the rotor contains both an IM and a PMSM, they can be designed separately. Once both rotor designs are done they can be combined into one rotor. The PMSM is designed first as the magnetic braking torque must be known to design the cage torque curve. During the design of the PMSM the possible cage location must be kept in mind.

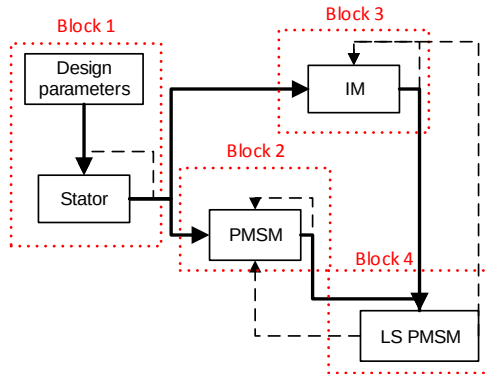


Figure 2: LS PMSM design processes

Figure 3 contains the design steps of Block 1. This block has a 2-step design: in the first step the machine is sized to determine rotor diameter (D_r) and active length (l). In the second step the winding factor (k_w), turns per coil (N_s) and the coil slot is determined. This design method was adapted and compiled from [10] and [11].

Once the stator design is complete the design must be verified to determine if the selected flux density values are achieved. This is done through static FEM analysis. The PMSM (Block 2) and IM (Block 3) design flow diagrams are shown in Figure 4 and 5,

respectively. Both Block 2 and 3 was compiled using [10 – 12].

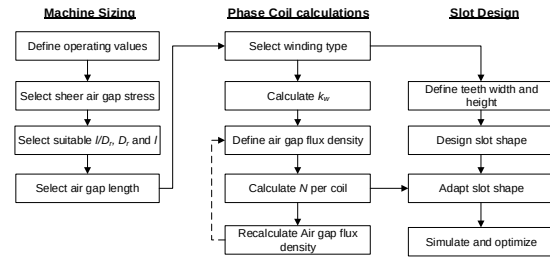


Figure 3: Design flow diagram of Block 1 [6]

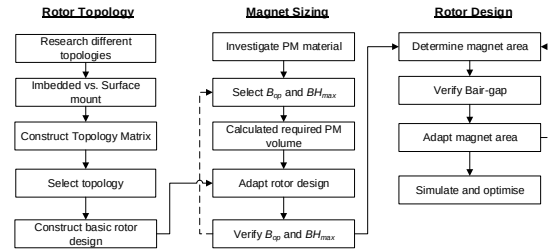


Figure 4: Design flow diagram of Block 2 [6].

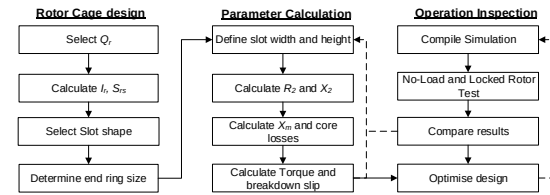


Figure 5: Design flow diagram of Block 3 [6].

Once both the PMSM and IM rotors sections have been designed and verified the two must be combined to form the LS-PMSM rotor. The performance of the LS-PMSM can then be compared to the two independent machines. Theoretically at steady state the LS-PMSM's performance will be similar to that of the PMSM and during transient operations to the IM although there will be difference due to the presence of T_m . For the static FEM simulations FEMM is used and for the performance comparison with the calculated model, Maxwell[®] RMxpert is used.

3. DESIGN AND VERIFICATION

In this section the afore-proposed method is implemented in the design of a prototype LS-PMSM machine.

3.1 Design Specifications

Table 1 contains the design specifications of the prototype machine. Besides the specifications in Table 1, the starting current of the machine is required to be similar to that of an IM with the same power rating. Furthermore the prototype must fit in a

132 size frame or smaller, enabling the machine to be used as a direct replacement for an existing IM.

Table 1: Prototype Machine Specifications

Specification	Value
Power (kW)	7.5
Phase	3
Line Voltage (V)	525
Line Current (A)	± 10
Number of Poles	4
Preferred Line Connection	Star
Rated Speed (RPM)	1500
Duty Cycle	S1

3.2 Machine Sizing

To determine l and D_r two options are available. The first option is to use the values of the IM counterpart and change either l or D_r while the second option varies both parameters. In both options (1) is used

$$\begin{aligned} T_{rated} &= \sigma r_r S_r \\ T_{rated} &= \sigma r_r (2\pi r_r l) \\ T_{rated} &= \frac{\sigma \pi D_r^2 l}{2} \\ \sigma &= \frac{2T_{rated}}{\pi D_r^2 l} \end{aligned} \quad (1)$$

with T_{rated} being the rated torque of the machine and σ the tangential stress or sheer air-gap stress of the machine [10]. The tangential stress is the main torque producing component when it acts on the rotor surface. If σ is selected to be within limits as set out in [6] only the l and D_r is unknown. By fixing one of the two unknown and performing a variable sweep the other can be selected out of the graph. The selected value is then re-checked by fixing the selected value and performing a sweep again. A graph similar to Figure 6 can then be compiled. The l/D_r ratio must be checked to aid in final selection with the empirical ratio rule i.e.

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

with p being the number of pole pairs [10]. The rotor diameter D_r must not be selected too small as this complicates the rotor design and limits the cross sectional area. The values of l and D_r for the design is 115 mm and 113.5 mm respectively. The air-gap (δ) length is selected as 0.5 mm.

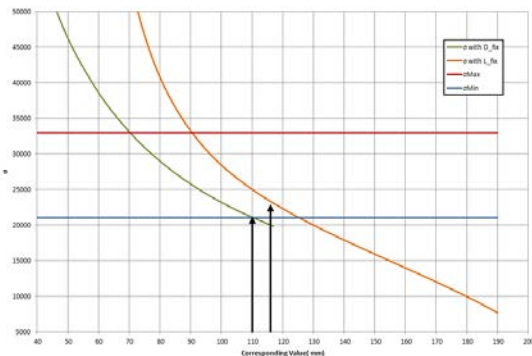


Figure 6: Variable sweep of LS PMSM design

3.3 Stator Design

Once the machine is sized the next step is to determine the winding factor (k_w), number of turns per coil phase (N_s) and the required slot area. To do this, the air gap flux density (B_δ) must be selected. For a PMSM the value of B_δ is usually between 0.85-1.05 T depending on the rotor topology, as for an IM the values is between 0.7-0.9 T [10]. By selecting the value as 0.85 T both machine's boundaries are respected.

In [6], a harmonic comparison study was done to select the winding layout. A 36 slot, overlapping double-layer layout was selected as it provided the best slot harmonics of the options. The coils are short pitched by one slot. Table 2 contains the winding configurations that were investigated with Q_s as the number of slots, q the number of slots per pole per phase, k_d the distribution factor, k_p the pitch factor and k_{sq} the skewing factor.

Table 2: Compilation of Various Winding Factors [6]

Q_s	q	k_d	k_p	k_{sq}	k_w
24	2	0.9659	1	0.9886	0.9549
24	2	0.9659	0.9659	0.9886	0.9223
36	3	0.9597	1	0.9949	0.9548
36	3	0.9597	0.9848	0.9949	0.9403
48	4	0.9576	1	0.9971	0.9548
48	4	0.9576	0.9914	0.9971	0.9466

As it is very difficult to skew the rotor bars and PM with respect to the stator it was decided to skew the stator slots by one slot pitch instead. This technique is not that common in mass produced IMs but is used in specific cases.

To calculate N_s , k_w is used in

$$\begin{aligned} N_s &= \frac{\sqrt{2} E_m}{\omega_e k_{w1} \Phi_m} \\ N_s &= \frac{\sqrt{2} E_m}{\omega_e k_{w1} (\alpha_{pa} B_\delta \tau_p l')} \end{aligned} \quad (3)$$

with Φ_m being the PM flux [10]. Note the expansion of Φ_m in (3), where the air gap flux density B_δ and pole area is used. Table 3 contains the values and the definition of each symbol in (3). The pole arch coefficient is dependent on the rotor topology and seldom exceeds 0.85. The actual value can only be calculated once the final rotor has been designed.

Table 3: Values for Calculations of N_s

Symbol	Value	Description
E_m	303V	Phase voltage
ω_e	100 π rad/s	Electrical angular velocity
k_{w1}	0.9403	Winding factor
α_{ac}	0.8	Pole arch coefficient
B_δ	0.85 T	Air gap flux density
τ_p	0.089 m	Pole pitch
l'	0.115 m	Active length

Using the values listed in Table 3, N_s is calculated as 208. This value is rounded to 216 to provide 36 turns per slot. When substituting this value into (3), B_δ is recalculated as 0.868 T which is an acceptable increase of less than 3%.

The required slot size can be calculated using N_s , the wire's copper area and the fill factor (ff). The rated current is calculated as 10A. The stator coil current density (J_s) must be selected to calculate the required size of copper wire [10]. By selecting J_s as 6.5 A/mm² the required copper area per conductor is ± 1.6 mm² thus a SWG17 (1.422 mm $\varnothing$) can be used. However to gain a better fill factor, two SWG19 (1 mm $\varnothing$) is used. By doing so, J_s is re-calculated as 6.3 A/mm². Thus a single slot contains 72 conductors with a copper area of 60 mm². For the LS-PMSM the fill factor is selected as 0.5, thus the required slot area is 120 mm². To further aid in designing the slot shape the stator slot height (h_s) and tooth width (b_{st}) is calculated by selecting the flux density values in the teeth and back yoke as set in [6]. Table 4 contains the selected flux density values and calculated values of h_s and b_{st} . The tooth flux density is at its maximum on the d-axis and gradually decreases towards the q-axis. Thus the value in Table 4 is the peak value and not the average value.

As indicated in Figure 3, once h_t and b_{st} is calculated the slot shape must be selected and optimized. The final slot shapes parameters are listed in Table 5. Once the design is finalised the stator parameters (stator winding resistance (R_l) and the reactance (X_l)) can be calculated. R_l is calculated using the combined method of [10] and [13]. Table 6 contains R_l and the relevant performance factors (λ – not to be mistaken for flux linkage) used to calculate the stator leakage reactance. The leakage effects of skewing can only be incorporated once the initial inductance is calculated [10].

Table 4: Stator Slot Height and Width Parameters

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

Table 5: Stator slot dimensions

	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

Table 6: Calculated Stator Parameters.

Parameter	Value
R_l	1.40544 Ω
Slot leakage (λ_{su})	2.8836 Wb
End winding leakage (λ_{sw})	0.24197 Wb
Zig-zag leakage (λ_{sz})	0.1785 Wb
X_l without skewing	2.343 Ω
Skew leakage (L_{ssq})	75.424 μ H
X_l	2.3668 Ω

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 to check if they are in a good agreement. A simple surface-mounted PM rotor that generates $B_\delta = 0.85$ T may be used to verify the stator design as shown in Figure 7. A full pitch surface mount PMSM

is used as the flux density values of the PM is the flux density value of the air-gap. Thus by setting remanence value of the magnet to 0.85 T, the correct B_δ is obtained for verification. Model A is between two d-axes and Model B between two q-axes. The stator coils in both models are not excited. The simulation results of both models are listed in Table 7. The calculated value differs from that selected values in Table 4 as both the slot height and width was reduced during the slot optimization. M400-50A lamination material is used for the stator.

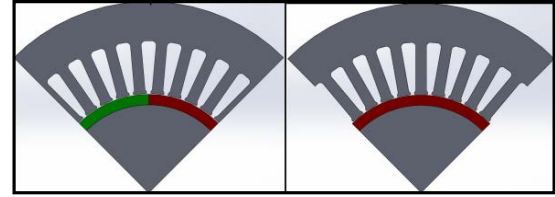


Figure 7: Stator verification models

Table 7: Stator Verification Information

	Stator Yoke Flux Density	
	FEM	Calculated
Model A	1.337 T	1.22 T
Model B	1.383 T	
	Stator Teeth Flux Density	
	FEM	Calculated
Model A	1.206 T	1.3 T
Model B	1.33 T	

From the table it can be seen that the calculated flux density values correlate with the FEM simulated ones.

3.4 Rotor Design

As indicated by Figure 2 the rotor design comprises of two parts. The PMSM rotor is designed first to simplify the design process as the PMs influence the space availability of the rotor the most. Furthermore the breaking torque must be known in order to design the cage to overcome it.

3.4.1 PMSM Rotor Component

Although many rotor topologies of PMSM exists, this study only considers four topologies namely surface mount magnets (SMM), slotted surface mount magnets (SSMM), imbedded radial flux magnets (IRFM) and imbedded circumferential flux magnets (ICFM) as shown in Figure 8.

Surface mounted topologies (SMM and SSMM) exhibit good steady-state performance, but suffer from poor transient operation capabilities [6]. Imbedded radial flux topology leads to high magnet consumption [6,14]. Thus the ICFM topology was selected as it provides the highest B_δ in comparison with the other 3 topologies for the same PM volume [14]. The key design consideration of ICFM topology is the large leakage flux through the shaft. This can be greatly reduced by using a non-magnetic shaft material.

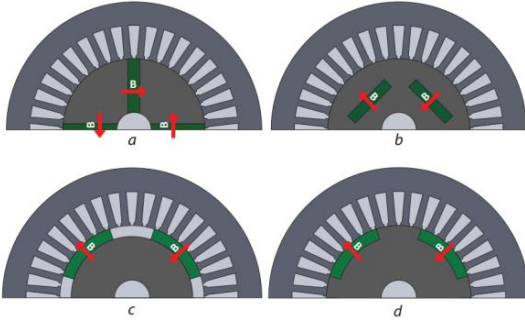


Figure 8: PM rotor topologies: a) ICFM; b) IRFM; c) SMM and d) SSMM [6, 14]

To determine the minimum PM volume to provide an airgap flux density of 0.85 T the assumption was made that there is no leakage flux in the rotor, thus $\Phi_m = \Phi_\delta$ and the PM volume (Vol_m) can be calculated with

$$Vol_m = \frac{Vol_\delta B_\delta^2}{\mu_0 (-H_m B_m)} \quad (3)$$

with Vol_δ the air gap volume and $H_m B_m$ the energy product of the operation point of the magnet [12]. By replacing $H_m B_m$ with the HB_{max} value of the selected PM grade, the magnet provided the maximum magnetic energy per material volume. As the active length of the machine is known only the height and thickness of the magnet must be selected. To provide adequate space for the cage the magnet height was selected as 26 mm. Table 8 contains the magnet thicknesses (t_m) as calculated with (3) for different PM grades.

Table 8: PM Thickness Sizing for Different Grades

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

All the magnets in Table 8 provide the same B_δ as ICFM utilizes flux concentration to form the poles [6]. It was decided to use a N33 grade PM and to increase the thickness to 6mm to accommodate for loss in magnetic energy due to leakage flux. Figure 9 indicates that the selected magnet volume provides the required energy (red line) at both rated (blue line) and operating (green line) temperature. The intersection points are the operation point of the magnets. The selected PM dimensions provided the required B_δ value.

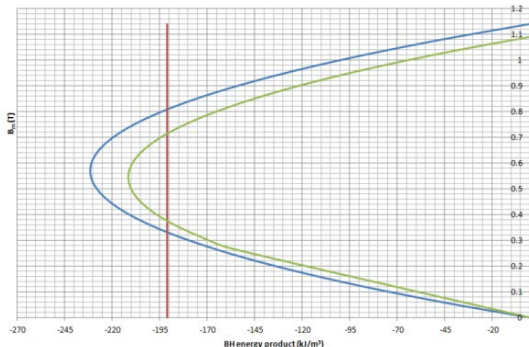


Figure 9: BH energy plot of the selected PM grade

The synchronous torque (T_{syn}) of the PMSM is the sum of electromagnetic torque (T_{em}) and reluctance torque (T_{rel}) components and is calculated using the general torque equation for a PMSM. The torque vs. load angle curve for the PMSM is indicated in Figure 10. The values used to plot Figure 10 is d -axis reactance, $X_d = 29.329 \Omega$, q -axis reactance, $X_q = 85.841 \Omega$ and back-EMF, $E_0 = 186.362$ V. These values were calculated as in [10] and [16].

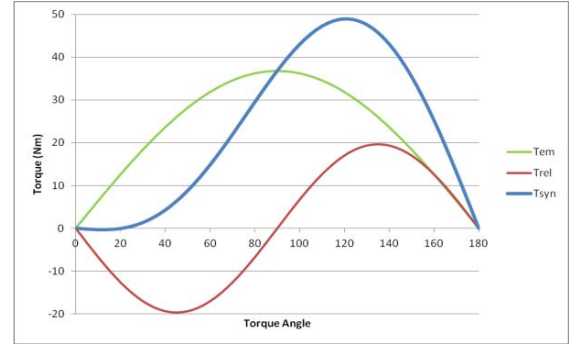


Figure 10: PMSM torque vs. load angle curve.

The breaking torque in an LS-PMSM was first investigated in 1980 by V.B Honsinger and published in his well-known paper [15]. T_m is calculated as a function of slip with (4). Figure 11 contains T_m 's plot with a peak value of ± 14 N.m at $s = 0.045$.

$$T_m = \frac{3p}{\omega(1-s)} \left[\frac{R_1^2 + X_q^2(1-s)^2}{R_1^2 + X_q X_d(1-s)^2} \right] \left[\frac{R_1 E_o^2(1-s)^2}{R_1^2 + X_q X_d(1-s)^2} \right] \quad (4)$$

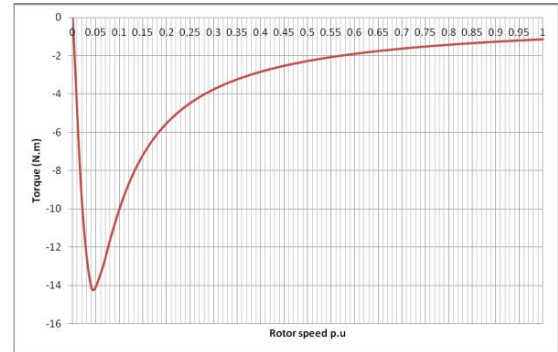


Figure 11: Breaking torque component of the PMSM in the LS PMSM prototype

3.4.2 IM Rotor Component

The first step in the IM design is to select the number of rotor slots (Q_r). As a slot is required above each PM to limit the leakage flux Q_r must be divided by four. For the prototype rotor Q_r is selected as 24 with information gained from [10] and [11]. For ease of manufacturing round rotor slots will be used.

As with the stator the rotor yoke and tooth flux density must be selected to calculate the slot height and width. As the PM span the rotor yoke, the remanence value of the PM is used. This provided a yoke height of 26.63 mm. The maximum tooth flux density occurs on the d -axis and gradually decreases between each tooth towards the q -axis. By selecting the maximum value of 2.2 T the tooth width is

calculated as 7.425 mm. The next step is to determine the remaining dimensions of the round slot as in Figure 12. This is done with the aid of the slot performance factor (λ_{ru}) equation that directly influences the rotor inductance (L_2). By minimizing λ_{ru} , L_2 is also minimized. λ_{ru} is calculated with (5). 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 $\lambda_{ru} = 3.078$. Along with λ_{ru} , the end ring performance factor (λ_{rer}), must be determined before L_2 can be calculated. The dimensions of the end rings were determined as 12 mm by 12 mm providing $\lambda_{rer} = 0.37007$.

$$\lambda_{ru} = 0.47 + 0.66 \frac{b_4}{b_1} + \frac{h_1}{b_1} \quad (5)$$

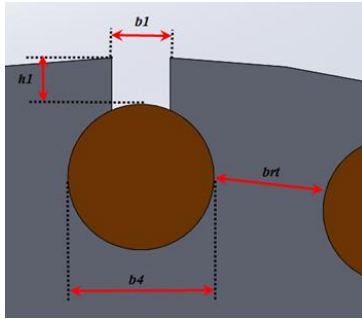


Figure 12: Rotor slot dimensions

To calculate R_2 , the rotor bar and end-ring resistance must be calculated, these values are calculated as 99 $\mu\Omega$ and 3.53 $\mu\Omega$ respectively as in [10, 11]. Once R_2 and L_2 is calculated in rotor reference frame these value must be transformed to the stator reference frame. The transformed values are $R_2' = 2.187 \Omega$ and $X_2' = 1.615 \Omega$ with X_2' being the rotor reactance in the stator reference frame.

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. Table 9 contains the relevant torque and current values. The torque vs. slip curve will be provided in Section 5.

Table 9: Starting and Breakdown Parameters

Parameter	Symbol	Value
Starting Current	T_{start}	55.93 A
Starting Torque	T_{start}	130.67 N.m
Breakdown Torque	T_{bd}	153.59 N.m
T_{db} slip speed	S_{db}	0.511

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 design of LS-PMSM prototype rotor as shown in Figure 13. To ensure that a single lamination can be used two saturation zones are use between the PM and shaft as well as the PM and rotor bar. The original gap between the rotor bar and PM was 1.425 mm. However the PM leakage flux as determined from static FEM simulations was 18% of

the total PM flux. By increasing the shaft diameter with 1 mm the gap was reduced to 0.952 mm resulting in a 10% PM leakage flux which is acceptable according to the PM sizing done in 4.3.1.

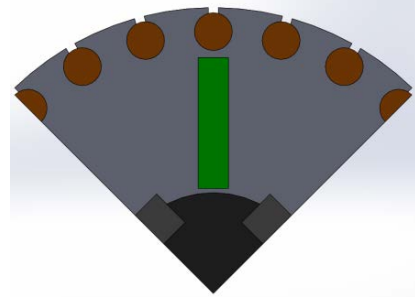


Figure 13: Quarter section of LS-PMSM Rotor

To verify the LS PMSM rotor the air gap flux, rotor yoke and teeth flux density will be used. For the FEM simulation the machine is driven a full load line current ($I_{line} = 10A$) at synchronous speed, Δf between the rotor and stator is zero. Table 10 shows the verification results. The tooth flux density is checked on the d -axis between two teeth.

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 as the flux only passes the air gap over the pole arch. If the flux over the pole is used a lower average flux density value will be obtained. The average flux density over the pole arch is calculated as 0.84 T which is acceptable. Thus the conclusion can be made that the simulation model is an accurate representation model of the LS PMSM machine.

Table 10: LSPMSM Rotor Verification Information

	Rotor Yoke Flux Density		
	FEM	Calculated	% Difference
Tooth 1	2.29 T	2.2 T	4
Tooth 2	2.082 T		6
	Rotor Teeth Flux Density		
	FEM	Calculated	% Difference
	1.208 T	1.17 T	3.2

4. PERFORMANCE PREDICTIONS AND SIMULATIONS

In this section the torque profile and the back-EMF properties of the prototype will be investigated. The calculated and simulation results of the asynchronous and synchronous torque components will be compared. This is done to ensure that the method used to design the prototype is correct. The back-EMF peak value is calculated and determined from the FEMM model. The peak back-EMF value is a key component in calculating the breaking torque of any LS-PMSM.

4.1 Asynchronous Torque Profile

In initial comparisons it was found that ignoring the skin effect on the rotor bars provided a significant

error in torque calculations. Once the effect was incorporated into the design the simulated and calculated torque curves are within acceptable range. Figures 14 provide a comparison between the cage torque with and without the skin effect while Figure 15 contains the revised T_{asy} curve and its components. The comparison between the calculated and simulated T_{asy} curves is given in Figure 16. The simulated results were obtained by using Maxwell[®] RMxpert.

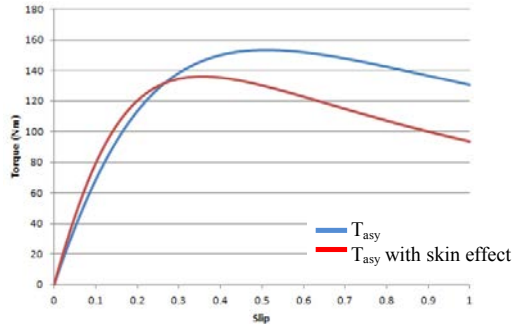


Figure 14: Cage torque with and without skin effect

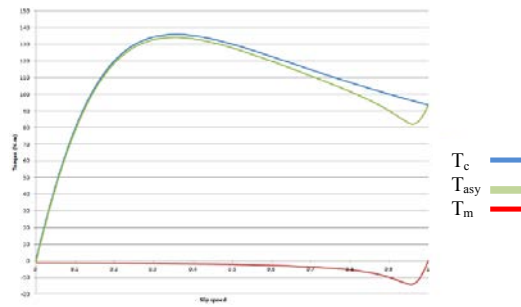


Figure 15: Torque vs. slip of different torque components (including skin effect)

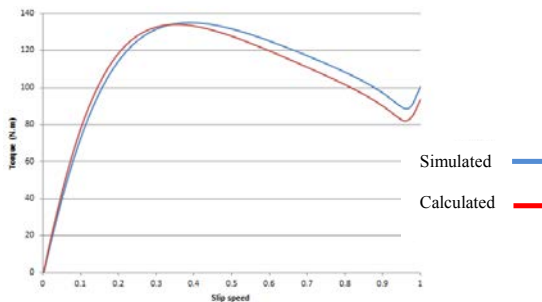


Figure 16: Simulated torque curve vs. calculated torque curve

Table 11: Simulation vs. Calculated Starting and Breakdown Results.

	Calculated	Simulated	% Difference
I_{start}	52.37 A	52.84 A	< 1 %
T_{start}	93.604 N.m	100.94 N.m	7.2 %
T_{bd}	136.01 N.m	135.20 N.m	0.5 %
s_{bd}	0.356	0.4	11%

Table 11 contains the breakdown and starting values for both the simulated and calculated machine.

4.2 Synchronous Torque Profile

The last comparison between the proposed model and the Maxwell RMxpert model is the torque versus load angle plot as indicated in Figure 17. The maximum

calculated torque is 48.9 N.m at $\delta = 120^\circ$ and the simulated maximum torque is 48.39 N.m at $\delta = 119^\circ$. Both the maximum torque and angle are within range of each other.

Table 12 contains the simulated and calculated parameters. The skin effect is included in the calculated parameters as the RMxpert simulation also incorporates the skin effect, thus providing an accurate resistance and inductance comparison.

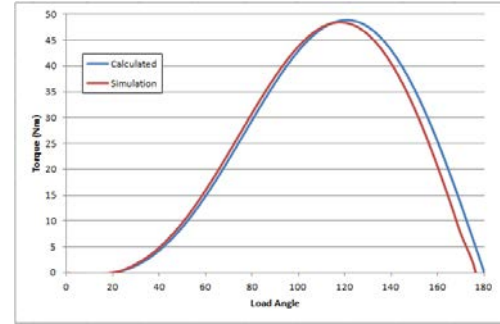


Figure 17: Load angle of simulated torque vs. calculated torque

Table 12: Parameter Comparison

	Proposed Method	Maxwell RMxpert	% Difference
R_l	1.428	1.423	< 1 %
R_2'	1.787	1.679	6 %
X_l	2.387	3.02	26 %
X_2'	1.6473	1.802	9.3 %

All the parameters except for X_l are within 10%, the effect of changing this value so that it is within 10% of the calculated value is investigated. By reducing the X_l the T_{start} and I_{start} increases. Thus the conclusion can be that the simulation package uses a different method to calculate the starting and breakdown values.

4.3 Back-EMF

The back-EMF of an LS-PMSM is a function of the PM flux linkage (λ_{pm}) with the stator coil phases with respect to the rotational speed of the rotor. The flux linkage of each phase is calculated in FEMM, by setting the phase current's value to zero. This simulates the rotor rotating at synchronous speed. To calculate the peak back-EMF value one of the phases must be aligned with the rotor d-axis. Once the flux linkage value of each phase is extracted the RMS back-EMF value is calculated as 186.167 V. This value correlates with the RMS phase back-EMF value calculated in ANSYS Maxwell[®] as 190.19 V.

5. CONCLUSION

The design method presented in this paper proved to present an adequate fit to the simulated results as indicated in Section 4. An efficient design approach for LS-PMSM has been presented in the paper. By dividing the LS-PMSM rotor design into two separate designs i.e. IM and PMSM, the final rotor's performance can be readily represented by using

classical electrical machine theory. To validate the proposed design approach, the calculated results are compared with the simulation results from both a FEM and a commercial design package. A good agreement is achieved.

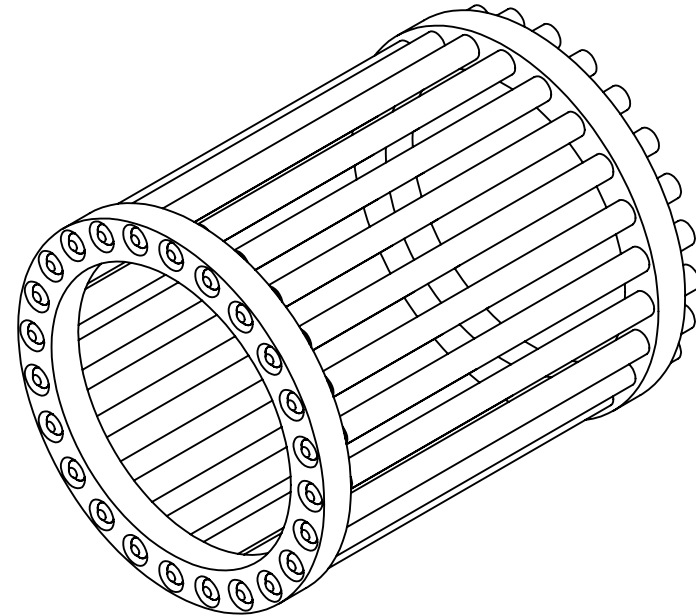
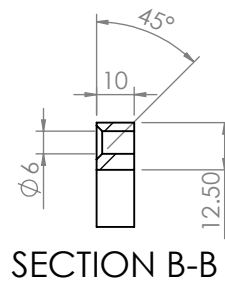
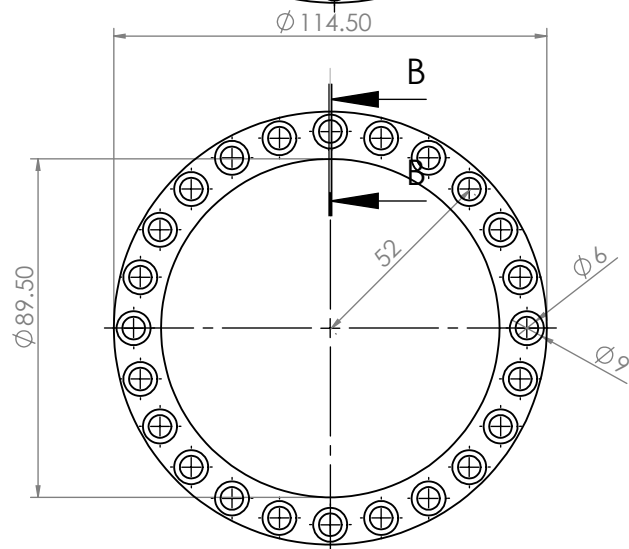
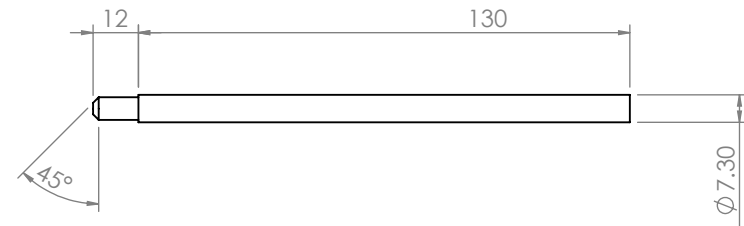
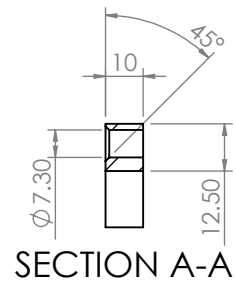
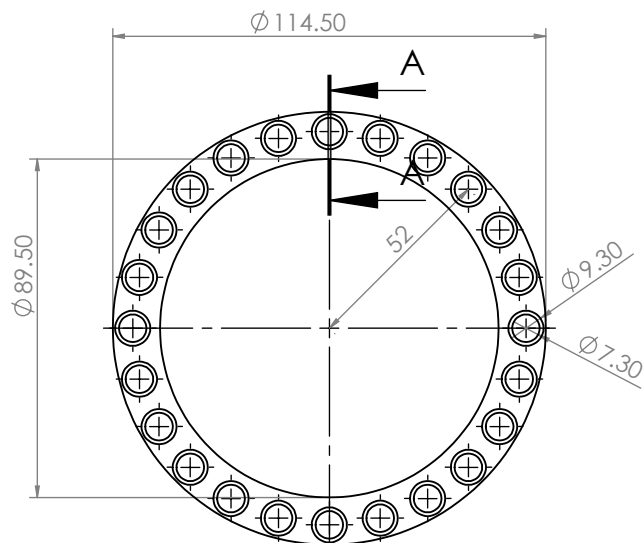
During the initial design of the cage the skin effects was neglected, which caused a poor match between the calculated and simulated cage torque curves. Once the skin effect was incorporated the two torque curves correlated well. In future designs this effect must be incorporated in the design.

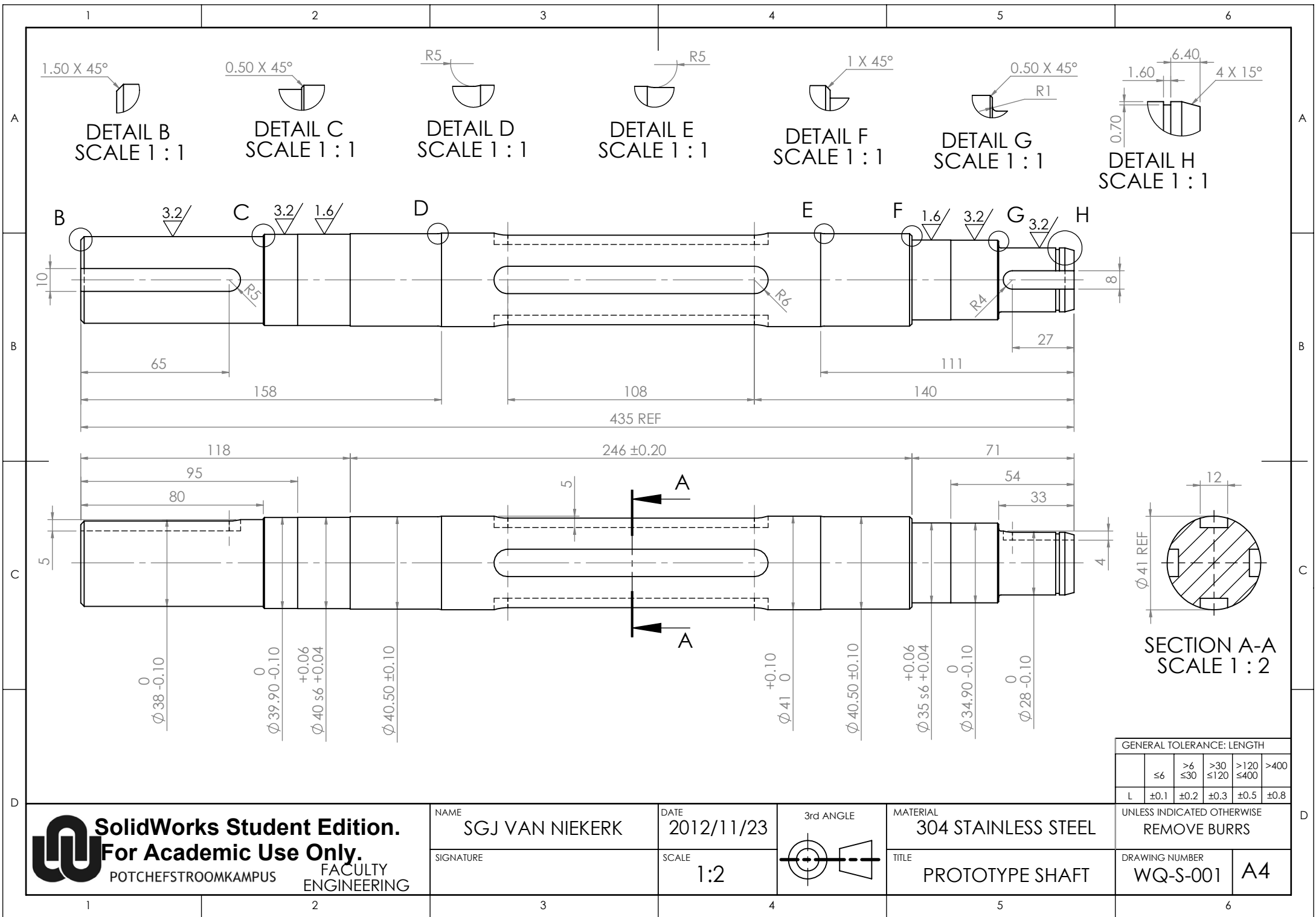
The calculated stator reactance differed greatly from the simulated value. Investigation into other calculation methods must be done and compared with the current method to refine the calculation of this parameter. Once this is done the proposed method in this paper can be seen as a viable design tool.

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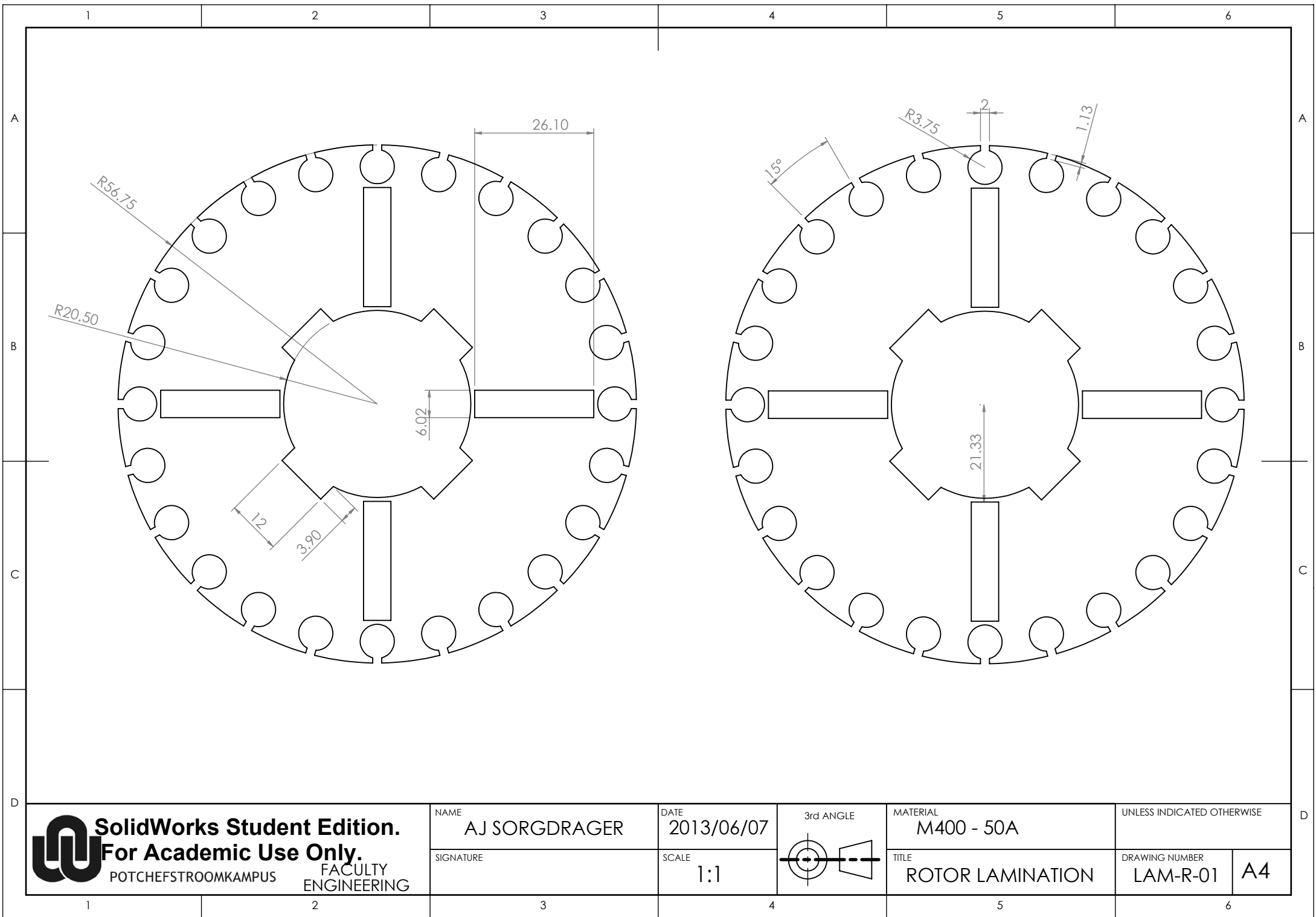
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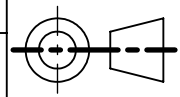
Appendix F: Technical Drawings





GENERAL TOLERANCE: LENGTH					
	≤6	>6 ≤30	>30 ≤120	>120 ≤400	>400
L	±0.1	±0.2	±0.3	±0.5	±0.8



NAME AJ SORGDRAGER	DATE 2013/06/07	 3rd ANGLE	MATERIAL M400 - 50A		UNLESS INDICATED OTHERWISE	
			TITLE ROTOR LAMINATION	DRAWING NUMBER LAM-R-01	A4	

