This is an electronic reprint of the original article.
This reprint may differ from the original in pagination and typographic detail.

Mölsä, Eemeli; Saarakkala, Seppo E.; Hinkkanen, Marko

Influence of magnetic saturation on modeling of an induction motor

Published in:

Proceedings of the 23rd International Conference on Electrical Machines, ICEM 2018

DOI:

[10.1109/ICELMACH.2018.8506746](https://doi.org/10.1109/ICELMACH.2018.8506746)

Published: 03/09/2018

Document Version

Peer reviewed version

Please cite the original version:

Mölsä, E., Saarakkala, S. E., & Hinkkanen, M. (2018). Influence of magnetic saturation on modeling of an induction motor. In *Proceedings of the 23rd International Conference on Electrical Machines, ICEM 2018* IEEE. <https://doi.org/10.1109/ICELMACH.2018.8506746>

Influence of Magnetic Saturation on Modeling of an Induction Motor

Eemeli Mölsä, Seppo E. Saarakkala and Marko Hinkkanen

Abstract—This paper deals with modeling of a saturated induction machine. In control algorithms, the inverse- Γ model is widely used. However, the saturation models are typically based on the T or Γ model. Transformation between these models is easy if the machine is magnetically linear, but since the motor saturates, the modeling becomes more complex. In this paper, we study the properties of the Γ and inverse- Γ models and the transformation between these two models when the saturation is taken into account. The phenomena related to model transformation are studied by means of analytical equations. Also, the functionality of the models is studied by means of laboratory experiments.

Index Terms—Induction motor, magnetic saturation, Γ model, inverse- Γ model.

I. INTRODUCTION

Speed-sensorless, model-based control of an induction motor is widely used in industrial applications. Modeling of an induction motor is discussed, e.g., in [1]–[7], but there are still open issues related to the effect of the saturation characteristics of the motor.

The T model is often considered as a reference model in literature, where the Γ and inverse- Γ models are derived from. Typically, the inverse- Γ model is used in motor control due to the implementation aspects, e.g., an easier implementation of the torque equation. However, the saturation phenomena are easier to include in the Γ model. For example, in [8], the inverse- Γ model, including the saturation, is used in a controller and observer structure. In [9], the saturating Γ model is used. The magnetic saturation phenomenon is present in all electrical machines. In many commercial motors, the main flux (magnetizing branch) is significantly saturated already at the rated operating point. An often used way to model the saturation of the induction motor is to model the stator inductance as a function of the stator flux, e.g. [10]. Furthermore, the low-power induction motors are often equipped with closed rotor slots, making the rotor leakage flux saturate highly as a function of rotor current (i.e., torque) [11]. Thus, both inductances of the Γ and inverse- Γ models are nonlinear. These phenomena will affect also the model transformation making the coupling factor depend nonlinearly on the fluxes and currents. Hence, the selection of the motor model for different purposes is not a self-evident

Fig. 1. Induction motor models: (a) Γ -model with saturating L_M and L_σ ; (b) inverse- Γ -model with saturating L'_M and L'_σ . These nonlinear parameters cause nonlinearity also in R'_R .

issue. In [12], the influence of the motor model selection is studied by means of the finite element method (FEM). However, this paper does not apply any saturation model and the analysis is based only on the FEM-data.

The main contributions of this paper are:

- It is demonstrated that the selection of the motor model structure (Γ model or inverse- Γ model) is not a trivial issue. Even though it is possible to transform information between the models, the magnetic saturation effect will make the models depend nonlinearly on each other.
- The transformation between the Γ and inverse- Γ model structures is analyzed. The starting point for the transformation is a saturated Γ model.
- It is shown that the saturation effect is easier to take into account in the Γ model, making it a preferable choice for the identification and control algorithms.

The properties of the models are compared by means of steady-state experiments using a 2.2-kW induction machine.

II. MOTOR MODEL

The motor is modeled by using complex-valued space vectors, e.g., $u_s = u_{sd} + ju_{sq}$. The parameters of the inverse- Γ model are marked with primes, e.g. L'_M .

This work was supported in part by ABB Oy and in part by the Academy of Finland.

E. Mölsä, S. E. Saarakkala and M. Hinkkanen are with the Department of Electrical Engineering and Automation, Aalto University, Espoo, Finland (e-mail: olli.molsa@aalto.fi; seppo.saarakkala@aalto.fi; marko.hinkkanen@aalto.fi).

TABLE I
MAGNETIC MODEL PARAMETERS FOR 2.2-kW MOTOR.

Parameter	R_s	R_r	L_u	L_∞	c	d	r	s
Value (p.u.)	0.065	0.04	2.56	0.14	1.06	0.025	6	2

A. Voltage and Flux Equations

Fig. 1(a) shows the dynamic Γ model of the induction motor in stator coordinates [1]. The voltage equations of an induction motor are

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

$$\frac{d\psi_R}{dt} = -R_R \mathbf{i}_R + j\omega_m \psi_R \quad (1b)$$

where the stator voltage vector is denoted by $\mathbf{u}_s$, the stator current vector by $\mathbf{i}_s$, stator resistance by R_s , and the electrical angular speed of the rotor by ω_m . The rotor current vector is $\mathbf{i}_R$ and the rotor resistance is R_R . The stator and rotor flux linkages, respectively, are given by

$$\psi_s = L_M(\mathbf{i}_s + \mathbf{i}_R) \quad (2a)$$

$$\psi_R = \psi_s + L_\sigma \mathbf{i}_R \quad (2b)$$

where the stator inductance is L_M and the leakage inductance L_σ . The electromagnetic torque is

$$T = \frac{3p}{2} \text{Im} \{ \mathbf{i}_s \psi_s^* \} \quad (3)$$

where p is the number of pole pairs of the motor.

B. Magnetic Saturation

The stator and leakage inductances depend on the flux linkages (or the currents) due to the magnetic saturation. Because the amplitudes of the flux linkages change (e.g. due to the field weakening or loss-optimizing control), the magnetic-saturation effects should be modeled and taken into account in the control algorithms.

Fig. 1(a) shows the Γ model that is a preferred starting point for modelling a saturated induction machine [1]. The saturation characteristics of the Γ model (unlike those of the T model) can be, in theory, uniquely defined based on measurements from the stator terminals. Furthermore, since the effects of the closed rotor slots are to be modelled, the leakage inductance should be located at the rotor side.

The flux densities in the stator core depend mainly on the stator flux linkage [1]. The stator flux saturation characteristics of a 2.2-kW machine, computed with FEM, are shown in Fig. 2(a). The leakage flux saturation characteristics are shown in Fig. 2(b). Due to the closed rotor slots, the leakage flux has an easy route via the slot bridges at low rotor currents. When the rotor current (and the rotor leakage flux) increases, the thin bridges become fully saturated.

The saturation characteristics could be modelled using measured or computed look-up tables. Alternatively, explicit functions can be used, which typically simplifies the implementation of the model and allows extrapolation beyond the

Fig. 2. Saturation characteristics for a 2.2-kW machine: (a) stator flux magnitude; (b) leakage flux magnitude. The markers show the values from the data computed by FEM. The solid lines correspond to (4) with parameters in Table I.

data range used in the fitting procedure. We apply a saturation model similar to [13], [14]

$$L_M(\psi_s) = \frac{L_u - L_\infty}{1 + (\psi_s/c)^r} + L_\infty \quad (4a)$$

$$L_\sigma(\psi_\sigma) = \frac{L_u - L_\infty}{1 + (\psi_\sigma/d)^s} + L_\infty \quad (4b)$$

where $\psi_s = |\psi_s|$ and $\psi_\sigma = |\psi_\sigma|$ are the flux magnitudes, L_u is the unsaturated inductance, L_∞ is the fully saturated inductance, and c , d , r , and s positive constants. Fig. 2(a) also shows the curves corresponding to (4) parametrized by Table I. Induction machines may show some cross-saturation effects, especially if the rotor slots are skewed. If needed, these effects could be included in the saturation model [14], at the expense of more detailed characterization of the machine and increased number of model parameters.

III. ANALYSIS OF THE MODEL TRANSFORMATION

In this section, we analyze, how the coupling factor is influenced when the operating point of the motor changes. The analysis is based on the equations (1) and (2), in which the saturation is implemented using (4). The equations are parametrized according the Table I. Steady state operation is considered.

The parameters and variables of the Γ model can be transformed to those of the inverse- Γ model using the coupling factor [1]

$$\gamma = \frac{L_M}{L_M + L_\sigma} \quad (5)$$

giving the following parameters for the inverse- Γ model:

$$L'_\sigma = \gamma L_\sigma \quad L'_M = \gamma L_M \quad R'_R = \gamma^2 R_R \quad (6)$$

When this transformation is used, the models are mathematically equivalent in steady state (and in transients as well if the magnetics are linear). It can be seen that due to the magnetic saturation, the equivalent inverse- Γ model parameters depend on the stator flux and the rotor current. The rotor flux can be transformed to the inverse- Γ rotor flux as $\psi'_R = \gamma \psi_R$.

Fig. 3. Transformation of the Γ -model parameters to the inverse- Γ -model when the stator flux is kept constant and torque (and inherently rotor current) changes: (a) fluxes; (b) magnetizing inductance; (c) leakage inductance (d) rotor resistance.

A. Rotor Branch Saturation

As explained, the motor model can be transformed between Γ and inverse- Γ models using the coupling factor (5). The coupling factor, however, includes nonlinear inductances. As a result, the parameters in (6) will become nonlinear functions of the fluxes or currents.

The effect of model transformation is investigated using the Γ model [Fig. 1(a)] as a reference model, in which the magnetic model including the saturation phenomena is modeled using (4).

As an example, the effect of the coupling between the Γ and inverse- Γ model is shown in Fig. 3. The magnitude of the stator flux is kept constant and the motor is at steady state.

Fig. 3(a) shows that the rotor flux linkage drops as a function of torque due to the rotor current. Since the stator flux stays constant, the Γ model main inductance L_M is also constant as can be seen in Fig. 3(b). The leakage inductance in Fig. 3(c) drops rapidly as a function of the torque due to the rotor current. The same behaviour can also be seen in Fig. 2(b). Since the coupling factor (5) depends on the both inductances, the flux and inductance values at the inverse- Γ model differ from the ones in the Γ model in highly nonlinear way. Without the saturation phenomenon in the rotor branch, the coupling factor, and inherently the transformed parameters, would be linear in the case of constant stator flux.

The rotor resistance R'_R also depends on the torque [cf. Fig. 3(d)]. The rotor resistance increases approximately 30% from its no-load value when the torque is increased to 1.5 p.u. It can be noticed that the model transformation will make the modeling and identification of these nonlinear phenomena more complex.

B. Magnetizing Branch Saturation

The main (magnetizing) inductance depends on the stator flux amplitude according to (4a). Since the coupling factor (5) includes this term, the nonlinearity originates also from there. Fig. 4 presents an example of situation in which the torque is kept constant and the stator flux changes. Fig. 4(a) shows that the magnetizing inductance depends highly on the stator flux linkage. In Fig. 4(b), we can see that the coupling of the leakage inductance behaves nonlinearly as well.

IV. FITTING THE SATURATING MOTOR DATA IN THE MODELS

The dynamic Γ model is presented in Fig. 1(a). The stator inductance of this model is directly the magnetizing inductance L_M and its nonlinearity can be explicitly defined as a function of the amplitude of the stator flux as demonstrated in (4a) and the numerical values are easily obtained by means of measurements from the motor terminals, e.g., by applying a no-load test.

Fig. 4. Transformation of the Γ -model parameters to the inverse- Γ -model when the torque is kept constant and the d-component of the stator flux changes: (a) magnetizing inductance; (b) leakage inductance.

Furthermore, as explained in Section II-B, the closed rotor slot motors have saturating leakage inductance [cf. (4b)]. Thus, the physical nature of the motor makes the Γ model a more natural choice for the modeling purposes. However, the inverse- Γ model [Fig. 1(b)] has advantages from the control algorithm implementation point of view. For example, in case of the rotor flux oriented control, the torque becomes directly proportional to the magnitude of the rotor flux linkage and the quadrature component of the stator current as follows:

$$T = \frac{3p}{2} \psi'_R i_{sq} \quad (7)$$

In [1], the model transformation using the coupling factors is introduced. However, it is possible to transform the saturating magnetizing inductance using several approaches. Two different approaches are introduced to transform the information of a saturating stator inductance L_M of the Γ model to the corresponding saturating magnetizing inductance L'_M of the inverse- Γ model.

A. Saturation Model Transformation: Case A

The Γ model magnetizing inductance L_M , which depends on the amplitude of the stator flux linkage [cf. (4a)], is transformed to inverse- Γ model. The transformation results in having the magnetizing inductance L'_M as a function of the inverse- Γ model rotor flux linkage ψ'_R . The transformation can be done through the following steps:

Fig. 5. The transformation of the saturating magnetizing inductance from the Γ model to inverse- Γ model. (a) shows the stator flux linkage ψ_s of the Γ model, which is directly the flux over the magnetizing inductance L_M . (b) shows the stator flux ψ_s , which includes also the flux of the leakage inductance L'_σ . The rotor flux ψ'_R of the inverse- Γ model is the flux over the magnetizing inductance L'_M . In standstill, there is no rotor current and $i_M = i'_M = i_s$.

- 1) Solve the stator flux as a function of the magnetizing current: $\psi_s = \psi_s(i_s)$, cf. Fig. 5(a). In a no-load condition, there is no rotor current and the stator current goes to the magnetizing branch ($i_M = i'_M = i_s$). On other hand, the stator current can be expressed as $i_s = i_s(\psi_s)$. An example of the stator flux is shown in Fig. 2(a). Alternatively, the stator flux values can be obtained experimentally from a no-load test:

$$\psi_s = \frac{u_s - R_s i_s}{\omega_s} \quad (8)$$

where ω_s is the stator angular frequency. Using different stator voltage magnitudes, the stator flux values can be obtained as a function of stator current.

- 2) Next, in Fig. 5(b), calculate the rotor flux ψ'_R as:

$$\psi'_R(i_s) = \psi_s(i_s) - L'_\sigma i_s \quad (9)$$

Because the saturation characteristics of the leakage inductance L_σ are relatively complex to model, select a constant L'_σ .¹ Alternatively, selecting L'_σ relatively close to the fully saturated inductance L_∞ should lead to reasonable results, because according to Fig. 2(b), it represents the behaviour of ψ_σ relatively well.

- 3) Finally, calculate the inverse- Γ model magnetizing inductance as

$$L'_M(\psi'_R) = \frac{\psi'_R}{i_s(\psi_s)} \quad (10)$$

¹The value of the constant L'_σ can be experimentally obtained, e.g., by completing a pulse test or a high frequency sinusoidal excitation test with a stand-still rotor. The rotor deep-bar effect and the leakage flux saturation effect (e.g. due to closed rotor slots) is not considered. These assumptions are typical implementing the motor control algorithms [15].

Fig. 6. Transformation of the saturating magnetizing flux and inductance from Γ to inverse- Γ model. (a) shows the stator flux linkage ψ_s as a function of the Γ model magnetizing current i_M and the inverse- Γ model leakage flux ψ'_σ as a function of the stator current i_s . The inverse- Γ rotor flux linkage ψ'_R is presented as a function of inverse- Γ magnetizing current i'_M . (b) shows the Γ model magnetizing inductance L_M (stator inductance) as a function of stator flux linkage and the transformed inverse- Γ model magnetizing inductance L'_M as a function of rotor flux linkage. The L'_M does not correspond the real motor at higher flux amplitude values (dashed line).

Fig. 6 shows an example of the transformation. The stator flux is calculated as a function of magnetizing current using (2a) and (4a) parametrized by Table I, as shown in Fig. 2. Due to the assumption of the constant leakage inductance ($L_\sigma = L_\infty = 0.14$ p.u.), the inverse- Γ model does not correspond the real motor at higher flux linkage values. However, the model is valid at the normal operating range. In the Fig. 6, the scale is extended from the normal operating range to show the behaviour of the model at higher flux amplitudes (dashed line).

B. Saturation Model Transformation: Case B

Another approach to the transformation is to directly subtract the leakage inductance from the stator inductance:

$$L'_M(\psi_s) = L_M(\psi_s) - L_\sigma \quad (11)$$

Fig. 7. Block diagram of the experimental system.

Fig. 8. The stator current estimation using Γ and inverse- Γ models as in Fig. 1. The parameters of the Γ model are fitted to correspond the data measured from the 2.2-kW motor and the parameters of the inverse- Γ model are calculated using the method described in Section IV-A.

This approach is simpler than the Case A, but the model does not correspond the physical motor, since the inverse- Γ magnetizing inductance L'_M cannot depend on the stator flux amplitude ψ_s . This simplistic transformation, however, may be sufficient for standard motor control algorithms.

V. EXPERIMENTAL RESULTS

The effect of nonlinearity in the motor models is studied by means of experimental tests with a 2.2-kW cage induction motor. The test motor is supplied by an adjustable three-phase sinusoidal voltage supply and the load torque is produced by an external load motor connected to the shaft of the test motor. The dSPACE DS1103 processor board was used for data acquisition and sending the torque reference to the load drive. The configuration of the experimental setup is depicted in Fig. 7.

An example case, where the motor operates at steady state having sinusoidal voltage supply, is considered. In order to reduce the effect of iron losses, the supply frequency was chosen to be half of the nominal (25 Hz) and the supply voltage is set to 180 V, corresponding the electromotive force of the motor at nominal stator flux. The phase voltages, phase currents, mechanical speed and torque are measured.

Fig. 8 shows an example, where the Γ and inverse- Γ models are applied as current observers. The observers are based on the voltage equations (1) at steady state, and the flux equations (2) including the saturating magnetizing inductance. The saturating Γ -model magnetizing inductance

$L_M(\psi_s)$ is fitted to the FEM data, as shown in Fig. 2. The leakage inductance and the rotor resistance are set to constant values, $L_\sigma = L_\infty = 0.14$ p.u. (highly saturated value) and $R_R = 0.04$ p.u.

The inverse- Γ model is parametrized using the Γ model as a starting point. The saturating magnetizing inductance $L'_M(\psi'_R)$ is transformed using the guidelines given in Section IV-A. The parameters L'_σ and R'_R are transformed from the Γ model using the coupling factor (5). The coupling factor is calculated by using the unsaturated magnetizing inductance value and highly saturated leakage inductance value ($L_M = 2.42$ p.u. and $L_\sigma = L_\infty$). The stator resistance value $R_s = 0.065$ p.u. is used in both models.

Fig. 9. The measured and estimated stator current magnitudes. $U = 180$ V, $f = 25$ Hz. (a) shows the measured current magnitude and the estimated magnitudes as in Fig. 8. (b) shows the estimation error $i_s - \hat{i}_s$. The coupling factor for inverse- Γ model parameters is set to correspond the unsaturated magnetizing inductance ($L_M = L_u - L_\infty = 2.42$ p.u.) and highly saturated leakage inductance ($L_\sigma = L_\infty$).

In Fig. 9, we compare the current magnitudes as a function of the slip (i.e., torque), estimated by the Γ and inverse- Γ models, to the measured current magnitude. It can be seen that both the Γ and inverse- Γ models correspond to the measured data relatively fine. According to Fig 9(b), the current estimation error stays reasonably low even with the increasing slip. However, it is important to notice that the selection of the coupling factor influences on the functionality of the inverse- Γ model.

To demonstrate the importance of the proper coupling factor selection, the coupling factor (5) is calculated by

changing the magnetizing inductance to correspond the nominal stator flux (i.e., $L_M = 1.5$ p.u.) and having the other parameters unchanged. Fig. 10 shows the resulting current estimation from the Γ and inverse- Γ models. The current estimation error of the inverse- Γ model increases slightly when going to higher currents, being closest to the Γ model at relatively small currents. If the magnetizing inductance is selected to correspond a higher saturation state, then the current estimation of the inverse- Γ model deviates more from the measured values. It is worth noticing that if the coupling factor separately for corresponding operating point, the Γ and inverse- Γ models become equal.

Fig. 10. The measured and estimated stator current magnitudes. $U = 180$ V, $f = 25$ Hz. (a) shows the measured current magnitude and the estimated magnitudes as in Fig. 8. (b) shows the estimation error $i_s - \hat{i}_s$. The coupling factor for inverse- Γ model parameters is set to correspond the nominal magnetizing inductance ($L_M = 1.5$ p.u.) and highly saturated leakage inductance ($L_\sigma = L_\infty$). The inverse- Γ current estimation is slightly influenced especially at higher current values.

VI. CONCLUSIONS

In this paper, the modeling aspects of the induction motor for control and identification purposes are analyzed. Typically, the stator flux linkage is first defined as a function of the magnetizing current by means of no-load test. Then, the resulting data is fitted to either Γ or inverse- Γ model for the implementation of the motor control algorithms.

It is shown that the Γ model has many advantages compared to inverse- Γ model. Firstly, the magnetizing inductance can be directly expressed as a function of the stator flux, and

hence, defined by means of straightforward measurements from the motor terminals. Furthermore, the leakage inductance is at the rotor side, which enables modeling of the rotor leakage inductance saturation as a function of rotor current. These saturation models can be implemented using analytical functions.

The inverse- Γ model is, however, easier to implement in the control algorithms, since the torque and slip equations become simpler. Also, the measurement of the leakage inductance using a pulse test is easier. To include the nonlinearities modeled with the Γ model, the attention has to be paid on the saturation model implementation and the coupling factors have to be calculated at all operating points. Thus, it is demonstrated that the saturation effect is easier to take into account in the Γ model, making it a more preferable choice.

REFERENCES

- [1] G. R. Slemon, "Modelling of induction machines for electric drives," *IEEE Trans. Ind. Appl.*, vol. 25, no. 6, pp. 1126–1131, Nov./Dec. 1989.
- [2] S. D. Sudhoff, D. C. Aliprantis, B. T. Kuhn, and P. L. Chapman, "An induction machine model for predicting inverter-machine interaction," *IEEE Trans. Energy Convers.*, vol. 17, no. 2, pp. 203–210, June 2002.
- [3] C. Gerada, K. J. Bradley, M. Sumner, and P. Sewell, "Evaluation and modelling of cross saturation due to leakage flux in vector controlled induction machines," vol. 43, no. 3, pp. 694–702, May/June 2007.
- [4] S.-I. Moon, A. Keyhani, and S. Pillutla, "Nonlinear neural-network modeling of an induction machine," *IEEE Transactions on Control Systems Technology*, vol. 7, no. 2, pp. 203–211, Mar 1999.
- [5] M. L. Aime, M. W. Degner, and R. D. Lorenz, "Saturation measurements in AC machines using carrier signal injection," in *Conf. Rec. IEEE-IAS Annu. Meeting*, vol. 1, St. Louis, MO, Oct. 1998, pp. 159–166.
- [6] J. Pedra, I. Candela, and A. Barrera, "Saturation model for squirrel-cage induction motors," *Electric Power Systems Research*, vol. 79, no. 7, pp. 1054 – 1061, 2009. [Online]. Available: <http://www.sciencedirect.com/science/article/pii/S0378779609000091>
- [7] Y. Xu and D. C. Aliprantis, "Wound-rotor induction machine model with saturation and high-frequency effects," in *2014 International Conference on Electrical Machines (ICEM)*, Sept 2014, pp. 1838–1844.
- [8] M. Hinkkanen, L. Harnefors, and J. Luomi, "Reduced-order flux observers with stator-resistance adaptation for speed-sensorless induction motor drives," *IEEE Trans. Power Electron.*, vol. 25, no. 5, pp. 1173–1183, May 2010.
- [9] Z. Qu, M. Ranta, M. Hinkkanen, and J. Luomi, "Loss-minimizing flux level control of induction motor drives," *IEEE Transactions on Industry Applications*, vol. 48, no. 3, pp. 952–961, May 2012.
- [10] M. Ranta, M. Hinkkanen, E. Džala, A.-K. Repo, and J. Luomi, "Inclusion of hysteresis and eddy current losses in dynamic induction machine models," in *Proc. IEEE IEMDC'09*, Miami, FL, May 2009, pp. 1387–1392.
- [11] A. Boglietti, R. I. Bojoi, A. Cavagnino, P. Guglielmi, and A. Miotto, "Analysis and modeling of rotor slot enclosure effects in high-speed induction motors," *IEEE Transactions on Industry Applications*, vol. 48, no. 4, pp. 1279–1287, July 2012.
- [12] M. Hinkkanen, A.-K. Repo, and J. Luomi, "Influence of magnetic saturation on induction motor model selection," in *Proc. ICEM'06*, Chania, Greece, Sep. 2006, CD-ROM.
- [13] N. R. Klaes, "Parameter identification of an induction machine with regard to dependencies on saturation," *IEEE Trans. Ind. Appl.*, vol. 29, no. 6, pp. 1135–1140, Nov./Dec. 1993.
- [14] T. Tuovinen, M. Hinkkanen, and J. Luomi, "Modeling of saturation due to main and leakage flux interaction in induction machines," *IEEE Trans. Ind. Appl.*, vol. 46, no. 3, pp. 937–945, May/June 2010.
- [15] R. J. Kerkman, J. D. Thunes, T. M. Rowan, and D. W. Schlegel, "A frequency-based determination of transient inductance and rotor resistance for field commissioning purposes," *IEEE Trans. Ind. Appl.*, vol. 32, no. 3, pp. 577–584, May/June 1996.

BIOGRAPHIES

Eemeli Mölsä received the B.Eng. degree in electrical engineering from Metropolia University of Applied Sciences, Helsinki, Finland in 2012 and M.Sc.(Tech.) degree in electrical engineering from Aalto University, Espoo, Finland in 2015. He is currently working toward the D.Sc.(Tech.) degree at Aalto University. His main research interest is the control of electrical drives.

Seppo E. Saarakkala received the M.Sc.(Eng.) degree in electrical engineering from the Lappeenranta University of Technology, Lappeenranta, Finland, in 2008, and the D.Sc. (Tech.) degree in electrical engineering from Aalto University, Espoo, Finland, in 2014. He has been with the School of Electrical Engineering, Aalto University, since 2010, where he is currently a Postdoctoral Research Scientist. His main research interests include control systems and electric drives.

Marko Hinkkanen (M'06–SM'13) received the M.Sc.(Eng.) and D.Sc.(Tech.) degrees in electrical engineering from the Helsinki University of Technology, Espoo, Finland, in 2000 and 2004, respectively. He is an Associate Professor with the School of Electrical Engineering, Aalto University, Espoo. His research interests include control systems, electric drives, and power converters. Dr. Hinkkanen was the corecipient of the 2016 International Conference on Electrical Machines (ICEM) Brian J. Chalmers Best Paper Award and the 2016 IEEE Industry Applications Society Industrial Drives Committee Best Paper Award. He is an Editorial Board Member of IET Electric Power Applications.