Abstract—It is well-known that using expensive Litz wire is an effective solution to reduce eddy current losses in high-speed low-power electrical machines. However, in an offshore wind power application where a direct-drive 10 MW ironless-stator generator and full-scale converter are used, the use of Litz wire becomes also necessary. Meanwhile, the progress in the manufacture of Roebel winding and Continuously Transposed Conductors (CTC) makes these conventional cheap solutions promising to be employed for a cost-effective generator. This study examines the feasibility of using such technologies instead of Litz wire by studying the winding losses associated with each solution. Completely transposed windings with different number of subconductors per turn were considered. Both resistive and rotational losses are explored in this analysis by means of three-dimensional finite element method (3-D FEM) software (i.e. Ansys Maxwell) integrated with an advanced computing server of 48-core and 128 RAM. In terms of resistive loss, decisive factors are the surface and length of the strand, since paths followed by each subconductor in the interlacement process pose a small influence, especially because of the low frequencies in play and the fact that circulating currents are highly reduced in a complete transposition. On the other hand, rotational losses depend mainly on winding dimensions and orientation. Results show that both proposed technologies are still far from delivering the loss reduction performance provided by Litz wire.

Index Terms—Axial Flux Permanent Magnet (AFPM) generator, circulating currents, Continuously Transposed Conductors (CTC), eddy losses, Litz wire, Roebel transposition.

I. INTRODUCTION

A. State of the art

Conventional iron-stator electrical machine demands a heavy structure to counter the normal force between stator and rotor so that the air gap can be maintained. Ironless-stator large-diameter generator is capable of producing huge torque $T \propto r^2$ and the normal force between stator and rotor is negligible so an extremely lightweight generator can be expected, as is shown in Fig. 1. Even though the magnet load in ironless axial flux machine is quite low (<0.4T usually), there is significant tangential component, which also contribute to the eddy loss in winding [1]. The use of the full-scale converter injects much harmonics current into the stator windings and induces more eddy losses.

![Fig. 1. Concept of torque production ($\sigma_T$; force stress)](image)

Litz wire has brought its advantages to reduce eddy loss considerably, therefore, is being used in the renewable energy market [2]. However, this multiple strand flexible solution is considerably more expensive. Meanwhile, the new technology in manufacturing of conventional cheap transposed winding has evolved so much that interesting levels of loss reduction can be obtained when subjected to more complex transposition.

In this paper, CTC and Roebel bars have been modeled for a machine with specifications given in Table I. Effectiveness of both solutions has been then evaluated with a three-dimensional finite element method.

<table>
<thead>
<tr>
<th>Table I - Generator specifications.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated power</td>
</tr>
<tr>
<td>Rated speed</td>
</tr>
<tr>
<td>Rated voltage</td>
</tr>
<tr>
<td>Outer diameter</td>
</tr>
<tr>
<td>Inner diameter</td>
</tr>
<tr>
<td>Number of poles</td>
</tr>
<tr>
<td>Winding width</td>
</tr>
<tr>
<td>Winding thickness</td>
</tr>
<tr>
<td>Winding material</td>
</tr>
<tr>
<td>Magnets material</td>
</tr>
<tr>
<td>Rotor steel</td>
</tr>
</tbody>
</table>

B. Preponderance of current density

In an AFPM machine the stator winding is located in the air gap magnetic field. Energy generation is accomplished by applying movement to the rotor which,
together with its attached magnets, originates current in the stator windings, as illustrated by Fig. 2. The output power of the machine is function of both its torque magnitude $T$ and angular speed $\omega$, as shown in equation (1).

$$P = \frac{dW}{dt} = \frac{d}{dt} \left( \int F \cdot d\vec{x} \right) = \frac{d}{dt} \left( \int F \cdot Rd\theta \right) = \frac{Td\theta}{dt} = T\omega \quad (1)$$

Since wind generators are characterized by working at low speeds, a powerful machine necessarily implies high torques. The electromagnetic torque is directly proportional to the magnetic force. This force depends on the stator current and on the air gap magnetic field. In order to increase the output power, the tangential force on the machine will have to be enhanced. The rotor’s magnetic field is naturally limited by the hysteresis loop of its permanent magnets. Consequently, the rated current is the key variable in achieving high power machines. High current densities give rise to significant losses in the stator conductors, making the study and calculation of such losses a decisive factor in the designing process of the generators addressed in this research.

In addition, if this exposition to the magnetic field is not equally distributed by the subconductors, the flux linked with each one of them differs from one to the other, which results in different induced voltages. Thus, as the strands are short-circuited somewhere in the coil (when there is only transposition in the active region short-circuit happens in the end region; but the coil may also be totally transposed), eddy circulating currents arise between the subconductors. This fact naturally alters the resistive losses.

A. Resistive loss

The phenomena to be discussed relate to distortions in the distribution of current-density over the cross-section of conductors. These effects may be divided into three classes. Firstly, the skin effect due to disturbance of current density in a conductor due to the alternating magnetic flux linked with the same. It may be regarded as due either to imperfect penetration of electric current into the conductor, or to the greater reactance of the central core of the conductor with respect to the surface layer, whereby the current density is less on the inside than on the outside. In a uniform solid round wire, the skin effect is symmetrical with respect to its axis, while in solid wires of other than circular form, the skin effect is, in general, dissymmetrical. Secondly, the spirality effect found in spiraled stranded conductors and due to the reactance of the spirals. Lastly, the proximity effect, found in parallel linear conductors of any cross sectional form when in proximity, owing to the alternating magnetic flux from one penetrating the other [3].

1) Circulating currents

As a cost effective solution for eddy current losses reduction, parallel thin wires are used in every turn. However, this may create a new problem, which is unless a complete balance of induced EMFs among the individual conducting paths is achieved, a circulating current between any of these parallel paths may occur, causing circulating eddy current losses [4]. The current within each strand is the sum of two components [5]:
   i. input current (AC stator current) that flows uniformly into every strand;
   ii. circulating current that differs from strand to strand which sums to zero over the cross-section of all the strands in one bar.

The circuit considered to represent the active region of the winding is shown in Fig. 3. $I$ is the stator current, while $i_k + I/n$ is the strand current, with $I/n$ being the fraction of stator current that flows uniformly into every strand and $i_k$ the circulating current, that differs from strand to strand. Magnetic coupling between strands is accounted for by mutual-inductances $L_{ij}$, while self-inductances $L_{kk}$ account for the influence of the internal magnetic flux produced by the current $I/n$. It should be noted that the magnetic coupling between different coils
is assumed to be small and, therefore, is disregarded in this analysis. Finally, $R_s$ is the resistance of the strand.

Applying Ohm’s law to one strand the voltage along the strand $V_k$ is found, as shown in equation (2).

$$V_k = R_s(i_k + I/n) + \sum_{l=1}^{n} L_{k,l} \frac{d(i_l + I/n)}{dt} \quad (2)$$

Applying Faraday’s law of induction to one strand, equation (3) is found.

$$\oint E \cdot ds = -\frac{d\psi_s}{dt} \Rightarrow V_k - V = -\frac{d\psi_s}{dt} \quad (3)$$

$V$ is the unknown voltage along the winding with respect to the fictitious conductor that, along with current filament of strand, closes path $s$.

B. Rotational loss

Components of the magnetic field present in the air gap originate an electric field with the orientation illustrated in Fig. 4. Since it is inside a conductive material, this electric field will then give rise to eddy current loops inside the conductor. As it can be seen in Fig. 4, this current loops influence each other. For instance, top and bottom surfaces from the red loop have the same direction of lateral surfaces of the blue one, which means the resulting current in the conductor is function of both tangential and axial flux. This implies losses cannot be computed in separate considering components of magnetic induction independently of each other as it is done in [4]. The only option is to consider the system as a whole. The relationship between the components of magnetic induction vector and the components of electric field vector is as shown in set of equations (4).

$$\begin{align*}
\frac{\partial B_x}{\partial t} &= \frac{\partial E_y}{\partial z} - \frac{\partial E_z}{\partial y} \\
\frac{\partial B_y}{\partial t} &= \frac{\partial E_z}{\partial x} - \frac{\partial E_x}{\partial z} \\
\frac{\partial B_z}{\partial t} &= \frac{\partial E_x}{\partial y} - \frac{\partial E_y}{\partial x}
\end{align*} \quad (4)$$

Fig. 5 illustrates the relationship between the dimensions of subconductor. It is clear that winding’s length is much bigger than both winding’s width and thickness. This type of geometry allows it to be assumed that the variation of component $z$ of the electric field along X is small when compared to the variation of component $x$ of the electric field along Z. Likewise, the variation of component $y$ of the electric field along X is small when compared to the variation of component $x$ of the electric field along Y. Thus, the problem can be reduced to system (5).

$$\begin{align*}
\frac{\partial B_x}{\partial t} &= \frac{\partial E_y}{\partial z} \\
\frac{\partial B_y}{\partial t} &= \frac{\partial E_z}{\partial x} \\
\frac{\partial B_z}{\partial t} &= \frac{\partial E_x}{\partial y}
\end{align*} \quad (5)$$

As it stands, there is still interdependence between the components of magnetic induction and the system remains too complex to be solved because there is no information regarding the charge density inside the conductor. However, in an AFFM machine the longitudinal component of the magnetic flux is almost inexistente and can be disregarded when compared to the axial and tangential components. This acceptable approximation greatly simplifies the problem, allowing rotational losses to be calculated directly from $P_j = \int \sigma E^2 dV$, provided the electric field inside the
Conductor is easily obtained by solving system of partial differential equations (6).

\[
\begin{align*}
\frac{\partial B_x}{\partial t} & = -\frac{\partial E_y}{\partial z} \\
\frac{\partial B_z}{\partial t} & = \frac{\partial E_x}{\partial y} \\
\frac{\partial B_y}{\partial t} & = \frac{\partial E_z}{\partial x} 
\end{align*}
\]  

(6)

Using phasor notation, \( B_y = B_{ye} e^{j\omega t} e^{-j\phi} = B_{ye} e^{j\omega t} \), \( B_z = B_{ze} e^{j\omega t} \) and \( E(y, z) = E_{ye} e^{j\omega t} \), the solution is found in last equality of set of equations (7).

\[
\begin{align*}
j\omega B_y & = -\frac{\partial E}{\partial z} \Rightarrow \tilde{E} = -j\omega z B_y + f(y) + C \\
j\omega B_z & = \frac{\partial E}{\partial y} \Rightarrow f(y) = j\omega z B_z \\
\tilde{E} & = j\omega y B_z - j\omega z B_y + C 
\end{align*}
\]  

(7)

To determine constant \( C \), the constitutive relation that states \( \nabla \cdot J = 0 \) must be considered. Using Gauss’ theorem, \( \int_{V} J \cdot n \, dS = 0 \) is obtained, which means that the number of \( J \) lines entering a given volume is equal to those leaving it. Basically, this fact indicates that current loops flow around a symmetry axis, as illustrated in Fig. 6, having the current density symmetric values with respect to it. Since \( J \) is directly proportional to \( E \), the boundary condition for the coordinate system adopted in Fig. 6 is \( E_{y}(y = 0, z = 0) = 0 \), which implies \( C = 0 \). With the electric field defined, the total rotational loss in the conductor is simply found as shown in equation (8), where \( c \) is the length of the conductor.

\[
\begin{align*}
\Delta P_{c} & = \int_{V} \sigma E_{c}^{2} dV \\
& = \int_{V} \sigma \left( \frac{j\omega (yB_z - zB_y)}{\sqrt{2}} \right)^{2} dV \\
& = c \int_{-\frac{L}{2}}^{\frac{L}{2}} \int_{-\frac{L}{2}}^{\frac{L}{2}} \sigma \left( \frac{j\omega (yB_z - zB_y)}{\sqrt{2}} \right)^{2} dydz 
\end{align*}
\]  

(8)

III. APPROACH

Since the large scale wind generator under analysis works at low frequencies, resistance limited approach is used to evaluate eddy current losses [2], which implies the following assumptions:

i. as conductor dimensions are small, flux produced by eddy currents has a negligible influence on the total field [6];

ii. conductor dimensions are smaller than the skin depth: the eddy current loss due to the load current is ignored, as it is only a very small percentage of the loss due to the air gap magnetic field [6].

Both solutions presented in this study concern using parallel wires with smaller cross sections instead of one thick conductor. Attending to Faraday’s law of induction, this may create a new problem: unless a complete balance of induced EMFs among the individual conducting paths is achieved, a circulating current between any of these parallel paths may occur, causing circulating eddy current losses [1]. Basically, not only time derivative of flux linkage due to rotor magnets’ movement is different from strand to strand, but also time derivative of flux linkage in each strand due to current that flows in every other strand also differs from one strand to another. By transposing the wires in such a way that each parallel conductor occupies all possible layer positions for the same length of the coil, the induced EMFs in all parallel conductors are equalized and circulating currents can generally be ignored in a heavily twisted coil, which is the case of the solutions presented in this paper (i.e. 360 degrees transposition) [1]. Alternating current that flows in stator conductors originates magnetic field that alters the one coming from the rotor, as illustrated in Fig. 7. The influence of this armature reaction on the eddy current losses is usually insignificant [1], and is therefore disregarded in this analysis. However, it is intrinsically considered in the Joule losses analysis, where the...
magnetic field created by the current in one strand influences the impedance of the remaining strands, as it will be seen further on this report.

![Fig. 7. Application of Ampère’s law to the current flowing threw a cylindrical conductor - Source: [7].](image)

The most straightforward approach to solve the problem would be to completely model the whole segment of the machine, with the transposed coils in between the double-sided rotor. Then to run a transient simulation on the model by rotating the segment of the rotor at rated speed. Such approach would give an exact replica of the real world environment, and losses would be computed considering circulating currents, rotational and resistive losses all together. However, several problems arise when trying to perform this method. Firstly, designing the end-winding region where different turns connect is a task of great complexity. Since the behavior of a transposed winding is to be evaluated, the simple method of designing a phase coil as a massive cross section conductor (and simply input to the software the number of turns in play for it to compute internally) does not work. It is necessary to model and transpose each turn individually, which means the coil has to be built by connecting turn by turn at the end winding region. Although several methods were attempted in order to achieve a satisfactory model of one coil, intersections between different strands kept occurring. Of course it would have been possible to turn to a more powerful three-dimensional software (e.g. SolidWorks) just for designing purposes and then convert it into a Maxwell model, but two reasons advised to search for other solutions: 1. the idea of testing cases with several situations (i.e. several number of strands and different types of winding) required the coding to be kept only about the Maxwell environment; 2. analysis were performed on the active region devoid of the end-winding region, and a considerably amount of time was consumed just to compute the initial mesh, so, even considering a good design of the coil, with the end region, was achieved, simulation would most likely not return any results due to lack of computational power. Moreover, the method followed to calculate eddy current losses allows a solution to be found in a magnetostatic environment, since field periodically variation can be computed using Fourier analysis along pole-pitch. To run a transient solution type would not only be an uncertain step considering the complexity of the design itself, but would also produce loss results that would not allow distinctions to be made between resistive and rotational losses. Consequently, it would not be possible to depict loss sources of each solution and understand what can be changed to achieve a more competitive winding.

IV. TRANSPOSED WINDING

The armature of the machine is composed by nonconductive material that supports phase current conductors. Instead of a single massive cross section, each coil turn is built from several sub conductors of rectangular cross section. The model under analysis consists of several strands placed in two adjacent rows. In order to reduce losses due to circulating currents, 360 degrees transposition is applied in the active region of the machine. The main difference between the two types of transposed wire taken into account is illustrated in . Basically, while transposition process in CTC is physically stepped, i.e. each strand shifts position along a small slant and then develops its path perpendicularly to coil cross section plan, in Roebel each strand extends itself obliquely to coil cross section plan.

![Fig. 8. 2-D profile comparison between CTC and Roebel.](image)

Fig. 9 illustrates the sizing of a CTC. Being $T$, $W$ and $L$ the thickness, width and length, respectively, of one turn of the winding without applied transposition, $t = T/(n+1)$, $w = W/2$ and $l = L$ are the thickness, width and length, respectively, of each CTC’s strand with $n$ sub conductors per row. The extra space needed in height is to allow shifting between rows. In a 360 degrees transposition, the length each strand stands in one of the $2n$ possible positions (or the transposition pitch) is $dx = l/(2n)$ . Longitudinal distance between the beginning/end of a slope and the transposition pitch is $delta = dx/10$. During a shift in the same row, a strand travels $dx = delta/2$ along the winding length direction, while a shift between rows means $dy = dx/5$ displacement in the same direction. Naturally, shifts in the same row mean a displacement of $dz = t$ along axial direction of the machine, while a shift between rows entails a $dy = w$ displacement along the winding width direction. Fig. 10 illustrates a Roebel bar. Same logic of CTC is applied, except thickness of each strand is $t = T/(n+2)$ , being the space lost to allow shifting between rows one strand’s height longer than CTC. This is due to the fact of each strand displacing itself obliquely. During a shift between rows, a strand travels $t$ upwards and downwards shifting sideward in between. Each of these three displacements entails a longitudinal progression of $dx/3$.
A. Resistive loss

The goal is to determine stator’s winding conductive performance under a range of frequencies around the rated speed of the machine, so that some understanding can be gained about the influence subconductors have on each other’s resistance. To account for the skin effect, a two-dimensional model is run for only one straight subconductor with the same cross section dimension it would have on a transposed arrangement. Then, the proximity effect is evaluated by running a two-dimensional model in a stranded winding without applied transposition. It should be noted that when studying the proximity effect, the skin effect is automatically considered. Finally, the spirality effect is accounted for by running a three-dimensional simulation on the transposed winding. Naturally, the skin and proximity effects are intrinsically considered. As already discussed in section III of this paper, the analysis is done solely considering the winding itself (i.e. rotor’s magnetic field is disregarded and separately accounted for as rotational loss).

1) Resistance

To evaluate solely the skin effect on each strand, it is necessary to study the latter isolated. Thus, tests were performed on a straight conductor with the same cross section of one strand for every transposition type in study. It was verified that, regardless of the technology, there is no measurable difference between the resistance of one isolated strand and the resistance of the same subconductor performing in a stranded wire. In other words, for the small dimensions and the array of low frequencies in study, the skin and proximity effects play the same role in shaping the resistive loss of the wire. Tests performed for a CTC with two strands per row are described below. Same procedure was followed to the remaining technologies, being the conclusions deducted here extensible to the latter. To evaluate the proximity effect, a stranded nontransposed winding is tested. Fig. 11 illustrates the distribution of the magnitude of current density vector over the cross section when performing under the rated frequency. It is verified that, although a rearrangement in the current density is noticeable, the differences are not big enough to cause a measurable change in the resistance of the conductor.

To evaluate the spirality effect, a stranded and transposed winding is tested using a three-dimensional model. Fig. 12 shows the distribution of the magnitude of current density vector over the cross sections when performing under the rated frequency. It is verified that the distribution of current density remains the same as in the nontransposed case and that the resistance raise is only due to the winding length increase that naturally occurs as a result of the interlacement.
Although the machine rotates at low speeds, the grid connection process involves electronic converters, which generally work at higher frequencies, introducing high order current harmonics in the stator. In order to evaluate this effect, each of the transposed technologies was studied under frequencies of 1 kHz, 10 kHz and 100 kHz. Results are presented in Table II. As expected, the resistance of each subconductor increases with the frequency, and is no longer approximated by the DC resistance. However, the increment is only measurable in the hundredth of kHz and even that is not drastic enough to make these solutions useless for higher speeds applications. Other conclusion to be pointed out is that, contrarily to what is expected in high frequency systems with parallel wires, there still is no significant difference between the impedance of each subconductor, proving the utility of the transposition.

Table II - Comparison between high and low frequency resistances.

<table>
<thead>
<tr>
<th>Winding type</th>
<th>Low frequency [mOhm]</th>
<th>1 kHz [p.u.]</th>
<th>10 kHz [p.u.]</th>
<th>100 kHz [p.u.]</th>
</tr>
</thead>
<tbody>
<tr>
<td>CTC 4</td>
<td>5.27</td>
<td>1</td>
<td>1</td>
<td>1.12</td>
</tr>
<tr>
<td>CTC 6</td>
<td>7.03</td>
<td>1</td>
<td>1</td>
<td>1.09</td>
</tr>
<tr>
<td>CTC 8</td>
<td>8.80</td>
<td>1</td>
<td>1</td>
<td>1.07</td>
</tr>
<tr>
<td>CTC 10</td>
<td>10.56</td>
<td>1</td>
<td>1</td>
<td>1.05</td>
</tr>
<tr>
<td>Roebel 4</td>
<td>7.02</td>
<td>1</td>
<td>1</td>
<td>1.05</td>
</tr>
<tr>
<td>Roebel 6</td>
<td>8.77</td>
<td>1</td>
<td>1</td>
<td>1.05</td>
</tr>
<tr>
<td>Roebel 8</td>
<td>10.53</td>
<td>1</td>
<td>1</td>
<td>1.04</td>
</tr>
<tr>
<td>Roebel 10</td>
<td>12.28</td>
<td>1</td>
<td>1</td>
<td>1.03</td>
</tr>
</tbody>
</table>

It is very difficult, if not impossible, to build a three-dimensional model of a Litz wire. Even if a draw was achieved, the software would most likely not be able to cope with the large number of subconductors involved and, most important, with their high degree of transposition. However, in order to have some term of comparison between the transposed winding modeled in this report and the more conventional Litz wire solution, the proximity effect test described before was applied to one turn of the machine. Basically, sets of circular subconductors with 0.25 mm of radius were spread over half surface of the cross section to account for the filling factor. This resulted in a model with 97 strands organized in horizontal rows of 5 or 4 subconductors. Fig. 13 shows the distribution of the magnitude of current density vector over the cross section when performing under the rated frequency. As it can be seen, contrarily to what happens in the CTC and Roebel transpositions, there is no symmetry in the model, which results in every strand having its own unique resistance distinct from the other subconductors. The average resistance per strand at rated frequency is 0.0855 mOhm, which means $P_{j\text{str}} = R_{str}i_{str}^2 \approx 0.086$ mW of Joule losses per strand or a total resistive loss of $P_{\text{str}} = \text{Total strands}.P_{j\text{str}} \approx 8.3$ mW in the modeled half turn, as circulating currents are completely absent in this kind of technology.

2) Circulating currents

Although it is simple to theoretical formulate the problem, performing in its completeness the task of determining the circulating currents between strands is not straightforward, due to the complexity of a design loyal to the real environment, and the subsequent heavy computational effort required. As already seen, circulating currents are mostly function of the induced voltage on each subconductor, which is directly related with the time derivative of the flux linked with each subconductor. The higher the differences between these induced voltages are the more significant circulating currents turn out to be. What happens in the real world is that a complete roll of winding is purchased to achieve a determined transposition degree in the active region. However, for price reasons, wire from the end winding region is also made of the same roll, which means that the strands are not short-circuited every half turn. In fact, the whole phase circuit is stranded even when it travels from one coil to the other. In this analysis, only one half turn is considered, which means that the impedance of each strand disregards the influence of: 1. other turns in the same coil; 2. end winding region; and 3. surrounding coils. Naturally, the influence of these elements’ magnetic induction on the flux linkage of the half turn is also not considered. More importantly, the fact that each turn is short-circuited in the active region considerably decreases the voltage drops between strands, which means that results for the circulating currents presented next should not be seen quantitatively, but as a comparator between technologies instead, since the conclusions drawn are extensible to a real environment situation. As a result, the approach followed to solve the problem is to determine the time variation of the flux linked with each strand by rotating a segment of the rotor along its radial length. Then, as the obtained plots are periodic in terms of the whole machine, circulating currents can easily be determined by performing Fourier analysis on them and solving system described in section II.A.1 of this paper, using phasor notation instead of time dependent functions.
3) Loss results

The resistive loss is calculated individually for each strand, including the circulating current, before adding it up to find the loss of the half turn. Total resistive loss of the machine is simply given by equation (9), where \( N \) is the number of turns, \( m \) is the number of phases and \( p \) is the number of pole pairs. Results are presented in Table III.

\[
P_j^\text{Total} = 2P_j^{\text{half-turn}} \cdot N \cdot n_e = 2\left(\sum P_j^{\text{strand}}\right) \cdot N \cdot (mp) \tag{9}
\]

<table>
<thead>
<tr>
<th>Transposition type</th>
<th>Total loss [kW]</th>
</tr>
</thead>
<tbody>
<tr>
<td>CTC – 2 strands per row</td>
<td>498.4</td>
</tr>
<tr>
<td>CTC – 3 strands per row</td>
<td>443.8</td>
</tr>
<tr>
<td>CTC – 4 strands per row</td>
<td>418.3</td>
</tr>
<tr>
<td>CTC – 5 strands per row</td>
<td>399.0</td>
</tr>
<tr>
<td>Roebel – 2 strands per row</td>
<td>665.7</td>
</tr>
<tr>
<td>Roebel – 3 strands per row</td>
<td>570.0</td>
</tr>
<tr>
<td>Roebel – 4 strands per row</td>
<td>501.0</td>
</tr>
<tr>
<td>Roebel – 5 strands per row</td>
<td>467.7</td>
</tr>
<tr>
<td>Litz wire</td>
<td>0.34</td>
</tr>
</tbody>
</table>

B. Rotational loss

Considering a straight conductive filament exposed to a magnetic field whose variation is the same along both filament’s length and thickness, then the eddy loss in such conductor can be calculated directly by equation (8) along the whole volume. However, in an ironless stator AFPM machine, the flux density waveform is not a pure sinusoid. This fact, along with the air gap magnetic field geometry exhibiting a 3-D nature (i.e. different behavior along machine’s axial and radial directions), implies that calculating eddy current losses merely analytically is subjected to significant errors [4]. Moreover, in a transposed winding, field orientation taken for the calculation varies according to strand positioning. Taking as reference the subconductor axial section, the two magnetic field components of interest for (8) must be parallel to strand’s section height and width, which are not the same as the axial and tangential components seen from the machine perspective. For instance, magnetic field coordinates in a strand shifting from one row to the other are different from the ones in a strand shifting upwards. Consequently, the analysis of each subconductor has to be performed discretizing its volume in such a manner that the above particularities are accounted for. Using basic trigonometry on strand’s segment crossed by the slice under analysis, it is possible to find magnetic field coordinates according to strand’s section.

Fig. 14 illustrates the approach chosen for volume discretization. To account for the flux density waveform in the air gap being close to a trapezoid, which varies along the axial direction of the machine, the concept of layer is introduced. To account for the flux density varying along the radii of the machine, the concept of slice is introduced. Each slice is an arc along the pole pitch and is built from several layers, which basically are copies of each other displaced along strand’s cross section thickness. In the active region, group of turns from 1 to N/2 and from N/2+1 to N experience the same flux density variation. Consequently, losses in upper and lower turn groups can be inferred solely from turns 1 and N, respectively. This is not valid in the end winding region, since the perimeter of the arc connecting half turns is larger in outer turns (e.g. turns 1 and N) than in inner turns (e.g. turns N/2 and N/2+1). Although this points to studying each strand of each turn individually, this would increase computational efforts unworthily, since field variation in end region is negligible when compared to the active zone. Likewise, differences between turns in there may be ignored and the whole analysis reduced to turns 1 and N, referred as lower and upper turn, respectively.

![Fig. 14. Slices and layers (orange) of a strand (green) in a CTC.](image)

The procedure followed to calculate eddy rotational losses is described below:

i. discretize strand length into slices;

ii. discretize strand height into layers;

iii. extract axial and tangential magnetic field along each layer;

iv. compute an average layer representative of the segment of the conductor in play;

v. for the average layer, perform Fourier analysis along pole-pitch for axial and tangential magnetic field to obtain \( \overline{B_y} \) and \( \overline{B_z} \) of equation (8);

vi. integrate equation (8) over the dimensions of the conductor in play to obtain the loss within one slice;

vii. loss within the strand (half turn) is the loss of all slices, where \( s \) is the total number of slices:

\[
\Delta P_{\text{str}} = \sum_{k=1}^{s} \Delta P_{\text{slicek}}
\]

viii. loss within the half turn is the sum of the loss in each of its strands, where \( sT \) is the total number of strands:
\[
\Delta P_{\text{turn}} = \sum_{k=1}^{n} \Delta P_{\text{turn}_k}
\]  

(i) Loss within the phase coil is the loss of every turn:

\[
\Delta P_{\text{phase}} = 2 \times \left[ \frac{N}{2} \Delta P_{\text{phase}_1} + \frac{N}{2} \Delta P_{\text{phase}_2} \right]
\]  

(x) Total loss is the sum of losses in every coil, where \( nc \) is the total number of coils, given by the product between the number of phases and the number of pole pairs:

\[
\Delta P_{\text{total}} = nc \times \Delta P_{\text{phase}}
\]

Due to the rotary symmetry, only one pole-pitch is necessary to model the entire machine [4].

Fig. 15 and Fig. 16 illustrate the magnetic induction vector and magnitude, respectively, along selected layers of the machine.

**VI. CONCLUSIONS**

Obtained results are quite clear and, as already expected before the study, the conclusion is that both proposed technologies are still far from delivering the loss reduction performance provided by Litz wire. The reason behind this is the considerable difference in the cross section surface available for the current to flow. As in Roebel winding and CTC additional space is needed to make the transposition physically possible, there is a waste of the available surface, which leads to an increase in the winding resistance and a subsequent rise in the resistive loss. Naturally, this waste is larger the lower the number of strands is, since surface of the strands is bigger. As Roebel transposition requires one more strand of additional space than CTC, resistive losses have the pace presented in Fig. 19, where a direct relation between this kind of loss and the available conductive surface is easily noted. On the other hand, the much smaller cross section surface of subconductors in Litz wire implies this technology is using almost ten times more strands than the best solutions of the other two options, reducing the resistive loss to an unreachable competitive edge.

A curious and, at first, unexpected result of this work is the inverse proportionality between the number of strands and obtained rotational losses, as illustrated in Fig. 19. It would seem logic to think that if subdividing the conductor in several subconductors serves the purpose of reducing eddy current effects, then the more intense that division is, the sharpest rotational loss reduction would
be. It is true that using a stranded conductor instead of a massive one provides better rotational loss performance, as it will be shown further in this section. However, since a stranded wire is built from several parallel circuits, transposition has to be performed in order that the induced EMFs in all parallel conductors are equalized, consequently reducing circulating currents. When the conductor is transposed, the changes in its dimensions orientation along the transposition path may cause the components of the flux that contribute to the rotational loss to mutual increase each other. Understandably, components of the flux may also align in a way they cancel each other out. The number of winding direction changes is proportional to the number of strands, and what these results indicate is that, in this two specific types of transposition, most direction changes in transposition steps cause the flux components to increase their effective value for the rotational loss. Another fact that also contributes to the increase of rotational loss with the number of strands is the available conductive surface, which is larger in the cases with more sub conductors. Fig. 19 indicates that, at the end, an apparent random pace is followed by total winding loss, which implies a balance must be made between resistive and rotational loss in order to find the best solution for given dimensions. Nevertheless, it is not clear that the process of subdividing winding conductors is work that pays off in terms of total winding loss reduction.

An issue that should be discussed is the advantages of using a three-dimensional simulation. It should be pondered if it is possible or not to achieve similar results with a two-dimensional model and if the computational effort spent is worthwhile. Rotational losses calculation in AFPM machines is usually described in the literature (e.g. [4]) as a two-dimensional problem. However, following the same method in a three-dimensional environment provides more accurate results not only due to the flux density varying along the radii of the machine, but also because it is not possible to account for the magnetic flux variation within a transposed coil using a two-dimensional model. In terms of resistive loss, there is now no doubt similar results could be achieved merely by recurring to the analytical expression that gives the resistance of a conductor. This happens because of the low frequencies under study. Basically, skin, proximity and spirality effects are unnoticeable and the impedance of the winding is simply only a function of its dimensions. However, the extremely heavy three-dimensional simulation run was not in vain since self and mutual inductances of the subconductors were needed for circulating currents determination. Highly dependent on the geometry, this parameter requires a three-dimensional model.

Fig. 19. Total winding loss balance.

REFERENCES