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Improved Analytical Model of Induction Machine for Digital Twin Application

Victor Mukherjee, Tatjana Martinovski, Aron Szucs, Jan Westerlund, Anouar Belahcen (Senior Member, IEEE)

Abstract – This paper presents a saturable analytical model of induction machines, with a systematic approach for segregating the electromagnetic losses. The proposed model is based on the equivalent circuit of the machine, which has been augmented to account for different loss components. The segregation of different loss components in stator and rotor have been improved by considering the loading, skin effect and field-weakening operation. The required parameters of the model are identified from a well-defined set of finite element analysis of the machine. The proposed model and the identification methodology have been tested on two different induction machines. In both cases, the finite element computations are validated with laboratory measurements, and the analytical model is validated against the finite element model. The results show that the proposed model is more accurate than the conventional ones from the literature, and thus it can be used as a component while building a digital twin of the induction machine, and inserted into a virtual simulation software for real-time thermal management for example.

Index Terms– Digital twin, equivalent circuit, induction machine, iron loss, magnetic saturation, skin effect.

I. INTRODUCTION

INDUCTION machine is one of the most widely used machines in the low and medium voltage section. Over the decades, numerous studies have been conducted to model induction machines with analytical and finite element (FE) methods [1]. Analytical models are fast and can be used to do the preliminary design of the machine [2], which is later improved with computationally heavy FE based models. Also, for real-time application, the analytical model can be quite useful as the computational time is equally important with computational accuracy. The analytical models can be broadly classified into two types- a. physical model and b. equivalent circuit-based model. The physical model contains different physical machine parameters like stator and rotor dimensions, airgap, winding information etc. The primary way of using the physical model is to solve the flux density mainly in the airgap and sometimes also in other parts of the

machine. Following that, post-processing is done to solve the torque and machine losses [3]. However, for certain applications like the digital twin of an electrical machine, this can be problematic.

A digital twin of an electrical machine is a relatively new concept, and it can mean different models based on the requirement of the applications. Normally in the context of an electrical machine, digital twin refers to a model, which is more accurate than traditional analytical model and takes significantly less time to compute than the FE model. Normally, a digital twin model is used in a real-time system where the sampling rate can be 10-100 μ s. A digital twin can be used for condition monitoring as well as for large-scale system optimisation. In this work, the digital twin is developed for a large-scale system optimisation which can approximate the loss and temperature distribution of different machines in a multi-machine industrial system. Since the system-level optimisation happens between different components like drives, motors, loads etc., it is preferred to use the least physical information while developing the model for the machine. From this perspective, the equivalent circuit-based model can be a very good solution, as it does not require too many physical parameters of a machine. However, additional efforts are required to compute the iron losses with the conventional equivalent circuit model.

The iron losses in an electrical machine are widely computed using Jordan's or Bertotti's model. Jordan's model divides the iron loss into two components, which are the eddy current losses and the hysteresis losses. In [4], a traditional Γ -equivalent circuit model is extended with an iron loss resistor to take into account the eddy current losses and the hysteresis losses in the constant power and field-weakening region. The advantage of this model is its capability to compute the losses dynamically, i.e. as a function of time. However, in this model, the total iron loss of the machine is computed without segregating the stator and rotor iron loss components. Also, this model does not consider the effect of the loading. It is shown in [5], that the loading of the machine has a significant impact on the losses of the induction machine. With increased loading, the top of the rotor slot openings saturates more due to the influence of the stator slot harmonics and thus the losses increase. In [6] and [7], the loss segregation at loading and no loading are proposed. However, the losses are not separated into eddy current and hysteresis loss components. In addition, the stator losses and the rotor losses, which increase with the loading, are taken into account together as a stray load loss component. The disadvantages of this approach are the vague distribution of the losses in the machine along with the

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limitation of operational points like ignoring the field-weakening region. As a result, the model cannot be used properly in a thermal network for a wider operational range. To resolve the above-mentioned issue, this work provides a systematic way of loss segregation in the different parts of the machine, through an improved equivalent circuit model of the induction machine. The work presented in this paper is an extension of the work presented in [4]. The contribution of the paper can be listed as:

- (a) The drawback of the model in segregating the losses in the stator and rotor along with its distribution is overcome. The effect of loading and rotor skin effect is taken into account separately by introducing new components which model the extra losses.
- (b) The Γ -equivalent circuit is improved by adding an additional resistance component for modelling the rotor iron loss. Moreover, the rotor losses in the field weakening region are identified.
- (c) The developed model is suitable to be used as a component to build a digital twin of induction machines. The model can provide input as losses for the thermal model in a system design in real-time. The accuracy of the loss computations is verified with two differently rated induction machines with measurement and FE computations.

II. INDUCTION MACHINE MODEL FROM LITERATURE

A. Γ Model of Induction Machine

The Γ -equivalent circuit model from [4] is shown in Fig. 1. The machine model is given as,

$$\frac{d\psi_s}{dt} = \mathbf{u}_s - R_s \mathbf{i}_s - \omega_s \mathbf{J} \psi_s \quad (1a)$$

$$\frac{d\psi_r}{dt} = \mathbf{u}_r - R_r \mathbf{i}_r - \omega_r \mathbf{J} \psi_r \quad (1b)$$

where the orthogonal matrix $\mathbf{J}$ is given as,

$$\mathbf{J} = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix} \quad (2)$$

ψ_s , ψ_r , $\mathbf{u}_s$, $\mathbf{u}_r$, R_s , R_r are the stator flux vector, rotor flux vector, stator voltage vector, rotor voltage vector, stator resistance and rotor resistance. The induction machine is modelled in a coordinate system which is rotating at an arbitrary angular speed of ω_s whereas the rotor is rotating with an angular slip frequency of ω_r . In this work, the rotor is short-circuited as we are dealing with cage induction machines, i.e. $\mathbf{u}_r = 0$. The iron loss resistance R_{Fe} in Fig. 1 is used to model the total iron loss in the motor. The iron loss current can be computed as,

$$\mathbf{i}_{Fe} = \frac{\mathbf{u}_s - R_s \mathbf{i}_s}{R_{Fe}} \quad (3)$$

The stator and rotor flux linkages are computed as,

$$\psi_s = L_M (\mathbf{i}'_s + \mathbf{i}_r) \quad (4a)$$

$$\psi_r = \psi_s + L_\sigma \mathbf{i}_r \quad (4b)$$

where $\mathbf{i}'_s = \mathbf{i}_s - \mathbf{i}_{Fe}$ and L_M , L_σ is the magnetising and leakage inductance. The magnetising current is computed as $\mathbf{i}_M = \mathbf{i}'_s + \mathbf{i}_r$. The leakage flux is computed as $\psi_\sigma = L_\sigma \mathbf{i}_r$.

Fig. 1. Γ -equivalent circuit as proposed in [4]

B. Identification Methods of Losses

The iron loss resistance is a function of the voltage across the iron loss resistance, and the stator flux magnitude, which are referred to as u and ψ_s respectively. It can be expressed as,

$$R_{Fe}(u, \psi_s) = \frac{R_{Ft}}{1 + k \psi_s^{n-1} / u} \quad (5)$$

where R_{Ft} , k , and n are positive constant model parameters. The no-load iron loss P_{fe} at steady state is computed as a sum of the eddy current loss P_{ed} and P_{hys} , and are given as,

$$P_{fe} = P_{ed} + P_{hys} = \frac{u^2}{R_{Ft}} + \frac{k \psi_s^{n-1} / u}{R_{Ft}} = \frac{\omega_s^2 \psi_s^2 + k |\omega_s| \psi_s^n}{R_{Ft}} \quad (6)$$

The parameters k , R_{Ft} and n are fitted using (6) at different ω_s and ψ_s , using a curve-fitting algorithm. More detail on the implementation will be discussed in sections III and IV while discussing the improved model. With the model in (6), the additional losses due to the loading are ignored. The losses, which can be computed using this model are

- a. the total iron loss P_{fe} as given in (6),
- b. the stator resistive loss given as $P_{res,s} = 3R_s i_s^2$,
- c. the rotor resistive loss given as $P_{res,r} = 3R_r i_r^2$.

In the proposed improved model, as described in section III, the following improvements will be accomplished:

- a. the iron loss in the stator and rotor will be computed separately,
- b. the effect of the loading on the stator iron loss will be incorporated,
- c. the skin effect in the rotor resistive losses will be considered.

III. IMPROVED INDUCTION MACHINE MODEL

A. Proposed Γ Model of Induction Machine

The Γ -equivalent circuit model proposed in this paper is shown in Fig. 2.

Fig. 2. Γ -equivalent circuit proposed in this work.

To compute the iron loss for stator and rotor separately, R_{Fe} in Fig. 1 is split into stator iron loss resistance $R_{Fe,s}$ and rotor iron loss resistance $R_{Fe,r}$ in Fig. 2. The rotor iron loss is caused mainly due to the slotting effect appearing from the stator side. Hence, $R_{Fe,r}$ is kept in the stator side of the equivalent circuit in parallel with $R_{Fe,s}$ for the simplicity of

the computation. So, the iron loss current i_{Fe} is computed as,

$$i_{Fe} = \frac{u_s - R_s i_s}{R_{Fe,eq}} \quad (7)$$

where $R_{Fe,eq}$ is the equivalent iron loss resistance of the machine, which is computed as $R_{Fe,eq} = R_{Fe,S} \parallel R_{Fe,R}$. From the machine model, the stator and rotor flux linkages can still be computed similarly as shown in (1), (2) and (4).

It is worth to mention that the inductances L_M and L_σ are not considered constant, rather the magnetic saturation of the machine is modelled by considering the inductances as a function of the fluxes [8]. The magnetising inductance and leakage inductance are modelled as,

$$L_M(\psi_s) = \frac{L_{Mu}}{1 + \alpha \psi_s^a} \quad (8)$$

$$L_\sigma(\psi_\sigma) = \frac{L_{\sigma u}}{1 + \beta \psi_\sigma^b} \quad (9)$$

where L_{Mu} and $L_{\sigma u}$ are the unsaturated constant inductance identified from the equivalent circuit as shown in Fig. 1 and 2. The constant parameters α , β and b are identified by the indirect method as described in [8]. In brief, different no-load operational points are computed by varying the input voltages. At no-load, the rotor flux $\psi_R = 0$, which means that after ignoring i_{Fe} , $i_s = i_M$ and $\psi_s = \psi_M = (u - R_s i_M) / \omega_s$. Therefore, the magnetizing inductance can be also written as $L_M = \psi_s / i_s$. Note that i_m , i_s and i_R are the magnitudes of i_s , i_M and i_R respectively. The cost function, which is minimized to identify α and a is given as,

$$f(L_M) = \frac{\psi_s}{i_s} - \frac{L_{Mu}}{1 + \alpha \psi_s^a} \quad (10)$$

For identifying the leakage inductance parameters, different loading points are computed by varying the voltages and the slips [8]. In this case, $\psi_\sigma = \psi_M - \psi_s = (R_R / \omega_R) i_R - \psi_s$. Therefore, the leakage inductance can be also written as $L_\sigma = \psi_\sigma / i_R$. The cost function which is minimised to identify β and b is thus given as,

$$f(L_\sigma) = \frac{\psi_\sigma}{i_R} - \frac{L_{\sigma u}}{1 + \beta \psi_\sigma^b} \quad (11)$$

B. Identification Methods of the Proposed Model

The iron losses are separated for the stator and rotor using $R_{Fe,S}$ and $R_{Fe,R}$ respectively. It is worth to mention that both $R_{Fe,S}$ and $R_{Fe,R}$ are considered as a function of u and ψ_s . Since at no-load, the rotor slip frequency is 0, the influence of the rotor flux is not considered for the computation of the rotor iron losses. The stator and rotor iron loss resistance are expressed as,

$$R_{Fe,S}(u, \psi_s) = \frac{R_{Ft,S}}{1 + k_S \psi_s^{n_S - 1} / u} \quad (12a)$$

$$R_{Fe,R}(u, \psi_s) = \frac{R_{Ft,R}}{1 + k_R \psi_s^{n_R - 1} / u} \quad (12b)$$

where $R_{Ft,S}$, $R_{Ft,R}$, n_S , n_R , k_S and k_R are positive, constant model parameters. The no-load stator iron loss $P_{Fe,S,nl}$ and no-load rotor iron loss $P_{Fe,R,nl}$ are computed as,

$$P_{Fe,S,nl} = \frac{\omega_s^2 \psi_s^2 + k_S |\omega_s| \psi_s^{n_S}}{R_{Ft,S}} \quad (13a)$$

$$P_{Fe,R,nl} = \frac{\omega_s^2 \psi_s^2 + k_R |\omega_s| \psi_s^{n_R}}{R_{Ft,R}} \quad (13b)$$

where $P_{Fe,S,nl}$ and $P_{Fe,R,nl}$ computed by the FE at different stator flux level and frequencies are used to fit $R_{Ft,S}$, n_S , k_S , $R_{Ft,R}$, k_R and n_R . However, the iron losses computed using (13) are not sufficient to take into account the loading effect. To overcome this issue, the extra stator iron losses due to the loading are introduced as $P_{Fe,S,sl}$ where ‘sl’ stands for stray load loss. It can be thus expressed as,

$$P_{Fe,S,sl} = P_{Fe,S,l} - P_{Fe,S,nl} \quad (14)$$

where $P_{Fe,S,l}$ is the stator iron loss at loading. To model $P_{Fe,S,sl}$, it is considered as a function of the slip and stator voltage. Subsequently, it is computed as,

$$P_{Fe,S,sl} = k_1 |\omega_s|^{n_{S1}} u_s^{n_{S2}} \quad (15)$$

Different values of $P_{Fe,S,sl}$ computed at different rotor slip frequencies and stator voltages can be used to fit k_1 , n_{S1} and n_{S2} . The influence of the loading on the rotor iron losses is ignored in the proposed model as the rotor slip frequency is very small to influence the eddy current and hysteresis losses when the machine is operating near the nominal point. However, this can be a source of error if this model is used to identify the losses in the entire torque-speed range when the machine is supplied as direct online.

The rotor bar losses are conventionally computed only from the DC resistance of the bar and ignoring the skin effect in the equivalent circuit. To consider the skin effect, it often requires more understanding of the machine’s physical parameters. One option is to express the rotor resistance with different branches in the equivalent circuit to model the skin effect variations in the different part of the rotor bar [9]. However, to keep the equivalent circuit simple, this option is not used in this work. Since the goal of this work is to use the least machine’s physical parameters, the skin effect is taken into account with a polynomial function of the stator angular speed and stator current amplitude. The rotor bar loss due to the skin effect, $P_{res,R,skin}$ is computed as,

$$P_{res,R,skin} = k_2 |\omega_s|^{n_{R1}} i_s^{n_{R2}} \quad (16)$$

where k_2 , n_{R1} and n_{R2} are non-negative constant parameters and can be fitted similarly and at the same time while fitting (15) i.e. during different loading operational points. The *lsqnonlin* [10] method from MATLAB curve fitting tool is one option, which is also used in this work to fit the data. The total rotor resistive loss $P_{res,R}$ is formulated as,

$$P_{res,R} = 3 R_R i_R^2 + k_2 |\omega_s|^{n_{R1}} i_s^{n_{R2}} \quad (17)$$

IV. STUDIED MACHINES

Two induction machines of nominal power 45 kW referred to as IM₁ and 250 kW referred to as IM₂ respectively are studied in this analysis. The machines’ parameters are provided in Table I.

The mesh and the flux density for the IM₁ and IM₂ are given respectively in Fig. 3 and 4. Fig. 4 shows that the top of the rotor bar is more saturated than the other part of rotor yoke. The stator slot harmonics are the primary reason for it. The skin effect of the rotor bar can be seen in Fig. 5, which has occurred due to the high-frequency slot harmonics. The rotor losses are therefore more on the top of the bar than the bottom part. Both IM₁ and IM₂ are prototyped in the ABB factory in Helsinki, Finland. The measurement results in the factory test [11] and corresponding FE computations are given in Table II. During the factory test, the machine is operated at no-load, full load and different partial load to measure the current, torque, efficiency etc. It is good to mention that during the FE computation, the mechanical losses are imposed as a load torque, based on the retardation test of the machines.

Fig. 3. Mesh of the (a) 45 kW and (b) 250 kW induction machine

Fig. 4. The flux density at a given time instant of the (a) 45 kW and (b) 250 kW induction machine at nominal operational point computed using time-stepping FE solver [1].

Fig. 5. Loss distribution of the (a) 45 kW and (b) 250 kW induction machine at nominal operational point computed using time-harmonic FE solver [1].

V. RESULTS

A. Parameter Identification from Loss Components

For the easiness of reading, only the example of IM₁ will be used to show how to segregate the different loss components and to fit different parameters with those loss

components. To compare the validity of the developed model with FE results, both IM₁ and IM₂ will be used. The parameter identification of IM₁ for no-load operation are performed for 3 different flux levels : 0.37 p.u., 0.74 p.u. and 0.81 p.u. where 1 p.u. flux is computed for nominal voltage U_n and nominal frequency f_n as $U_n / 2\pi f_n = 1.27$ Wb. The three corresponding frequencies are 5 Hz, 25 Hz, and 50 Hz respectively.

Parameter	Value for IM ₁	Value for IM ₂
Power (P_n)	45 kW	250 kW
Voltage (U_n)	400 V	500 V
Current (I_n)	81.94 A	351.76 A
Power factor	0.8476	0.8493
frequency (f_n)	50 Hz	50 Hz
connection	Delta	Delta

For the field-weakening region, the voltage cannot increase with the frequencies higher than 50 Hz. So, the constant voltages are chosen based on the previous flux levels (0.37 p.u., 0.74 p.u. and 0.81 p.u.) at 50 Hz which are 148 V, 296 V and 324 V. The fitting of (13) for the constant flux region and the field-weakening region is shown in Fig. 6 [10].

Fig. 6. Fitting of iron losses of IM₁ from (13) at no-load for the stator in (a) constant flux-region, (b) constant voltage region, and for the rotor in (c) constant flux-region, (d) constant voltage region.

TABLE II
MEASUREMENT VERIFICATION OF THE MACHINES

Parameter	Computed for IM ₁	Measured for IM ₁	Computed for IM ₂	Measured for IM ₂
Current at no-load (A)	31.09	31.2	129.4	129.3
Input power at no-load (kW)	1.1	0.97	3.18	3.84
Torque at full load (Nm)	289.65	289.65	1603.5	1607.0
Rotor speed (rpm)	1483.7	1482.1	1490.6	1488
Current at full load (A)	82.74	83.2	351.7	352.2
Input power at full load (kW)	48.0	48.7	258.8	260.7
Output power at full load (kW)	45.0	45.2	250.3	250.5
Input power at 25% load (kW)	12.85	13.1	67.1	67.5
Output power at 25 % load (kW)	11.25	11.3	62.5	63.3
Input power at 50% load (kW)	24.42	24.8	130.4	131.4
Output power at 50% load (kW)	22.51	22.7	124.9	126.0
Input power at 75% load (kW)	33.76	33.8	194.1	195.8
Output power at 75% load (kW)	36.13	36.7	187.4	188.4

There is not any fixed rule on how to select those flux levels or frequencies, but a good approach is to choose those points as uniformly distributed in the operational range of the flux levels and frequencies. To identify the parameters for (15), three different flux levels (0.74 p.u., 0.81 p.u. and 0.88 p.u.) and three different frequencies (40 Hz, 45 Hz and 50 Hz) are selected to find the difference of the stator iron loss at different loads and no-load. The slip during the loading cases is applied for 25 %, 50 %, 75 % and 100 % of the nominal load. The fitting of $P_{Fe,S,sl}$ is presented in Fig. 7. Similarly, to identify the parameters for (16), three different flux levels (0.74 p.u., 0.81 p.u. and 0.88 p.u.) and three different frequencies (40 Hz, 45 Hz and 50 Hz) are selected to model the skin effect of the rotor bar for different stator currents. The stator current is computed at different loading points, which are 25 %, 50 %, 75 % and 100 % of the nominal load. The fitting of $P_{res,R,skin}$ is presented in Fig. 8. All the different fitting parameters as identified in (13), (15), and (16) are listed for IM₁ and IM₂ in Table III.

TABLE III
IDENTIFIED PARAMETERS FROM FITTED RESULTS

Parameter	Value for IM ₁	Value for IM ₂
k_S	447.42	551.72
n_S	2.11	2.02
R_{FLS}	724.92 Ω	296.47 Ω
k_R	5.22	4.03
n_R	2.3	1.03
R_{FLR}	811.68 Ω	579.00 Ω
k_1	0.0004	0.0014
n_{S1}	1.75	1.65
n_{S2}	1.86	1.84
k_2	0.0081	0.0001
n_{R1}	0.7	0.86
n_{R2}	1.41	1.72
a	0.059	0.025
a	13.27	7.09
β	0.0003	0.047
b	20.96	9.48

Fig. 7. Fitting of stray load losses of IM₁ from (15) at (a) 0.74 p.u., (b) 0.81 p.u., and (c) 0.88 p.u.

Fig. 8. Fitting of rotor bar's skin effect losses of IM₁ from (16) at (a) 0.74 p.u., (b) 0.81 p.u., and (c) 0.88 p.u.

B. Comparison of Proposed Model with FE

The comparison of the proposed model and FE models are done at 1 p.u. flux which has not been used during the fitting. This has been done deliberately to see how the model behaves in other magnetic states, which are not used in the model fitting. The comparison of the different loss components at the nominal operating point is shown in Fig. 9 for IM₁ and IM₂. The losses calculated using the proposed model show close similarities with the FE models. The stator iron loss is slightly underestimated, and the rotor resistance loss is slightly overestimated for both machines in comparison to the FE computed results.

Next, the machine is simulated with rated voltage at different slips to compute the different loadings. The proposed model along with the model as described in [4] are compared for the total electromagnetic losses with the FE

computed losses. The results are presented in Fig. 10, which quantify the improvements of the proposed model than the traditional model in estimating the total loss at different operational states.

Fig. 9. Comparison of the different component of losses at the nominal operation point for (a) IM₁, and (b) IM₂.

Fig. 10. Comparison of the total electromagnetic losses at different operational point for (a) IM₁, and (b) IM₂.

It is worth to mention that the proposed model is meant to be used in a dynamic simulation. However, verifying the model with a dynamic simulation using FE is challenging as it requires a coupling between the control software and finite-element solver. Moreover, measuring losses for cyclic

loading with experiments are also challenging. Hence, in this work, the proposed model is verified at steady state. In future, the verification of the proposed model with dynamic FE simulation and experiment is intended.

VI. CONCLUSION

In this work, a systematic approach to identify the different losses of an induction machine is presented. The proposed model is based on an equivalent circuit model, which is extended for different loss computations. To create the model, 33 different operational states are computed with the FE method to fit the model. The proposed dynamic model can separate the total core loss in constant flux and field-weakening region in rotor and stator components, which is absent in previous researches. Also, the influence of skin effect of the rotor bar and loading of the motor have been considered separately for the stator and rotor iron losses in the proposed model which are ignored in previous works. Once the model is created, it can be used in the real-time system to provide the loss estimation in the different parts of the machine which can be then used as inputs for the thermal model. Therefore, this model is suitable to be used as a component of the digital twin for thermal management of a large-scale industrial system. The model is applied to two different induction machines. The FE computations are validated with the laboratory measurements, and the proposed model is validated with FE for different analysis. In the future, the impacts of the model coefficients and selection of the operational states can be investigated further for improving the accuracy of the model and speeding its identification.

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VIII. REFERENCES

- [1] A. Arkkio, "Analysis of induction motors based on the numerical solution of the magnetic field and circuit equations," Ph.D. dissertation, Lab. Electromech., Helsinki Univ. Technol., Espoo, Finland, Dec. 1987
- [2] J. Pyrhonen, T. Jokinen, and V. Hrabovcova, *Design of rotating electrical machines*, 2nd ed., Chichester, UK: Wiley, 2013, pp. 301.
- [3] D. Zhang, T. Liu, H. Zhao, and T. Wu, "An analytical iron loss calculation model of inverter-fed induction motors considering supply and slot harmonics," *IEEE Trans. Indus. Elec.*, vol. 12, pp. 9194-9204, Dec. 2019.
- [4] M. Ranta, M. Hinkkanen, E. Dlala, A. Repo and J. Luomi, "Inclusion of hysteresis and eddy current losses in dynamic induction machine

- models," in *Proc. 2009 IEEE International Electric Machines and Drives Conference, Miami, FL, 2009*, pp. 1387-139
- [5] R. S. Kanchan and R. R. Moghaddam, "Experimental validation of a novel core-loss model including additional harmonic losses for online energy efficient control of induction motors," in *Proc. 2017 IEEE International Electric Machines and Drives Conference (IEMDC), Miami, FL, 2017*, pp. 1-6.
- [6] E. Dlala, O. Bottauscio, M. Chiampi, M. Zucca, A. Belahcen and A. Arkkio, "Numerical investigation of the effects of loading and slot harmonics on the core losses of induction machines," *IEEE Trans. Mag.*, vol. 48, no. 2, pp. 1063-1066, Feb 2012.
- [7] Yamazaki, A. Suzuki, M. Ohto, T. Takakura and S. Nakagawa, "Equivalent Circuit Modeling of Induction Motors Considering Stray Load Loss and Harmonic Torques Using Finite Element Method," *IEEE Trans. Mag.*, vol. 47, no. 5, pp. 986-989, May 2011.
- [8] T. Tuovinen, M. Hinkkanen, and J. Luomi, "Modeling of Mutual Saturation in Induction Machines," *IEEE Industry Applications Society Annual Meeting*, 2008.
- [9] G. J. Rogers and D. Shirmohammadi, "Induction Machine Modelling for Electromagnetic Transient Program," *IEEE Power Engineering Review*, vol. PER-7, no. 12, pp. 37-38, Dec. 1987.
- [10] "lsqnonlin", MathWorks, n.d. [Online]. Available: <https://se.mathworks.com/help/optim/ug/lsqnonlin.html>. [Accessed: November 17, 2019]
- [11] IEEE Draft Standard Test Procedure for Polyphase Induction Motors and Generators," IEEE P112/D5, pp.1-110, Jan 2015.

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