Impact of Converter Modulation Strategies on the Losses in a Traction Motor

EMIL FJÄLLSTRÖM
Abstract

With an increased demand on high-efficiency electrical machines in traction applications, better understanding in the losses induced by the converter is needed. The harmonic losses from the converter are non-negligible part of the total losses in an induction machine, and these harmonic losses have been studied here with a focus on iron, skin-effect and proximity losses. The main goal of the thesis is to advise how the modulation method can be changed to lower these time harmonic losses in an induction machine. A model of the induction machine was created in the FEM(Finite Element Method) software FLUX.

The iron, the proximity and the skin-effect losses were simulated in the FLUX model with different modulation method to analyse the time harmonic losses due to respective phenomena.

The modulation method at higher operation is square-wave modulation, this is due to that maximum fundamental voltage has been reached. At lower operation, pulse width modulation is used, and the switching frequency can be increased to lower the time harmonics losses in the motor.

The iron losses simulated in FLUX show that the losses increase with frequency even if the motor enters flux weakening operation at higher frequencies.

The proximity phenomena gives rise to a distinct increase of resistive losses in the conductor closest to the air gap in the stator winding.

The skin-effect phenomena affects the losses at square-wave modulation of high frequencies mostly as the time harmonics are larger and of low order.

Keywords: Induction Machine, FEM, FLUX

Sammanfattning

Med en ökad efterfrågan på högeffektiva elektriska maskiner i traktionsapplikationer, måste man ha en högre förståelse i de förluster som induceras av frekvensomvandlaren. Förlosten från frekvensomvandlaren är en icke försumbar del av de totala förlusterna i induktionsmaskinen. Förlosten har här analyserats med fokus på järn-, skineffekt- och proximityförluster. Huvudsyftet med examensarbetet är att ge råd i hur moduleringsmetoden kan ändras för att sänka övertonsförlusterna i en induktionsmaskin. En modell av induktionsmaskinen skapades i FEM(Finita Elementmetoden) mjukvaran FLUX.

Järn-, proximity- och skineffektförlusterna simulerades i FLUX-modellen med olika moduleringsmetoder för att analysera övertonsförluster på grund av respektive fenomen.

Moduleringsmetoden vid högre drift är fyrkantmodulering, det är på grund av att maximal fundamental spänning har uppnåtts. Vid lägre drift har pulsbreddsmodulering använts och switchfrekvensen kan ökas för att sänka övertonsförlusterna i motorn.

Järnförlusterna som simulerades i FLUX-modellen visar att förlusterna ökar med frekvensen även när motorn utsatts för fältförsvarning vid högre frekvenser.

Proximityfenomenet ger upphov till en tydlig ökning av resistiva förluster i ledaren närmast luftgapet i statorlindningen.

Skineffektfenomenet påverkar förlusterna på fyrkantsmoduleringen av höga frekvenser, mest på grund av att övertonerna är större och av låg ordning.

Nyckelord: Induktionsmotor, FEM, FLUX
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6.5 Material set-up ................................................................. 40
6.6 Electrical Set-up ............................................................... 40
6.7 Result .............................................................................. 41
  6.7.1 Operation Point 2 ......................................................... 41
  6.7.2 Operation Point 4 ......................................................... 43
  6.7.3 Operation Point 7 ......................................................... 45
  6.7.4 Operation Point 8 ......................................................... 47
  6.7.5 Operation Point 9 ......................................................... 49
  6.7.6 Operation Point 10 ....................................................... 51
  6.7.7 Operation Point 12 ....................................................... 53
  6.7.8 Analysis of the Operation Points ................................. 55
6.8 Summary ........................................................................... 60

7 Discussion .......................................................................... 61
  7.1 FLUX as a Simulation tool for Induction Machines .............. 61
  7.2 Drawing the Geometry in FLUX ....................................... 61
  7.3 Switching from Steady State Application to Transient Application in FLUX ................. 61
  7.4 Simulation of time harmonic rotor bar losses ...................... 61

8 Conclusions ........................................................................ 63
  8.1 Conclusions from the project ........................................... 63
  8.2 Future work .................................................................... 63
**List of Symbols**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>QQS</td>
<td>Number of stator slots</td>
</tr>
<tr>
<td>QQR</td>
<td>Number of rotor bars</td>
</tr>
<tr>
<td>p</td>
<td>Pole number</td>
</tr>
<tr>
<td>$R_{SCR}$</td>
<td>Rotor short-circuit ring resistance</td>
</tr>
<tr>
<td>$L_{SCR}$</td>
<td>Rotor short-circuit ring inductance</td>
</tr>
<tr>
<td>rpm</td>
<td>Rotations per minute</td>
</tr>
<tr>
<td>$R_S$</td>
<td>Stator winding resistance</td>
</tr>
<tr>
<td>$f_{stator}$</td>
<td>Frequency of the applied voltage or current in the stator winding</td>
</tr>
<tr>
<td>$R_{end}$</td>
<td>Resistance of the end-windings</td>
</tr>
<tr>
<td>$L_{end}$</td>
<td>Inductance of the end-windings</td>
</tr>
<tr>
<td>$k_h$</td>
<td>Material parameter hysteresis losses in iron laminations</td>
</tr>
<tr>
<td>$k_e$</td>
<td>Material parameter excess losses in iron laminations</td>
</tr>
<tr>
<td>$B_{m}$</td>
<td>Amplitude of fundamental flux density</td>
</tr>
<tr>
<td>f</td>
<td>Frequency</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>Conductivity</td>
</tr>
<tr>
<td>d</td>
<td>Iron lamination thickness</td>
</tr>
<tr>
<td>$d_{initial}$</td>
<td>Initial/Real lamination thickness</td>
</tr>
<tr>
<td>$\delta_{skin}$</td>
<td>skin depth</td>
</tr>
<tr>
<td>$\rho$</td>
<td>Resistivity</td>
</tr>
<tr>
<td>$\mu$</td>
<td>Permeability</td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>Permeability in vacuum</td>
</tr>
<tr>
<td>$\mu_r$</td>
<td>Relative permeability</td>
</tr>
<tr>
<td>$B$</td>
<td>Magnetic flux density</td>
</tr>
<tr>
<td>$H$</td>
<td>Magnetic field strength</td>
</tr>
<tr>
<td>$k_f$</td>
<td>Stacking factor</td>
</tr>
<tr>
<td>$J$</td>
<td>Current density</td>
</tr>
<tr>
<td>$R_{coil}$</td>
<td>Resistance of one parallel coil excluding the end-winding resistance</td>
</tr>
<tr>
<td>o.p.</td>
<td>Operation point</td>
</tr>
</tbody>
</table>
1 Introduction

1.1 Background

The electrical machine is an important component in today’s society, without this invention the world would not move nor have any electricity. As this invention is so important, it is of great interest to study it to be able to understand the machine and possibly improve it. The most implemented electrical machine in traction application the last years is the asynchronous induction machine. With studies on the induction machine, one can optimise the design and control to minimise the losses in the system.

The induction machine in traction application is connected to a converter that transfers the electricity to the machine. Using semi-conductors as IGBT or similar, DC voltage is transformed into a switched AC voltage whose fundamental has the expected amplitude and frequency. The downside of this method is that the output from the converter always includes harmonics. These undesired harmonics are called time harmonics and they generate losses in the electrical machine. There are several ways to do these switchings and different methods are also referred to as modulation methods. The losses in the induction machine and the dependency of the used modulation method have been analysed in this project.

1.2 Objective

The goals of the project are listed below:

- Identify the different loss components in the induction motor
- Create a FLUX model of the induction motor
- Validate the FLUX model with measured data
- Calculate time harmonics losses in the FLUX model
- Make comments on the choice of modulation strategies with support from the results

1.3 Outline of Report

- Introduction of the thesis can be found in section 1.
- Previous studies in the field and what can be used from them is presented in the literature study in section 2.
- The program FLUX used for magnetic simulations and how to replicate the results is presented in section 3.
- The created model is verified in section 4 to prove that the performance of the model matches the measurements from the corresponding machine.
- The theory, implementation and alternative approach in calculation of the iron losses are in section 5.
- The theory, implementation and alternative approach in calculation of the proximity and skin-effect losses are in section 6.
- Discussion about positive and negative experience in the project is in section 7.
- The conclusion of the project and how it can be utilised is in section 8.
2 Literature Study

The traction system investigated in this study consists of the converter and the traction motor, more specifically an induction motor.

To supply the traction motor with power, the converter uses PWM [1] to get the desired speed and torque. A traction motor for train application has several operation points, it needs for example to operate at low speed with high torque and high speed with low torque. This puts high demands on the traction motor and the converter. To be able to fulfill the required performance, different PWM methods are used as some methods are better suitable for certain operation intervals [2]. Due to the PWM converter supply, the voltage waveform applied to the stator winding is not sinusoidal, and the harmonics increase the losses in the traction motor. These additional losses are referred as harmonic losses [3, 4, 5, 6, 7, 8].

To calculate the harmonic losses, a model is needed. The model is then to be implemented in the equivalent circuit commonly used for dynamic simulations of an induction machine [9].

In the stator winding, skin effect and proximity effect increase with frequency and can be of significant importance for the calculation of the winding losses [10, 11, 12, 13].

In the rotor bars, the harmonics also give rise to an increase of conduction losses [14, 15].

The iron losses in the rotor and stator and how to model them in the equivalent circuit is explained differently in each article [16, 17, 18, 19]. Some models are more advanced than others, but the basic alternative is a resistance parallel to the magnetization inductance. In previous work, the calculated iron losses had to be multiplied with a factor of 1.5-2 to fit the measured values. One reason for this considerable difference would be the negative effect of manufacturing on the magnetic properties of laminations [20]. The article also presents that the manufacturing process significantly alters the BH-curve of the material. The influence on iron losses caused by the time harmonics is explained in [20, 21]. The principle is that more harmonics create more losses and low Modulation Index, or Switching Frequency can be two causes of that.

The increase of stray losses has also been investigated [22, 23], but the loss increase due to the time harmonics may need further investigation.
3 Simulation of a Inverter-fed Induction Motor in FLUX

3.1 Introduction

FLUX 11.2 is a FEM simulation software for rotating machines [24]. In FLUX for magnetic simulations there are three different applications: Magneto Static, Steady State and Transient Analysis. In this master thesis, only Steady State and Transient Analysis are used and described for induction machines.

Steady state is a simplified simulation to find the steady state properties of a rotating machine. The machine in this application can only be fed with a pure sinusoidal voltage or current, because of this there is no possibility to simulate with time harmonics. The steady state application is good to test the model, to see whether it behaves as expected before the heavier transient analysis is made.

Transient analysis is a time-dependent simulation with a specified time step, can be used to simulate start-up and for calculating of losses at continuous operation including time harmonics. Transient analysis requires more time than steady state simulation. A steady state simulation can take a couple of minutes but a transient simulation a couple of hours to days.

3.2 Pre-simulation

Before the simulation is started, the engineer needs to know what type of simulation he/she will be doing: the starting characteristics or continuous operation at a specific operation point.

If the engineer investigates starting characteristics, then it is less complex. The engineer first defines the model of the machine, does a steady state simulation, deletes the steady state application and then straight on continues with a transient analysis application.

But if the engineer is simulating a machine at a specific operation point, then it is more complex. The reason is that the mesh needs to be identical in both steady state application and transient analysis application. This is because when the steady state analysis is done a transient start-up file needs to be exported to initialise the transient analysis, if the mesh then is different an error will arise. To avoid this, the engineer first defines the model for the machine and does a steady state analysis to see if the machine behaves as it should and then follows the steps explained below.

1. First save the working final design steady state model in a different name.
2. Delete any existing simulation result.
3. Delete the mesh.
4. Delete the infinite box.
5. Save the model and close it.
6. Open FLUX Supervisor and create a new project.
7. Import the geometry that was just saved.
8. Create the infinite box and mesh the geometry.
9. Save the project named “Empty”.
10. Open “Empty” and save it with another name “SteadyState”.
11. Create the steady state application and repeat what was done in the initial working project.
12. Simulate the steady state project and export the start-up file for transient analysis.
13. Close the "SteadyState" project.
14. Open “Empty” and save it with another name “Transient”.
15. Create the Transient application and initialise it with the start-up file.

By following the above steps, an identical mesh will be used in both applications. Each step will be explained more carefully in this chapter in a chronological order.
3.3 Geometry

When a new project is opened in FLUX, the "sketcher" is initially opened. The sketcher is a fast user-friendly tool to draw the machine, but if there is a wish to parametrise the geometry the sketcher should be avoided. This is because when a transformation is used in the sketcher the parameters will be removed and recalculated to values.

To parameterise a geometry is an advantage. If one length in the model is wrong the parameter can be changed and the whole model adapts, and if another motor needs to be simulated only the parameters need to be changed. If no parameters are used and a length needs to be changed, the whole design needs to be redrawn. To parameterise the geometry, the design then needs to be made in the main window.

To minimise the computational time, symmetry can be used. If the number of stator slots, \( QQS \), and rotor slots, \( QQR \), are dividable with an integer as for example the pole number, \( p \), two or four, then only one of these pieces of the machine need to be simulated. In this example, the \( QQS \) and \( QQR \) are dividable with \( p \) equals four and then only one-quarter of the machine needs to be simulated.

3.3.1 Stator

In the construction drawing of the stator for the machine all dimensions for the stator exist, all of these have to be expressed by a parameter to make the geometry parametrised. If one parameter in the model is later changed, the model will then automatically adapt.

An example of a drawn model is displayed in figure 1.

![Figure 1: Example of a stator model.](image)

A zoom of the stator slots is displayed in figure 2.
For the stator slot, two cases can be considered. In the first case all conductors can be drawn as in figure 2. In the second case, instead of drawing all conductors the whole slot can be set as a coil see figure 3. This action with a simplified model can be used if the detail of each conductor is of no interest and it also allows the mesh to have fewer elements. Fewer elements in the mesh have two main advantages. First the computational time is less and second the model requires less memory and can then be simulated on a smaller computer. For a simulation of proximity and skin-effect losses, the conductors need to be drawn in detail.

![Figure 2: Example of a model of a stator slot.](image)

![Figure 3: Simplified model, not used in this project.](image)

### 3.3.2 Rotor

The principle with parametrisation is the same for the rotor as for the stator, and the geometry can be found in the construction drawing of the rotor. A drawn model of a rotor is displayed in figure 4. The model in this project has other rotor topology investigated in [14].
3.3.3 Complete Machine

Up to this point only a quarter of the machine is drawn. To set up so it will be simulated as a whole machine the function periodicity is used, see figure 6 and 7.
The geometry still needs boundary conditions to the outside world, so an infinite box is introduced outside the stator, see figure 8.
3.3.4 Mesh

The built-in automatic mesh in FLUX is good and can be used for performance simulations. However, for the proximity and skin-effect simulations the density of the mesh in the conductor needs to be evaluated, this will be explained later in section 6.4. The automatic mesh is shown in the figure 9.

3.4 Steady State Simulation

The steady state application is defined for one frequency and cannot be used for simulation of harmonics. When the application is started the parameters for the steady state need to be set up. These are the mechanical properties for the steady state, the materials properties for the different parts of the machine, the electrical properties of the machine with the applied current or voltage.
3.4.1 Mechanical set-up

The machine has two mechanical parts with different properties, the rotor and the stator. The settings for the mechanical set-up are displayed in figure 10.

The stator is fixed, and all regions from the middle of the air gap and out belong to this mechanical part.

The rotor is moving, and all regions from the middle of the air gap to the centre belong to this mechanical part. In the rotor mechanical set-up, the slip also needs to be set as it is an induction motor, see equation 1. There is also a possibility to choose inertia and FLUX will warn if no inertia is set, but it is not a must to calculate with inertia and in this project no inertia was used.

\[
slip = \frac{f_{\text{stator}} \frac{2}{p} - \text{rpm}/60}{f_{\text{stator}} \frac{2}{p}}
\]  

Equation 1

Figure 10: Mechanical settings for steady state.

3.4.2 Material set-up

The machine consists of four different materials, whereas three of them need to be chosen. It is the shaft iron, the iron lamination, aluminium rotor-bars and the copper winding.

The winding is not set to be copper as it does not matter for the calculation because the resistance of the winding is set separately in the electrical set-up.

The shaft iron and the iron lamination material properties can be imported with the FLUX material manager. In this project, the shaft iron is defined as FLU_STEEL_1010_XC10 and the iron lamination is FLU_M600_50A, which both exist in the material manager.

When simulating in the steady state application an important assumption for the iron material needs to be made so the performance of the model matches the measurements of the machine. If the simulated machine is powered with a current source the magnet field should be assumed to be sinusoidal, and if it is a voltage source the B-field should be assumed sinusoidal. If this assumption is not made, the machine may not produce expected torque. This was an issue that was faced in this project for lower frequencies, but the machine still behaves as expected at a higher frequency at nearly twice the base speed.

The material for the rotor bars is Aluminium. This material needs to be created manually. The chosen material properties of Aluminium are as follows: permeability is equal to 1 and isotropic resistivity equal to 0.4676E-7 [Ωm]. The resistivity is high due to the operation temperature of the machine which is 150 [°C]. The resistivity is defined at 20 [°C] and then it needs to be recalculated with the use of the temperature coefficient with a temperature rise of 130 [°C]. It then matches the mentioned resistivity at 150 [°C].
3.4.3 Electrical set-up

In figure 11 the electrical circuit for the model is displayed. The electrical circuit is a coupling of an electrical simulation and a FEM simulation. The three cylinders to the right symbolise the three coils in the stator slots, the coil resistance is calculated analytically as explained on page 172 in [8]. The three inductances are the inductances in the end-windings for each phase. There are also two current sources which act as the applied current. The reason why it is only two is that FLUX calculates the third current automatically, to balance the system according to Kirchhoff’s current law.

In the bottom, there is one component left, and that is the squirrel-cage. In the squirrel cage three parameters need to be given: the number of rotor-bars, the inductance of the short-circuit ring, $L_{\text{SCR}}$, and the resistance of the short-circuit ring, $R_{\text{SCR}}$. The number of bars is known from the geometry, the total number of bars QQR divided by in this example the symmetry of four gives QQR/4 bars. The short-circuit inductance and resistance are calculated analytically [8], but these values will be changed later to trim the performance of the machine.

The electrical circuit represents the whole machine because of the symmetry and periodicity defined in the geometry. This was also checked by comparing the performance of a simulation of the whole machine with a divided machine and the performance matched.

The circuit is now defined, but the electrical set-up in the geometry still needs to be done. The induction machine used in this project has a dual-layer winding with shorted coil pitch of two slots, see figure 12.

Each of the groups in figure 12 above are defined as one region in the geometry, region $A^{-}$, $B^{+}$, $C^{-}$ and $A^{+}$, see figure 13. In the positive region, the current has a positive orientation and in the negative regions the current has a negative orientation. Region A is connected to coil A, region B is connected to coil B and region C is connected to coil C. In the electrical set-up for these regions FLUX asks for turns per phase. In the geometry, the number of conductors per phase is 40, but this machine has parallel coils, and to take that into account the turns per phase is cut in half [8]. In the region $A^{-}$ and $A^{+}$ only half of the coil slots is in model because of the symmetry, also here the geometrical number of conductors is cut in half. The settings are displayed in figure 14.

An alternative method to set the negative region is to have a negative orientation of the current is to change the direction of the component in the circuit. In figure 11 above each component in the circuit
there is a small square. This square represents the direction of the component. So instead of changing the orientation in the settings for region C_MINUS the component C_MINUS in the figure could be rotated, and the current would flow in the other direction.

Figure 13: Coil regions in the geometry.

Figure 14: Settings for the coil regions. Red: Choice of component, Blue: number of conductors and Green: orientation of current

The rotor-bars in the geometry also need to be connected to the electrical circuit. For each region of the rotor-bars the region is set to a solid conductor and can then be connected to respective squirrel-cage rotor bar, see figures 15 and 16 below.
3.4.4 Parametric Calculation of the Rotor Short-circuit Resistance and Inductance

After the first steady state simulation the torque produced by the machine might not match the desired torque at all. In this project the simulated torque was 16 Nm when desired was around 1000 Nm.

Two parameters that can be adjusted to find the matching torque are $L_{SCR}$ and $R_{SCR}$ from the electrical set-up. The relationship between the torque and the parameters is not linear so a parametric simulation can be done numerically in FLUX to find the matching values. This is done by creating a solving scenario that changes these two parameters within a user defined interval, see figure 17.
The result can then be plotted in a 3D graph, and the parameters can be identified. The parameters might have a good match at one operation point, but not as good at another one so a compromise needs to be done to find the overall best match of the parameters. The result for one operation point is presented in figure 18 and 19.
Analysing the result in figure 18 and 19 the best fit to desired values of torque and voltage of $L_{SCR}$ and $R_{SCR}$, are $R_{SCR} = 5E-8 \, \Omega$ and $L_{SCR} = 2E-7 \, H$, respectively.

### 3.4.5 Export of Transient Start-up file

As described in the Pre-simulation, the transient start-up file can be exported when the engineer has decided that the final design is achieved. The engineer needs to follow the steps described and re-simulate the steady state scenario. When the re-simulation is done, it is time to export the transient start-up file, see figure 20. The start-up file minimises the time for the transient simulation to reach steady state, see also the description in section 3.2.

![Transient start-up file export](image)

**Figure 20:** Transient start-up file export.

### 3.5 Transient Simulation

The transient analysis application is used to simulate transient behaviour, such as start-up characteristics and torque over time. The transient application is also the preferable for simulation of losses and can include harmonics or external waveforms. The mechanical, material and electrical set-ups need to be done again for the transient analysis.

Initialisation by file is chosen when starting the application. The initialisation by file minimises the computational time by using the result from the steady state start-up file as initial values. The simulation mode in the transient application is “Imposed Speed”, this means that the shaft rotates with a user defined constant speed.
3.5.1 Mechanical set-up

The mechanical set-up is done as described for Steady State with one difference. The slip is not put in as a parameter. In the transient analysis the rotations per minute, rpm, of the shaft is used in the “Imposed Speed” option. The rpm can easily be calculated as the slip is known. In this project, the rpm value used is from measurements.

3.5.2 Material set-up

For the material set-up there is only one difference from the Steady State set-up and that is that the assumption of sinusoidal magnet field or B-field does not need to be made.

3.5.3 Electrical set-up

The electrical set-up has one change from the steady state model. In the electrical circuit, the input to the current sources should be the exact curve of the current. Before only the rms value of the sinusoidal current was set. Now it is the waveform expressed with sin or cos with the magnitude and phase-shift. Observe that the current waveform will preferably match the magnitude of the steady state start-up file. The steady state start-up file contains the first solved step for the transient simulation, a difference in the current waveform would result in more transients before the simulation converges.

3.5.4 Tabulated Input

If the input waveform is non-sinusoidal, it is preferable to use a tabulated input, instead of writing the formula for the current/voltage described in the electrical set-up. The waveform needed can be created in MATLAB[25] and saved in a .txt file with two columns, column one being time and column two being the value of the waveform. If the waveform has in total 20 periods and the FLUX transient simulation is 30 periods FLUX will not give a warning. The simulation will continue to simulate with the last value of the waveform as a magnitude of a DC current/voltage. To avoid this, the length of the imported waveform needs to be longer than the simulation time. Two waveforms need to be imported, one for each current/voltage source in the electrical circuit with the corresponding phase-shift.

The advantage of using the tabulated input is that the formula for the waveform with its entire harmonics does not need to be described with a single formula. The fundamental is easily described with the formula, but the harmonics are more complex as they have other phase-shift than the fundamental. The field in which the formula is entered has a restriction of a number of characters; this also limits the use of a more complex formula.

The instructions for tabulated input are displayed in figure 21.

![Figure 21: Settings for tabulated input. To open the right window press the arrow in the red circle.](image)
3.6 Calculation of Losses in Post-Processing

There are three losses that can be calculated in the model, conductive losses in the windings and the rotor bars and also the iron losses.

3.6.1 Conductive Winding Losses

The losses in the windings can be calculated with the use of a sensor in the stranded coil. The result from the sensor is first presented after it has been evaluated. The evaluation can be done by right clicking on the sensor and pressing evaluate sensor. The settings for the stator coil sensor are displayed in figure 22.

![Figure 22: Settings for stator coil sensor.](image)

3.6.2 Conductive Rotor Bar Losses

The losses in the rotor bars can also be calculated with a sensor in each region of the rotor bars. The losses calculated will be only for a quarter of the machine, and the total losses are $p$ times larger with $p = 4$. The sensors here also need to be evaluated to present the result. The settings for the rotor bar sensor are displayed in figure 23.
3.6.3 Iron Losses

The model used for iron losses is the Bertotti iron loss model [26], this will be explained in detail in section 5.
4 Verification of the Simulation Model with Measurements

4.1 Introduction

The FLUX FEM model described in this project is a model of an existing induction traction motor at Bombardier Transportation. In this chapter, the model will be verified by using measurements on that motor. Measurements of the motor have been analysed to use the same measured input for the model as for the motor in the lab to compare the output torque and speed.

In section 4.2 - 4.3 current sources were used, and the model was applied with the same fundamental current as the existing motor.

In section 4.4 - 4.5 voltage sources were used to provide the same test scenario as for the existing motor. During the tests, the fundamental of the voltage is set to analyse the phase current in the motor.

4.2 Performance versus Speed

The traction motor is designed to deliver a certain torque at a certain speed. In figure 24 the performance is compared between the simulation results and the measurements of the motor.

![Comparison Between Measurements and Model](image)

Figure 24: Comparison between the motor measurements and the model simulation results. The arrows show the operation points analysed in section 5 and 6.

At the operation points, the model deviates from the measurements in performance but with acceptable accuracy. For example, at stator frequency 21, 40, 56 and 179 Hz; the deviation is 2.58%, 4.37%, 1.02% and 1.64% respectively.

4.3 B-field in Air Gap

The maximum value of the fundamental flux density in the air gap for three of the four chosen operation points is listed below in table 1. The magnitude of the flux density normal component of around 1 p.u. is consistent with what the motor is designed to have for the two first operation points. The third operation point is at flux-weakening operation. It can be observed in the table that the flux density is of smaller magnitude.
Table 1: B-field in the air gap.

<table>
<thead>
<tr>
<th>Operation point</th>
<th>(B_{\text{air}})</th>
</tr>
</thead>
<tbody>
<tr>
<td>21 Hz</td>
<td>1 [p.u.]</td>
</tr>
<tr>
<td>40 Hz</td>
<td>1 [p.u.]</td>
</tr>
<tr>
<td>179 Hz</td>
<td>0.43 [p.u.]</td>
</tr>
</tbody>
</table>

4.4 No-load Test
A no-load test has been performed on the model where the shaft is rotating synchronously with the magnetic field created by the stator. The measurements and the FEM result are presented in figure 25.

![Figure 25: Comparison between motor and model for no-load test.](image)

The no-load test was initially simulated in the steady state application. The magnitude of the current for a higher voltage is much higher compared to the measurements. To get a valid simulation result, a transient analysis was made. The result from the transient simulation follows the measurements and does deviate as the result from the steady state simulation. The difference was a higher flux density in the motor with the steady state application than in the transient application. The deviation started in the steady state application when the flux density in the iron lamination reached saturation level, this is probably because of the assumptions made in the steady state application.

The model gives acceptable prediction of the motor behaviour during the no-load test.
4.5 Locked Rotor Test

The locked rotor test is done when the shaft is not rotating at all, but the magnetic field is still rotating in the stator. The result of the comparison between the measurements and the model can be seen in figure 26.

![Locked Rotor Test of Simulation Model vs Measurements](image)

Figure 26: Comparison between motor and model for Locked rotor test.

The steady state model does not follow the curve of the locked rotor test at any point that can be seen in the figure.

The rotor temperature is suspected to be the reason why the predicted current values differ so much from the measurements. The model created in this project is a model of a warm traction motor in operation with a rotor and stator temperature of 150°C. Even in a no-load operation the external fan is controlled to keep that temperature of the motor. The locked rotor test is done with a cold motor with an initial test temperature of 20°C. The resistance and the inductance of the motor are temperature dependent and the resistance increases with temperature. At cold operation, one could then assume that the resistance should be lower and then also allow a higher current to pass through the motor.

To simulate a cold motor the parameters used in the model were recalculated to fit an operation temperature of 20°C. The parameters that were changed are: Aluminium resistivity for the rotor bars, $R_s$, $R_{SCR}$ and $L_{SCR}$.

4.6 Summary

The calculation well agree with the measurements. The model in FLUX passes the no-load test and the locked rotor test. The results from the transient model are more accurate when the motor is under larger stress than in the steady state model and iron in saturation is present. With the acceptable performance
in both no-load test and locked rotor test one can conclude that the model in FLUX delivers acceptable performance.
5 Iron Losses

5.1 Introduction

In the electrical machine, iron is used as a magnetic conductor as it has better magnetic properties than air. The magnetic field in the iron changes direction all the time as the applied current is an alternating current. This causes losses in the iron in the form of heat. A number of action has been taken to reduce these losses. One of the action is that the iron is divided into thin sheet called laminations, to reduce induced current in the iron. The Iron lamination is an alloy. Different alloys have different magnetic properties, and some alloys are more suitable for different tasks.

The iron material used in this induction machine has the same magnetic property as Cogent M600-50A [27]. The iron material data used in this report are from Cogent.

In this section, the calculation of iron losses in the induction motor will be analysed.

5.2 Theory

There are several iron loss models available for estimation of the losses, a comparison is presented in [28]. The model was simulated in FEM software - FLUX. There are two different options available in the software for calculating the iron losses: Bertotti model [26] and Loss Surface model [29]. The Bertotti model was used in this project.

Calculation of iron losses for a motor is normally far from accurate. This is due to numerous assumptions made in the equations and the algorithms so they would be reasonable to use.

The foundation for the iron loss model is the measurements of the iron properties in the Epstein frame. Here the material M600-50A is used, it has measurements for only 50 Hz. When adapting the equations to these measurements they will only be valid at 50 Hz. At all other frequencies the assumption is made that magnetic behaviour is similar, however, it is incorrect.

When one measures magnetic properties of iron lamination the most common way of doing it is to use the Epstein frame[30]. The measuring set-up with the Epstein frame is geometrically very different from the iron lamination geometry in a stator and rotor. It is not correct to presume that the magnetic behaviour is the same in any geometry.

During manufacturing of a motor the shape of the stator and rotor is punched from the lamination sheet. This process damages the iron and changes the magnetic property of the iron. In the next manufacturing process the laminations are stacked and welded together. This also damages the iron lamination and changes the magnetic property of the iron lamination. The welding is in principle a magnetic short-circuit for the eddy currents. It is interesting to know that the welding together with the punching increases the actual iron losses in the motor up to 50 % [20].

5.2.1 Iron loss model for FLUX

The Bertotti iron loss model has three parts: the hysteresis losses, the eddy current losses and the excess losses. In the same mentioned order these parts can be seen in both equations 2 and 3. The Bertotti model is based on material properties of the iron, and two parameters need to be identified, see \( k_h \) and \( k_e \) in equation 2. This equation is the average loss model. The manufacturer of the iron has provided a data-sheet with average losses for different magnitudes of the flux density at 50 Hz [27].

\[
dP_{ave} = k_h B_m^2 f + \frac{\pi^2 \sigma d^2}{6} (B_m f)^2 + k_e (B_m f)^{3/2} \cdot 8.67 \quad [W/m^3] \quad (2)
\]

In the equation, \( k_h \) is a material parameter, \( B_m \) is the amplitude of the fundamental flux density, \( f \) is the fundamental frequency, \( \sigma \) is the conductivity of the iron, \( d \) is the lamination thickness and \( k_e \) is another material parameter.

In this case, measurements for only one frequency are given, therefore it is easier to determine the two material parameters needed for the equation. The method to find these parameters is numerical curve-fitting, it was done in MATLAB. There are two ways to do so in MATLAB. First method involves newer versions of MATLAB, where one can use MATLAB’s own toolbox for the curve-fitting. Second method involves an older version of MATLAB, where two for-loops can be programmed to change \( k_h \) and \( k_e \), until the smallest error has been found between the fitted curve and the measurements.
Two for-loops were implemented in MATLAB and the two parameters were found, see figure 27. The parameters estimated are $k_h = 174.5$ and $k_e = 2.699$. The measurements are given in $W/kg$ not $W/m^3$ - the dimensions of the Bertotti model. When the curve-fitting is completed, the equation 2 needs to be divided by the density of the iron material. The density of the iron material M600-50A is equal to 7900 $kg/m^3$.

![Graph of losses vs. B-field](image)

Figure 27: Curve-fitting for material parameters for Bertotti iron loss model.

The FEM software FLUX does not use equation 2 in the transient analysis as it is an average model. FLUX uses the direct model see equation 3 for the transient analysis. One reason is that the fundamental flux density needed for the average model is not known as the transient simulation only calculates the difference between the time steps. Therefore, the equation 3 is ideal, as the flux density derivatives can be calculated with the difference between the time steps.

\[
dP(t) = k_h B_m^2 + \sigma \frac{d^2}{dt^2} \left( \frac{dB}{dt}(t) \right)^2 + k_e \frac{d^2}{dt^2} \left( \frac{dB}{dt}(t) \right)^{3/2} \quad [W/m^3]
\]  

(3)

5.2.2 Alternative Iron loss Approach

As it was mentioned before, the manufacturer has provided measurements for only 50 Hz for material M600-50A, so the parameters would only be valid for 50 Hz. Material parameters with a frequency dependency can be created to make the parameters more accurate at other frequency than 50 Hz. This can not be done by observing the measurements of M600-50A as only measurements at 50 Hz exists, but in the same material series there are several other similar materials with measurements up to 2500 Hz. One can also assume that the material parameters for these other materials have the same frequency dependency as for M600-50A.

In the following example, the assumptions are also made that the excess losses are equal to zero. This is because the frequency dependency for the excess parameter does not have the same characteristics for the different iron materials. When the excess parameters were assumed to be equal to zero the hysteresis parameters had the same frequency behaviour for the different iron materials.
For higher frequency of the flux density the induced eddy current also has higher frequency, this observation leads us to the fact that skin-effect in the lamination needs to be considered. As explained in section 6.2.2 the skin-effect makes the current density to increase in the area closer to the surface of the material. Considering the skin-effect the iron losses induced by the eddy currents decrease. If the current density increases at the surface and the centre of the lamination gets less utilised the effective resistivity of the iron then increases. The resistivity increases whilst the conductivity decreases. If one observes equation 2 the eddy current iron losses are proportional to the conductivity, herewith decrease in conductivity is also decrease of iron losses caused by the eddy currents. This can also be taken into account if one creates a function for the parameter \( d \) with the relation to the skin depth. When the skin depth equals to less than half the lamination thickness, \( d_{\text{initial}}/2 = 0.50/2 = 0.25 \text{ mm} \), then \( d \) will be replaced with the value equal to the double skin depth \( \delta_{\text{skin}} \).

The equation for the skin depth is the following 4.

\[
\delta_{\text{skin}} = \sqrt{\frac{\rho}{\mu_0\pi f}} \quad [\text{m}]
\] (4)

Where \( \rho \) is the resistivity of the iron, \( \mu \) is the permeability of the iron and \( f \) is the frequency of the eddy currents. The difference with the section 6.2.2 is that here \( \mu \) is not equal to \( \mu_0 \), however, \( \mu = \mu_0 \cdot \mu_r \). \( \mu_r \) is the relative permeability of the iron and it depends on the B-field and the H-field, see equation 5.

\[
\mu_r = \frac{B}{H \mu_0}
\] (5)

The B-field and the H-field curves and their relation can be found in the material BH-curve, see figure 28.

Figure 28: BH-curve for Cogent M600-50A[27].

The resulting relative permeability for the material then follows the curve in figure 29.
One aspect that is not taken into account in the skin depth formula at this point is the resistivity and its temperature dependency. The resistivity is defined at $20^\circ C$, but the motor is operating at $150^\circ C$ with a possible higher resistivity. The previously mentioned fact has not been dealt with in this project, depending on material this phenomena might affect the results and has to be taken into account if one intends to use this method.

When all relationships are determined, the method should be implemented in the curve-fitting to find the desired material parameter $k_h$. The implementation conditions for the lamination thickness can be seen in equation 6.

$$d = \begin{cases} 
    d_{\text{initial}} & \text{if } \delta_{\text{skin}} > d_{\text{initial}}/2 \\
    2 \cdot \delta_{\text{skin}} & \text{if } \delta_{\text{skin}} < d_{\text{initial}}/2 \end{cases}$$

(6)

The frequency dependency from the hysteresis parameter can be seen in figure 30. The equation follows the relation $k_h = k_1 + k_2 \cdot f + k_3 e^{-f/k_4}$, which was determined by studying the curve shape.
Figure 30: The frequency dependency of the hysteresis parameter.

This can then be used to calculate the harmonic iron losses in a different part of the motor. In figure 31 the flux density has been taken from the FEM transient simulation by inserting a sensor in one stator tooth.

Figure 31: Flux density in stator tooth. Blue: sinusoidal applied current. Red: Square-wave.
By calculating the FFT of the curve in figure 31 the harmonic content can be analysed, see figure 32.

![Graph of FFT of flux density in stator tooth. Blue: sinusoidal applied current. Red: Square-wave.](image)

**Figure 32:** FFT of the flux density in the stator tooth. Blue: sinusoidal applied current. Red: Square-wave.

The flux density frequency components in figure 32 can be inserted in equation 2. Further on the iron losses for each harmonic can be calculated at that point with the use of the frequency behaviour of the hysteresis parameter $k_h$. The losses between the sinusoidal and the square-wave case can then be compared.

The result of the comparison shows that with square-wave there is a 12.44 % increase of iron losses in this point of the tooth, compared to sinusoidal case.

The two curves in figure 31 can also be used in the direct Bertotti equation, see formula 3. The frequency parameter in that equation is the fundamental frequency. The result of the direct Bertotti equation shows that the square-wave case has 14.46 % higher losses compared to the sinusoidal case.

### 5.3 Implementation in FLUX

In order to calculate the iron losses in FLUX, the Bertotti iron loss model is used as mentioned before in this section. The calculation of iron losses in FLUX is done in post-processing tool, which means that the transient simulations need to be finished before the iron losses can be calculated.

When giving the input data for the iron calculations, there are three tabs that are important: the coefficient tab, the geometry tab and the time tab.

The coefficient tab is where all the parameters/coefficients are set for the Bertotti model, see table 5.3 with all the parameters used in these calculations.

In the geometry the user can choose in which region one wants to calculate the iron losses, for example in the stator tooth or stator yoke.

The time tab is where the time intervals of the calculation are set. It is important that the time interval is equal to exactly one period. That is how FLUX calculates the fundamental frequency for the Bertotti equation in formula 3, if the interval is wrong the result is wrong. If there are too many time steps in the desired interval, the post-processing may take up to several hours.
<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k_h$</td>
<td>174.5</td>
<td>Hysteresis material loss parameter</td>
</tr>
<tr>
<td>$k_e$</td>
<td>2.669</td>
<td>Excess material loss parameter</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>3333333</td>
<td>Conductivity of the iron</td>
</tr>
<tr>
<td>$d$</td>
<td>0.0005</td>
<td>Lamination thickness in meters</td>
</tr>
<tr>
<td>$k_f$</td>
<td>0.94</td>
<td>Stacking factor</td>
</tr>
</tbody>
</table>

Table 2: Coefficient tab for Bertotti iron loss calculation in FLUX

5.4 Results

In figure 33 the iron losses during different operation points are displayed. The iron losses in FLUX are compared to sinusoidal applied current and to the iron losses when the measured current is used with harmonics. Additionally to the previous iron losses the result is also compared to another non-FEM analytical program.

![Figure 33: Iron losses for the motor are compared with current including harmonics and excluding them, as well as to an analytical program.](image)

One can observe in the figure that the increase of losses due to harmonics is nearly constant through the whole motor operation range. The iron losses in the motor increase with frequency, it can also be observed in the figure.

The analytical program does not behave as the FEM model. The sinusoidal losses increase in a level with the FEM harmonic losses at lower frequencies and at higher frequencies the iron losses decrease. The reason why iron losses begin to decrease over 100 Hz is because the motor at this point has passed a base speed and works at flux weakening. This means that over 100 Hz the flux in the motor starts to decrease more and more with an increased frequency; explaining the decrease of iron losses in the analytical program.

The increase in percent of the iron losses due to harmonics is presented in figure 34.
The increase of losses is nearly constant with the FEM model, but the sinusoidal iron losses are smaller at lower frequencies, consecutively making the increase factor higher. The factor then decreases as the sinusoidal iron losses increase and the proportion between the harmonics and the sinusoidal losses expands.

For the operation at 56 Hz with under-modulation, sinusoidal operation was compared to simulation results with the measured current using PWM modulation method. It was also compared to cases with an increased switching frequency equal to 1 kHz and 2 kHz. The purpose of this comparison was to identify how do the losses decrease with the higher switching frequencies. The result of this investigation can be seen in figure 35.
As one expects the iron losses caused by the harmonics decrease with an increased switching frequency. If one increases the switching frequency from around 550 Hz to 2 kHz, the harmonic iron losses are lowered to nearly half.

5.5 Summary

The iron losses were simulated in the FEM software FLUX. There were many assumptions made for these simulations in FLUX, and the actual iron losses are probably quite different from those values. Simulation of iron losses in FLUX can be used to analyse whether one set-up/geometry gives lower losses than another set-up/geometry, however, the result from such simulation does not reflect the losses of a real life motor. The simulated result can give a hint about the magnitude of the iron losses. In reality one needs to multiply the simulated iron losses with a factor of 1.5-2 to get more accurate result. This factor is the result of the before mentioned manufacturing damage which causes an increase of the iron losses.

The results of the calculations in this section show that an increase of switching frequency would decrease the iron losses, example for a fundamental frequency of 56 and a increase of switching frequency from around 550 Hz to 2 kHz would lower the losses to nearly half.
6 Proximity and Skin-effect Losses

6.1 Introduction

Proximity- and skin-effect are two physical phenomena that increase the losses in the winding. The combined loss increase form these two phenomena are studied in this section.

The proximity losses are eddy-current induced in the conductors from the crossing slot leakage magnetic field. The eddy-currents displace the current density in a conductor, what results in a non-uniform current density with higher resistive losses.

The skin-effect losses are connected to the applied current of the motor. With an increase in frequency, the current density shifts more and more towards the surface of the conductor.

6.2 Theory

6.2.1 Proximity Losses

With increased slot leakage flux density, the losses increase, see figure 36. There is low flux density in the upper stator slot, and the current density is nearly uniform. The closer to the air gap the higher slot leakage flux density and the losses are greater. Whilst the non-uniform current density is observed for the conductors close to the air gap. The conductors seen in the figure have the same current, if the current is uniform the losses are the same through whole cross-section of the conductor. When the current is non-uniform in conductors close to the air gap, the current density is close to zero at one side and twice the normal current density at the other side. The resistive losses are proportional to current in square, so the losses in conductors close to the air gap are much higher than in the last conductors situated higher up in the slot.

![Figure 36: Example of a FEM simulation of proximity effect](image)

6.2.2 Skin-effect Losses

When a non-DC current flows in a conductor the current density is not uniform. It is stated in theory that the current density is shifted to the surface depending on the skin depth for that specific frequency.
The skin depth is the depth in the conductor where the current density has fallen to \(1/e\) of \(J_z\) in equation 7[10], see also figure 37.

\[
J = J_z e^{-d/\delta} \quad [A/m^2] \tag{7}
\]

Figure 37: Current density in round conductor affected by skin-effect[10].

The skin depth itself is defined in equation 8[10]. In this equation, the parameter \(\rho\) is the resistivity of the material and is temperature dependent, the operation temperature of the conductor/winding needs to be known for doing such a calculation. Another important parameter is \(\mu\), in this case a copper conductor is equal to \(\mu = \mu_0\).

\[
\delta = \sqrt{\frac{\rho}{\mu \pi f}} \quad [m] \tag{8}
\]

The skin depth of Copper in a warm motor of 150°C can then be calculated, see figure 38. In the example in the figure, the skin-effect starts to make an affect on the losses at around 1060 Hz for a conductor with width 5 mm.
6.3 Geometry

To simulate proximity losses the whole machine was drawn, not only a quarter as it was initially done in this project. To calculate the proximity and skin-effect in FLUX, the name of the conductors in the winding needs to be changed. Before in figure 13 the name was assigned to each phase, however, now each conductor has its own name. A label-system was created to keep all the names in track, see figure 39.

The motor has three phases that explain the name of the following index “phase(A/B/C)”. The motor also has two coils connected in parallel for each phase, that explains the next index “parallel coil in phase (1-2)”. In one parallel coil there are eight coils connected in series, index “coil (1-8)”. Each of these coils has five turns, but one turn passes through the motor twice, that results in the last index “conductor (1-10)”. The “orientation (p/n)” index tells in which direction the conductor passes through the motor, p stands for positive and n for negative. Every odd number of conductor has a positive direction, even numbers have negative direction, correspondingly.

Such labelling has resulted in 480 different names of conductors in the geometry, taking into account that all conductors are going to be simulated. The assumption can be made that the losses in one parallel coil are the same as in all others, as both the flux density and the applied current are the same in all coils during one electrical period. This means that only one parallel coil needs to be simulated in detail with the use of this label- system, such a simulation results in only 80 detailed conductors.

In order to be able to label all conductors correctly the winding arrangement needs to be analysed.
This motor has a double layer diamond winding. Half of the coils starts with the first conductor in the lower layer, see figure 40.

![Figure 40: Single coil conductor order, under to top.](image1)

And the other half starts with the first conductor in the higher layer, see figure 41.

![Figure 41: Single coil conductor order, top to under.](image2)

The placement of each coil and whether they are situated from the bottom to the top, or on the contrary, can be observed in the technical document for the winding arrangement of that specific motor.

6.4 Mesh

The meshing in the conductor for the simulation of skin-effect and proximity effect needs to be considered. The number of elements and how far they are supposed to be from the surface of the conductor depends on the skin depth.

If a simulation of frequency X Hz results in a skin depth of Y mm, then the first element needs to be closer than Y mm to the surface. In figure 42 the mesh for the used conductor is presented, one can observe that the elements density is higher close to the surface for mentioned purpose.
6.5 Material set-up

In the initial material set-up, see section 3.4.2, no material was defined for Copper as the resistance for the winding was set in the electrical set-up instead. Now in order to calculate the proximity and skin-effect the material copper needs to be defined.

The material property for copper defined is permeability, which equals to 1 and isotropic resistivity, that equals to 0.26144E-7 Ωm which corresponds to a temperature of 150°C.

6.6 Electrical Set-up

The electrical set-up for proximity and skin-effect simulation is different from the initial model. In figure 43 the schematic is presented and also the label-system is shown.

The main difference between the simulation proximity and skin-effect is that the component “stranded coil conductor” can not be used, but “solid conductor” needs to be used instead. This is because stranded coil conductor expects the uniform current density within the conductor.

In the figure A2, B1, B2, C1 and C2 are the parallel coils and are simulated as stranded coil conductors. This is because of the mentioned assumption in the previous section that it is not necessary to calculate the proximity and skin-effect in all parallel coils, as the result is same in all five parallel coils. This is why the top branch for parallel coil A1 in the figure contains only solid conductors.

One problem that occurs when connecting solid conductors and a stranded coil conductor in parallel (that is done in phase A in the schematic) is unbalanced currents. The stranded coil conductor has a
fixed resistance, but because of the proximity and skin-effect the resistance varies, in the solid conductor. What then happens is that the current is not balanced between two parallel coils. The solution of avoiding this problem is to split the current source into two, it provides half of the current to each coil, and places them on each of the respective parallel branch in the schematic. With this solution, the same current flows in both parallel coils, and no balancing problem occurs.

As it was mentioned in the material set-up, the resistance of the windings was initially set in the stranded coil, but a solid conductor uses the resistivity of the material instead. Then a change needs to be made as the end-windings are not included in the geometry, the easiest way to solve is to introduce a resistance in the electrical schematic. \( R_{end} \) is added to the schematic in series with the inductance for the end-windings. The total resistance of the windings was \( R_s = 0.0333 \Omega \). With the geometry of the conductor and the resistivity of copper the resistance for one solid conductor is 0.347898 m\( \Omega \). There are 80 solid conductors in series, what leads to a resistance of \( R_{coil} = 0.0278 \Omega \), but there are two conductors in parallel as well, that results in an effective resistance of 0.0139\( \Omega \). The resistance of the end-windings then is \( R_{end} = R_s - 0.0139 = 0.0194 \Omega \).

\( R_{coil} = 0.0139 \Omega \) from above is the resistance in this model that is put in the five stranded coil conductor components: A2, B1, B2, C1 and C2

### 6.7 Result

The results in section 6.7.1 to 6.7.7 are displaying the mean losses for respective conductor position. The geometric meaning of the conductor positions are shown in figure 44, where each conductor is numbered from the air gap up to the stator yoke.

For each operation point simulation has been done with currents based on measurements, this is done for: Asynchronous PWM in o.p. 2, PWM-S9 in o.p. 4 and in all the square-wave cases in o.p. 7 to 12. The currents in the motor are not ideal and contain an unwanted DC current which affects the results and needs to be taken in to account. As the resistance of one conductor is known and the DC current is known, the DC losses can therefore be calculated. Then to exclude the DC losses from the results one can simply subtract them. The DC currents in the measurements used in this project varied from 0.0018 p.u. for o.p. 2 up to 0.05 p.u. for o.p. 10. The DC losses have been excluded in the results presented in this section.

![Figure 44: Geometric meaning of the conductor positions.](image)

#### 6.7.1 Operation Point 2

Operation point 2 is at 21 Hz with rated torque and fundamental power of 299 kW. This operation point is at under-modulation with a modulation index of around 0.36. Because of this the chosen modulation method is asynchronous PWM. Switching frequency of 550 Hz, 1 kHz and 2 kHz are compared to sinusoidal applied current. The results of this comparison can be seen in figure 45.
As it is seen in the figure with sinusoidal current, there is nearly no difference if one compares the losses at conductor one and conductor ten. At this frequency, the contribution of the skin effect is very weak for the fundamental as the frequency is very low. The skin depth is 18 mm, which is far larger than the conductor width. For the simulation with harmonics, the switching frequency gives rise to harmonic content. The skin depth is from 3.47 mm at 550 Hz to 1.82 mm at 2 kHz. This harmonic content results in higher losses due to skin-effect. The losses in conductor ten increase when the switching frequency is lowered.

According to the theory, the losses due to proximity effect should increase if the slot leakage flux density increases. The closer to the air gap the more slot leakage flux density. The phenomenon of proximity losses can then easily be observed in the figure as the losses for three simulations with PWM increase the closer it gets to the air gap. The difference between 550 Hz and 1 kHz is much larger than between 1 kHz and 2 kHz, this is because of the inductance of the motor which suppresses the higher frequencies very efficiently. There is a small increase of losses in the first conductor with sinusoidal current compared to conductor number ten. One can assume this increase is because of proximity effect. Therefore it is clear that in this particular case the harmonic content is the cause for the main increase of proximity losses.

In section 6.6 the resistance for one solid conductor was presented. This resistance is calculated for a conductor with uniform current density, in principle a DC current in a conductor without any crossing magnetic fields. In the results presented in this section the case is totally opposite, a AC current with crossing magnetic fields. This means that the effective resistance of the solid conductor is not the same as the initial “DC” resistance. The current in each conductor is known and so is the losses, with this information the effective resistance can be calculated. The same principle apply on all operation points.

For this operation point the effective resistance of the Asynchronous PWM modulation has been compared to the results in the sinusoidal case, see figure 46. The rms current of the Asynchronous PWM modulation is 1.0134 p.u. as it contains harmonics.
The effective resistance of the conductor in the active part of the winding increases the closer one gets to the air gap as seen in the figure. This increase is mainly due to the harmonics in the supplied current as the sinusoidal current only has a small visible impact on the first conductor closest to the air gap.

### 6.7.2 Operation Point 4

Operation point 4 is at 56 Hz with rated torque and fundamental power of 676 kW. This operation point is at under-modulation with a modulation index of around 0.88, because of this the chosen modulation method is called PWM-S9. Switching frequencies compared are: 550 Hz, 1 kHz, 2 kHz and also sinusoidal. The results can be seen in figure 47.
While observing conductor ten one can notice that from PWM at 550 Hz to sinusoidal there is an increase in losses. For 56 Hz the skin depth is 10.87 mm, this value is still larger than the conductor’s, so one can assume that fundamental frequency does not increase the skin-effect losses. However, the harmonics have much lower skin depth - it can affect the skin-effect losses, as PWM 2 kHz has the lowest losses of three. Then the losses increase up to PWM at 550 Hz with lowest switching frequency.

By observing the figure, it is clear that the highest losses are in the conductor closest to the air gap with highest crossing flux density. However, even at this frequency the sinusoidal current involves the increase in losses on the first conductor. The modulation method with highest losses is PWM at 550 Hz as it has lowest switching frequency. It can be observed that the harmonics created by the modulation method affect the losses more and more the closer one gets to the air gap. The difference between PWM 550 Hz and PWM 1 kHz is not neglectable, and it can be observed that the increase of switching frequency decreases the proximity losses. An increase of switching frequency to PWM 2 kHz lowers the proximity losses even more by nearly the same value as the difference between PWM 550 Hz and PWM 1 kHz.

For this operation point the effective resistance of the PWM-S9 modulation has been compared to the results in the sinusoidal case, see figure 48. The rms current of the PWM-S9 modulation is 1.0358 p.u.

Figure 47: Mean losses in the conductors.
The behaviour of the effective resistance in this operation point is similar to o.p. 2, with two differences. First, the increase of the effective resistance is higher than for o.p. 2. Second, the sinusoidal current affects not only the first conductor but up to the fourth conductor.

6.7.3 Operation Point 7

Operation point 7 is at 99 Hz with rated torque and fundamental power of 654 kW. This operation point is at maximum over-modulation with the use of square-wave waveform. To get the current waveform for the square-wave input, the measured current from this operation point was used in the simulation. Same principle applies on all operation points with square-wave modulation. The results can be seen in figure 49.
The fundamental frequency of 99 Hz has a skin depth of 8.179 mm, the value is still too large to make an impact on the skin-effect losses. The first large current harmonics are fifth and seventh, and they have an impact on the skin-effect losses.

The proximity losses (when one observe the sinusoidal case in the figure) have nearly no affect before the conductor one, then the losses have a very distinct increase for conductor one. For the square-wave case the proximity losses increase slowly the closer one gets to the air gap, but then on the last conductor it is the same distinct increase as for sinusoidal case.

For this operation point the effective resistance of the square-wave modulation has been compared to the results in the sinusoidal case, see figure 50. The rms current of the square-wave modulation is 1.0385 p.u.
For this operation point the sinusoidal case show that the larger increase of the resistance on the first conductor is because of the fundamental current. The square-wave modulation has an increasing difference to the sinusoidal case the closer one get to the air gap, this is because of the harmonics.

All operation points with square-wave modulation show similar behaviour both in the losses in the conductor and in the behaviour of the effective resistance.

### 6.7.4 Operation Point 8

Operation point 8 is at 109 Hz square-wave modulation with rated torque and fundamental power of 656 kW. The results can be seen in figure 51.
The fundamental frequency of 109 Hz has a skin depth of 7.795 mm, the value is still too large to make an impact on the skin-effect losses.

If one compares this operation point with number 7, the main difference is that the distinct loss increase for conductor number one is even higher with 109 Hz at number 8.

The rms current of the square-wave modulation is 1.0326 p.u., see figure 52 for the behaviour of the effective resistance.

Figure 51: Mean losses in the conductors.
6.7.5 Operation Point 9

Operation point 9 is at 129 Hz square-wave modulation with rated torque and fundamental power of 661 kW. The results can be seen in the figure 53.
The fundamental frequency of 129 Hz has a skin depth of 7.165 mm this value is also too large to make an impact on the skin-effect losses.

If one compares this operation point, with previous ones, the distinct loss increase for conductor number one is even higher at 129 Hz. All operation points with square-wave modulation have similar behaviour, with an increase in frequency the distinctive losses in conductor number one increase.

The rms current of the square-wave modulation is 1.0220 p.u., see figure 54 for the behaviour of the effective resistance.
Figure 54: Effective resistance for different conductors in the stator slot.

6.7.6 Operation Point 10

Operation point 10 is at 149 Hz square-wave modulation with rated torque and fundamental power of 669 kW. The square-wave modulation was also compared to a modulation method called EP5[2]. The results of this comparison can be seen in figure 55.
The fundamental frequency of 149 Hz has a skin depth of 6.667 mm, the value is still too large to make an impact on the skin-effect losses as well as in previous cases.

The EP5 modulation method has same proximity loss pattern as for square-wave, increase of losses the close one get to the air gap and a distinct increase for conductor number one. The difference here is that EP5 has a different harmonic spectra compared to square-wave, this cause the proximity losses to increase even more than for square-wave modulation.

The rms current of the square-wave modulation is 1.0156 p.u., see figure 56 for the behaviour of the effective resistance. Modulation method EP5 has not been compared to.

Figure 55: Mean losses in the conductors.
6.7.7 Operation Point 12

Operation point 12 is at 179 Hz square-wave modulation with rated torque and fundamental power of 657 kW. The results can be seen in figure 57.
The fundamental frequency 179 Hz has a skin depth on 6.082 mm which is also still too large to make an impact on the skin-effect losses.

The difference between sinusoidal and square-wave is smaller for this operation point than the lower ones, this is because of decrease of harmonic content as this higher frequency the inductance of the motor suppress the harmonics more efficient.

The rms current of the square-wave modulation is 1.0123 p.u. of the sinusoidal one, see figure 58 for the behaviour of the effective resistance.

Figure 57: Mean losses in the conductors.
At this high fundamental frequency the sinusoidal current affect the effective resistance on all conductors. The resistance increase closer to the air gap and has a nearly twice increase on conductor one. The square-wave modulation contains the contribution of the harmonics and the difference to the sinusoidal case increase closer to the air gap, same as in other operation points. The effective resistance increase most in this operation point and is as high as 2.14 times higher than the DC resistance on the first conductor.

6.7.8 Analysis of the Operation Points

The figures in operation point 2 and 4, figure 45 and 47, clearly present that an increase in switching frequency lower the resistive losses in the active part of the winding. The active part of the winding is the part of the winding existing inside the motor in the stator slot, by that excluding the end-winding. Table 6.7.8 state how large the decrease is for the two operation points if one increase the switching frequency from 550 Hz to 1 kHz, 2 kHz or use pure sinusoidal applied current.

<table>
<thead>
<tr>
<th>Switching Frequency</th>
<th>O.p. 2</th>
<th>O.p. 4</th>
</tr>
</thead>
<tbody>
<tr>
<td>550 Hz</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>1 kHz</td>
<td>5.24%</td>
<td>8.04%</td>
</tr>
<tr>
<td>2 kHz</td>
<td>6.71%</td>
<td>11.80%</td>
</tr>
<tr>
<td>sinusoidal</td>
<td>7.64%</td>
<td>15.97%</td>
</tr>
</tbody>
</table>

Table 3: Decrease in losses if switching frequency is increased from 550 Hz.

For the operation point 10 square-wave modulation was compared to the modulation method EP5, see figure 55. In the figure one could observe that the losses with modulation method EP5 were slightly higher, the loss increase was 7.27% of the active part of the winding.

In all operation points one could observe an increase in losses for conductor number one compared to conductor number ten. The relative loss increase for conductor number one compared to conductor
number ten is presented in figure 59 for sinusoidal case. The relative loss increase for conductor number one is nearly linear when the fundamental frequency is increased. This loss increase displayed in figure 59 is only due to the proximity effect as no time harmonics are presented in the simulation with pure sinusoidal current.

![Graph showing relative loss increase for conductor number one compared to number ten for sinusoidal current.](image)

Figure 59: Relative loss increase for conductor number one compared to conductor number ten for sinusoidal applied current. Combined results from all operation points.

For square-wave modulation, in operation point 7 to 12, the relative increase of losses for conductor number one compared to conductor number ten are of an average of 18.49 percentage points higher than pure sinusoidal, see figure 60.
Figure 60: Relative loss increase for conductor number one compared to conductor number ten for square-wave modulation. Results from operation point 7 to 12 where square-wave modulation was used.

There is not only a relative loss increase for conductor number one but for all conductors, except for conductor number ten which is the reference. The relative loss increase for all conductors is presented in figure 61 for sinusoidal applied current.
Figure 61: Relative loss increase for all conductors, conductor number ten for respective frequency is the loss reference. Sinusoidal applied current. Combined results from all operation points.

For square-wave modulation the loss increase is higher than for sinusoidal current, see figure 62. One reason for this is the relative high amplitude low order harmonics which create high flux density harmonics. The higher frequency of the flux density the closer the flux lines will be to the conductor, this then results in a higher slot leakage flux density in the stator slot.
Figure 62: Relative loss increase for all conductors, conductor number ten for respective frequency is the loss reference. Square-wave modulation. Results from operation point 7 to 12 where square-wave modulation was used.

The percentage point increase of losses for square-wave modulation differs depending on the position of the conductor in the stator slot. In figure 63 the average rise in losses is presented for each conductor in the stator slot.
Figure 63: The average percentage point increase of losses between square-wave modulation and sinusoidal operation for each conductor in the stator slot. Only results from operation point 7 to 12 where square-wave modulation was used.

6.8 Summary
In this project for proximity losses the whole machine was drawn and that resulted in four times as much elements as before in the FEM model. For further investigation, it can be possible to only simulate with a quarter of the motor. If possible then the computational time could be reduced to around half.

In the label-system, the orientation index $(p/n)$ should has been placed after the conductor index instead to have a list in chronological order. Now every odd and even has different orientation, and the lists did not follow chronological order what resulted in a longer time to set-up the model.

With the increase of frequency at over-modulation, the proximity losses for modulation operation become more similar to sinusoidal operation as the inductance of the motor suppresses more and more the harmonic content.

The losses of the conductor closest to the air gap have higher proximity losses than the other conductors. This loss difference also increases with frequency, this can be seen if one compares the figures for number 4 and number 12.

The losses due to skin-effect are not neglectable, however, compared to the proximity losses the magnitude of these losses is significantly smaller.
7 Discussion

7.1 FLUX as a Simulation tool for Induction Machines

The approach for this project was to create a FLUX model that would replicate the existing machine as accurately as possible.

One problem that has occurred during creation of the FLUX model was that the performance of the machine is extremely dependent on the rotor short-circuit ring coefficients: the inductance and the resistance. These coefficients are not calculated in FLUX. To calculate these coefficients, an analytical program was employed. The resulting torque was around 1.6% of the desired torque, at a certain rpm one desired to obtain around 1000 Nm, but the result from the FLUX model was 16 Nm. To solve this deviation in torque, parametric solving was implemented to find suitable values, see section 3.4.4.

The method works as the performance of the machine is known. If one needs to simulate a machine in FLUX without available measurements, then it is hard to answer the question what are the values of these coefficients.

Further analysis in ways to calculate these coefficients needs to be made, in order to be able to simulate potentially interesting designs in FLUX.

7.2 Drawing the Geometry in FLUX

The sketcher tool in FLUX similar to the CAD software and the design can be done much faster than before. This sketcher is perfect if one knows exactly what to draw and that the geometry will never be changed. In this project, the geometry was supposed to be adaptable and parametrised to any induction machine in Bombardier with the same layout. The sketcher used in the software is not programmed to use parameters. Therefore, the sketcher was not used in this project.

7.3 Switching from Steady State Application to Transient Application in FLUX

In section 3.2 there is a step-list for how to go from the steady state application to transient application, since the mentioned way is not an obvious one.

The first way that was tested was simply to delete the steady state application and to create a transient application in the same file. The mesh and geometry would be unchanged, and nothing in the physics would be changed. But the program FLUX can not handle this, and a lot of thought was put in the reason why it did not work. There was no error in the model it was just a flaw of FLUX.

In the end, the FLUX support has created the step-list explaining how to use their software. The method proposed will take some more hours than the initial obvious way. In this project, these errors/bugs in FLUX delayed the project with several weeks.

7.4 Simulation of time harmonic rotor bar losses

This project has not included the time harmonic rotor bar losses, one reason for this is explained in this subsection.

The losses in the rotor bars are dependent on the induced current in the rotor bars. The current in the rotor bars has a very low frequency, the same as the slip frequency. To understand the behaviour of that current at least one-quarter of a period needs to be simulated in order to extrapolate the rest of the period.

According to CEDRAT if one wants to calculate losses, for example iron losses, the recommended time steps for the input current/voltage per period should not be less than 40. This number is for a sinusoidal case in order for the current to have a good enough sinusoidal character. But if one wants to include time harmonics of a certain frequency then it has to be enough time steps at that frequency also. If the assumption is made that one also needs 40 time steps per period for the highest important harmonics one wants to consider, then the total number of time steps will be very high.

A simple example. At one operation point at 100 Hz there is a slip frequency of 1 Hz, square-wave modulation is used and highest important time harmonic is considered to be the seventh harmonic. As the slip frequency is 1 Hz and one period is 1 s long, one needs to simulate $1/4 = 0.25$ s to be
able to extrapolate the current. The seventh time harmonic has a frequency of 700 Hz, so one needs $700 \times 40 \times 0.25 = 7000$ time steps to be able to calculate the rotor bar current. With the detailed model used in this project one can approximately simulate 400-500 steps per day in a transient simulation. This means that one simulation of the rotor harmonics would take 14-17.5 days to complete, even without considering the initial transients in the simulation.

The problem with the long simulation time in FLUX is that not all processes in the solver are programmed for multiple cores. That means even if 12 cores were used in this project, adding 12 more cores will not necessarily reduce the simulation time by half.
8 Conclusions

8.1 Conclusions from the project

A FEM model for an induction machine has been created in the program FLUX. The model results of the no-load and the locked rotor test scenario are in a good agreement with the corresponding motor tests. So with these tests validated one can assume the model in FLUX delivers acceptable performance.

The results of the calculations in this project show that an increase of switching frequency would decrease the iron losses, for example a fundamental frequency of 56 Hz with an increase of switching frequency from around 550 Hz to 2 kHz would lower the losses to nearly half.

For proximity and skin-effect losses the whole machine needed to be drawn and that resulted in four times more elements in the FEM model. Further investigations are necessary, so it can be possible to simulate only a quarter of the motor. Hence, the computational time might be reduced to half.

The report shows that because of the proximity- and skin-effect, the first conductor closest to the air gap has distinct higher losses than the other conductors further away from the air gap. The distinct difference of losses between the first conductor and the other conductors also increases with frequency.

The simulation of proximity and skin-effect losses also presents that an increased level of lower order harmonics affects the losses of the first conductor mostly. The loss increase due to this phenomena then decreases the further the conductors are placed from the air gap. At square-wave modulation the lower frequencies had higher time harmonic conductive losses than the higher frequencies, the reason is that the inductance of the motor is higher and then gives rise to smaller time harmonics.

8.2 Future work

To continue the research one could do measurements of the induction machine with the modulation method of interest, for example with 1 and 2 kHz switching frequency for PWM modulation at lower fundamental frequencies. The results from the measurements could be compared to the models presented in this report and also used to see the direct loss impact of a change in switching frequency.

To investigate the proximity- and skin-effect more one could look into the conductor losses depending on the geometric position in the stator. In this project the mean losses of the conductors position related to the air gap was analysed. But what does the losses in the conductor looks like if the slot is next to a another slot with another phase, or the slot is between two slots with the same phase? As the proximity effect is depending on the slot leakage flux the results should not be the same and it would be interesting to know how much they differ.
References


