Eddy-Current Losses evaluation in hairpin wound motor fed by PWM Inverter

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Abstract—This paper presents different approaches proposed to evaluate eddy-current losses in hairpin wound motor fed by pulse width modulated inverter, underlining strengths and weaknesses of everyone and applying them, as test-cases, to a 200kW induction motor designed for a premium electric vehicle.

The first three approaches are based on the finite element analysis: the first, the more accurate one, uses a transient analysis that request long execution time and high memory usage, the second is based on tool that permits rapidly to evaluate the steady state conditions imposing a sinewave current and the third uses a single-slot simplified model. The fourth approach proposes a rapid monodimensional analytic method based on hypotheses that permit to consider the superposition effects of the Eddy-Current Losses of every current harmonic component. The fifth approach makes uses of the proposed analytic method by leveraging a rapid investigation of current harmonic content due to pulse width modulation based on a lumped parameter motor model.

In addition, the paper discusses important considerations about the use of the hairpin technology fed by pulse width modulation that can be drawn from the analysis of the obtained results.

Keywords—Induction motors, Electromagnetic model, Finite element analysis, Cage rotors, Motor simulation

I. NOMENCLATURE

n_s number of wires in the slot  
l_s stator stack length  
l_e half turn overhanging portion  
i imaginary unit  
l rms value of the wire current  
b bar wire width  
w slot width  
h bar wire height  
n number of the considered wire  
m magnetic permeability of the conductor material  
\rho electrical resistivity of the conductor material

II. INTRODUCTION

The interest in hairpin wound motor has increased in recent years and today they are used in several application, often driven by Voltage Source Inverter (VSI), [1]-[2]-[3]-[4].

In line with the new demands of power density improving of electrical machines, the high fill factor achieved by the hairpin technology represents a key feature for numerous applications, one for all the automotive field, [5]-[6]-[7]. On the other hand, eddy current losses due to skin and proximity effects due to the high operating frequency can represent an important drawback with a large cross-sectional area. In this context an accurate and rapid analysis of the eddy current losses introduced by the harmonic generated by the space vector Pulse Width Modulation (PWM) is very important to consider relevant effects at the design stage in order to let the stator geometry and hairpin winding be designed as a whole, to optimize the slot height and the number of layers, avoiding ineffective exploitation of copper. An example of such slot geometry ineffective design can be found in [1], where partial emptying of the slot, having a given non optimized height, is even beneficial at high frequency.

A one-dimensional eddy current analytic model for rectangular conductors of coils for turbogenerators was introduced in [9], which has been applied to transformer windings by [10] and [11] and to slot-bound series-connected conductor portions by [12]. The ac to dc resistance ratio increases due to fast changing non-sinusoidal current waveform during the commutation of a dc machine was already calculated in [13]. However a more general extension of the cited works [8]-[13] to non-sinusoidal current waveform can be found in [14]-[15], where the additional copper losses for every harmonic component, calculated by means of the cited methods, are superimposed to each other for electronic applications (dc–dc converters and transformer windings).

Since many years, the Finite Elements Analysis (FEA) tools permit the calculation of winding losses even if it requests much computational burden, leading to high time cost for coupling motor geometry optimization considering the winding losses computation. This computational cost drastically worsens if the designer intends to run FEM simulation accounting for PWM voltage supply [16].

This paper presents different approaches proposed to evaluate eddy-current losses in hairpin wound motor fed by pulse width modulated inverter and applied to an Induction Motor (IM) for vehicle traction application as test case.
In section III, the considered hypotheses and the equations of a proposed analytical method is presented: it is based on the principle of superposition of effects for the Eddy-Current Losses due to the harmonic components of the periodic distorted non-sinusoidal current, recalling the mentioned theory.

In section IV, a Lumped Parameter (LP) model is used and discussed as a tool for rapid determination of the harmonic content of the current with PWM VSI described in section V.

Finally, in section VI, different approaches, based also on the equations and model described in the previous sections, are explained and implemented to evaluate the Eddy-Current Losses of a proposed analytical method is presented: it is based on the principle of superposition of effects, as described below.

By integrating the product of the current density over the wire cross section, the power loss in the n-th wire can be calculated, as follows:

\[ P_{acn} = \rho blx \int_{0}^{h} j \cdot J^* dx \]  

By extending the sum of the result of the previous integration to all the \( n \) wires in the slot, then dividing the overall result by the number of works, e.g. [9], which is the earliest at our knowledge.

\[ k_{ac} = \frac{R_{ac}}{\rho} \frac{h}{\delta} \left[ \frac{\sinh 2\frac{h}{\delta} + 2}{\cosh 2\frac{h}{\delta}} \right] \frac{2}{3} \left( nL^2 - 1 \right) \frac{\sinh 2\frac{h}{\delta} - \sin 2\frac{h}{\delta}}{\cosh 2\frac{h}{\delta} + \cos 2\frac{h}{\delta}} \]  

Assuming sinusoidal variation over the time for the variables, the eq. 5 and eq. 6 have solutions in eq. 7 and eq. 8, respectively, which can be obtained by the same calculation procedure as in [10] and are valid for the n-th wire conductor in the slot. The first term in the numerator at their second member depends on the current in the n-th wire itself as it was alone in the slot, the second term accounts for the effect of currents in the n-1 wires piled up between the slot bottom and the considered one.

\[ P_{ac} = \rho \frac{\partial^2 H}{\partial x^2} = \mu \frac{\partial H}{\partial t} \]  

\[ P_{ac} = \rho \frac{\partial^2 j}{\partial x^2} = \mu \frac{\partial j}{\partial t} \]  

The eq. 3 and eq. 4 can be obtained from Maxwell equations eq. 1 and eq. 2, respectively.

\[ \nabla \times H = j \]  

\[ \nabla \times E = -\frac{\partial B}{\partial t} \]  

\[ \frac{\partial H}{\partial x} = j \]  

\[ \rho \frac{\partial j}{\partial x} = \mu \frac{\partial H}{\partial t} \]  

where \( x \) is the coordinate of an axis with origin at the bottom edge of the bare wire, parallel to slot walls and oriented toward the slot opening and \( n \) is the number of the considered wire, with numbering order ascending from bottom to top of the slot.

The eq. 3 and eq. 4 describing the magnetic field (current density) in the wire conductor:

\[ \rho \frac{\partial^2 H}{\partial x^2} = \mu \frac{\partial H}{\partial t} \]  

\[ \rho \frac{\partial^2 j}{\partial x^2} = \mu \frac{\partial j}{\partial t} \]  

with \( \delta = \frac{\rho w}{\pi f \mu b} \) defined as the skin depth at the frequency \( f \).
IV. Rapid Investigation of Harmonic Content of the Current Due to PWM in IM

Considering the hypotheses exposed in section III, the Eddy-Current Losses due to the periodic distorted non-sinusoidal current produced by the PWM VSI supply can be considered as a superposition of effects of the Eddy-Current induced by every sine wave of the current harmonic spectrum (content that consists of integer multiples of the fundamental frequency).

An analysis methodology that allows a rapid evaluation of the harmonic content with different PWM strategies is described and compared with respect to transient FEA model in terms of accuracy and calculation time.

In particular, the Current-Ripple due to the PWM is determined by imposing the modulated voltage to a constant LP model for IM. To better clarify and explain these approach, a thorough analysis is presented below.

LP models permits evaluating the transient performance of electric drives, [17]. They consist in software solutors of a set of non-linear differential equations with generally constant electric drives, [17]. They are defined as 

\[
\frac{dI_{st}}{dt} = -\frac{1}{T_i}I_{st} + \frac{K_r}{\sigma L_{sr}}\Psi_{ra} + \frac{\sigma}{\Delta x^2}\frac{d\Psi_{ra}}{dx} + \frac{V_{st}}{\sigma L_{sr}}
\]

\[
\frac{dI_{gb}}{dt} = -\frac{1}{T_i}I_{gb} + \frac{K_r}{\sigma L_{sr}}\Psi_{rb} + \frac{\sigma}{\Delta x^2}\frac{d\Psi_{rb}}{dx} + \frac{V_{gb}}{\sigma L_{sr}}
\]

\[
\frac{d\Psi_{ra}}{dt} = M_T I_{st} - \frac{1}{\tau_{r}}\Psi_{ra} - \omega_r \Psi_{rb}
\]

\[
\frac{d\Psi_{rb}}{dt} = M_T I_{gb} - \omega_r \Psi_{ra} - \frac{1}{\tau_{r}}\Psi_{rb}
\]

where \(I_{st}, I_{gb}, \Psi_{ra}, \Psi_{rb}, V_{st}, V_{gb}, \omega_r, \sigma\) are respectively the \(\alpha-\beta\) stator currents and voltages, the rotor fluxes and rotor electrical speed.

The parameters \(T_i, \sigma, T_r, K_r\) are defined as \(T_i = \frac{\pi^2 r_{st}^2}{\tau_{st} - \tau_{sr}}\), \(\sigma = 1 - \frac{M^2}{L_{sr} L_{sr}}\), \(T_s = L_s / R_s, T_r = L_r / R_r\), \(K_r = M / L_r\), where \(M, L_s, L_r, R_s, R_r\) are respectively the magnetizing, the stator and rotor inductances, the stator and rotor resistances. \(L_s\) and \(L_r\) can be express as sum of the magnetizing and the leakage inductances \(L_{ss}\) and \(L_{sr}\) respectively.

The electrical coupling of phase quantities with the \(\alpha-\beta\) orthogonal model quantities is based on the Clarke transformation. The \(\alpha-\beta\) voltage, input of the model, can be obtained in function of the time using the modulated three-phase voltage as described by the following relations:

\[
V_{st} = \frac{2}{3}V_a - \frac{1}{3}V_b - \frac{1}{3}V_c
\]

\[
V_{gb} = \frac{\sqrt{3}}{3}V_a - \frac{\sqrt{3}}{3}V_b - \frac{\sqrt{3}}{3}V_c
\]

where \(V_a, V_b, V_c\) are the three phase voltages.

The parameters in the differential equations (14)-(17) can be kept constant, as in the validation of the next section, or more accurately considered as function of a set of the state variables as suggested in the rest of the section.

In particular, a variation of the magnetizing inductance in function of the magnetizing current permits to consider with good approximation the effect of the material saturation for the following case of study the inductance \(M\) is constant in view of the fact that the machine is affected very little by saturation in all analysed working points). The rotor resistance can be varied as function of the slip: the approach to evaluate the rotor resistance for different frequencies can be analogous to the analytical one proposed in section III but it’s important to emphasise that, considering the effect of space non-fundamental harmonics neglected in the used LP model, the work frequency of the rotor bar currents and so their skin effects are generally low (the resistance is close to the DC value), [18].

The stator resistance, as seen in the previous section, can significantly vary with the frequency. Considering a constant stator resistance in the transient LP model, as in the test case validation in the next section, means considering the same value for each frequency, which can mean having a not good approximation in the transient dynamic for certain analysis, see the case study results in this regards.

An interesting alternative, subject of a subsequent article, is represented by the harmonic decomposition of the PWM voltage waveform and the application of each harmonic in a steady-state equivalent model with variable parameter as function of the frequency (as the stator and rotor resistances) or the state-variables amplitudes (as the magnetizing inductance).

In order to obtain the magnetizing inductance value considering all the current harmonic content that must also be adjusted, an iterative procedure can be considered.

V. Case Study and Validation

The case of study refers to a motor that has been developed in the frame of the Horizon 2020 project ReFreeDrive “Rare earth Free e-Drives featuring low cost manufacturing” as the traction engine of high power 200kW electrical vehicles [19].

Key points of the design have been: 1) cost reduction through the minimization of the motor size; 2) optimized shape of the windings by hairpin technology, which results in an average efficiency over the drive cycle greater than 94% (considering WLTP class 3 drive cycle) due to high slot fill factor and effective heat dissipation. Fig. 2 shows the cross section and main geometric data of the three-phase IM considered in this study, with its rated point performance summarized in table 1
TABLE I. Rated working point performance

| Rated point values                  |
|------------------------------------|
| Current 589 A (peak)              |
| Voltage 337 V (peak)              |
| Frequency 207 Hz                  |
| Flux 0.25 Wb                      |
| Torque 340 Nm                     |
| Speed 6000 rpm                    |
| Slip 0.0339                       |
| No-load current 224 A (peak)      |
| PWM Frequency 10 KHz              |

Table 1 shows the main geometric data of hairpin winding, while the lumped parameters of the three-phase IM considered in this study are summarized in Table II.

VI. RESULTS AND COMPARISONS

A Comparison of different approaches to evaluate the Eddy-Current Losses in the proposed case of study is explained and implemented.

For all the approaches, a widely employed Space Vector PWM strategy is considered, with symmetrical voltage pattern as regards the vertex of a triangular carrier, leading to seven changes of the feeding voltage in a PWM period. The voltage source inverter has been modelled by using ideal switches. The first approach is the more accurate one and is based on a “Full FEA” method: the Voltage PWM waveform is applied directly to the stator wire regions in a transient FEA tool that permit to evaluate the Eddy-Current Losses [20].

The most important drawback of this approach is the long execution time and high memory usage due to the need for highly discretized mesh in the wire and small time step (2e-7 s for simulation time greater than 100 ms) constrained by the high switching frequency. In fig. 4 the Ohmic-Losses distribution in the hairpin is shown when the steady state is reached. In particular, the figure can help to evaluate as the hypothesis of the orthogonality of the flux lines used in section III is necessarily an approximation.

The second approach is based on the use of a tool that permit to rapidly evaluate the state-space Eddy-Current Losses imposing a sinewave current and based again on FEA (Motorcad® in our case) [21]. Calculating the Eddy-Current Losses for every harmonic component, the total Eddy-Current Losses due to the PWM modulation can be well estimated by adding all the calculated contributions.

In particular, using the harmonic content of the current evaluated by “Full FEA” approach is possible to evaluate the good accuracy of the linear hypothesis even when the hypotheses of the section III are not considered. This second approach can be useful also to have a rapid and accurate estimation of Eddy-Current Losses due to the PWM modulation when the harmonic content is rapidly estimated by LP model, as explained in section IV and implemented as last approach. Fig. 5 show the magnetic field density that can be rapidly calculated imposing the fundamental harmonic by Motorcad® and can be “freeze” to evaluate the Eddy-Current Losses contribute for every other current harmonic component, assuming that do not contribute to change the saturation state.

A third compared approach can be called “single slot” one: it is based on static FEA but with a geometry very simplified that consists of only one slot with opportune boundary conditions without the presence of the rotor. With these approximations the mesh is sensibly reduced respect the first approach and there’s no need to wait for the steady state. This approach permits to evaluate how such approximations affect the evaluation of Eddy-Current Losses with an eye towards an analytic approach that doesn’t consider the saturation effect of the iron and the field induced by the rotor and other slots. The fig. 6 shows the geometry considered for this approach and the relative mesh.

Fig. 4. Ohmic-Losses distribution in the hairpin at the steady state

Fig. 5. Magnetic field density imposing the fundamental harmonic

Fig. 6. Geometry considered for this approach “Single Slot”
The fourth approach, unlike previous ones that are based on FEA method, estimates the Eddy-Current Losses with the analytical method described in section III and so as superposition effects of the Eddy-Current Losses evaluated by formulas 10. The fig 7 shows a comparison of the second, third and fourth approaches for the rated point reported in table II. In particular, fig. 7 reports the evaluation of the Eddy-Current Losses for each harmonic considering the same current harmonic spectrum estimated by the first “full FEA” approach (that estimate a total Eddy-Current Losses of 12868 W).

Fig. 7. Comparison of the second (green), third (purple) and fourth (cyan) approaches for the rated point

Fig 7 shows how the high-frequency components (in particular the 48th and 50th component) due to the PWM modulation give a non-negligible contribution in relation to the total Eddy-Current losses.

As last compared approach, the method proposed in section IV, based on the constant LP reported in table II, is implemented to evaluate the current harmonic content and the analytic method proposed in section III is used to estimate the Eddy-Current Losses due to the PWM modulation. Fig. 8 represents a comparison between the current evaluated by “full-FEA” and the last approach based on LP in function of the time, fig. 9 shows the correspondent harmonic content, useful to evaluate the Eddy-Current Losses.

Fig. 8. Comparison between the current evaluated by “full-FEA” (blu) and the LP approach (orange) in function of the time

As fig. 8 shows, the correspondence is not perfect due the approximation introduced by the LP model respect the “full FEA”, about the constant stator and rotor resistances and the saturation to mention only the more important, but even due to the cross-coupling and a first-harmonic model effects.

Table III shows an exhaustive comparison of Eddy-Current losses (and the percentual error respect the more accurate “full FEA” approach) for all the described approaches in different working points (constant torque zone until 6000 rpm and flux-weakening zone at 20000 rpm). It also shows the calculation time necessary to obtain the losses, already knowing the current spectrum (please remember that in the ”full FEA” case the losses are calculated at the same time as the current calculation and in the ”Analy. + LP” case the current spectrum is calculated by LP model with 48 s).

| Speed [rpm] | Torque [Nm] | Full FEA | SS FEA | Single slot Analy. | Analy. + LP |
|-------------|-------------|----------|--------|------------------|-------------|
| 500         | 34          | 77        | 74     | 72 (6.5%)        | 70 (8.4%)   |
|             | 170         | 1532      | 1489   | 1419 (7.4%)      | 1333 (13%)  |
|             | 340         | 6371      | 6046   | 5937 (6.8%)      | 5772 (9.4%) |
|             | 34          | 128       | 125    | 124 (3.4%)       | 122 (4.6%)  |
| 3000        | 170         | 2942      | 3025   | 2976 (1.2%)      | 2878 (2.2%) |
|             | 340         | 9316      | 9925   | 9830 (5.5%)      | 9640 (3.5%) |
| 6000        | 34          | 233       | 228    | 219 (5.5%)       | 210 (10%)   |
|             | 170         | 3808      | 3438   | 3352 (12%)       | 3180 (16%)  |
|             | 340         | 12868     | 12069  | 11876 (7.7%)     | 11491 (11%) |
| 20000       | 9           | 221       | 212    | 199 (9.8%)       | 194 (12%)   |
|             | 45          | 2723      | 2625   | 2467 (9.4%)      | 2380 (12.6%)|
|             | 90          | 7787      | 7561   | 7289 (6.4%)      | 6938 (10.9%)|
| Computational time | 8 days | 6 hrs | 35 min | 30 s | 30 s |

Table III. Eddy-Current losses (and percentual error respect “full FEA”) and computational time for different methods

Fig. 9. Comparison between the harmonic content of the current evaluated by “full-FEA” (blu) and the LP approach (orange)

This non-correspondence in function of the time has obvious influence on the harmonic content, as shown in Fig. 9, and so on the evaluation of Eddy-Current losses.
As can be seen, the proposed approaches are less and less precise in the order in which they were proposed. Note that the computational demand grows with increasing accuracy. To confirm this, the worst error for the different approaches are respectively: 9.7%, 12%, 16% and 29%. Certainly, the precision of the totally analytical method can be increased with the use of more accurate dynamic models, as suggested in section IV.

By monitoring the total Eddy-Current losses compared to the ones due to first harmonic alone, an important contribution in the total losses is given by the modulation. In fact, even if the amplitude of the harmonic components due to the modulation is modest, their high frequency contributes to increase the losses attributed to them.

VII. Conclusions

The present work aims at presenting different approaches to evaluate the eddy-current losses in hairpin wound motor fed by PWM Inverter and at validating them by means the transient FEA method, more accurate but even more wasteful in term of computational resources. A test case of IM for automotive application is considered and the important contribution of the losses due to the frequencies introduced by the PWM is underlined. The comparison also confirms the correctness of using the superposition of effects to simplify the evaluation of eddy current losses. In particular, a proposed analytical approach based on the mentioned theory and lumped parameter model is useful to evaluate them with reasonable approximation and minimal computational cost. Suggestions are also proposed to improve accuracy of the transient analysis based on the lumped parameter and will be the subject in next articles.

The good agreement between FEA and analytic model confirms that the rotor field has a negligible effect on the stator winding eddy current losses, for rectangular slot form-wound or hairpin winding, as already envisaged in [8, 22-23]. Therefore, future works could be based on the present and previous works [23], to set experimental winding loss determination methods for motors fed by PWM voltage, being their rotor removed.

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