Abstract

This paper presents a novel method for the rotor resistance identification of an indirect rotor field oriented (IRFO) control system of a saturated induction motor. A saturated induction motor is an induction motor with a variable (saturable) mutual inductance. Saturation of the mutual inductance appears in operating conditions with variable rotor flux levels in the IRFO control system of an induction motor. The rotor resistance identification is based on an inverse rotor time constant identification, which is based on the model reference adaptive system (MRAS) principle. The magnitude of the rotor flux has been estimated in the stationary reference frame by the voltage model (reference model) and by the current model (adaptive model). The error signal of the rotor flux magnitude of the two estimators is applied to drive a PI mechanism that provides a correction to the inverse rotor time constant. This paper presents an original analysis of the IRFO system stability by taking small deviations of mutual inductance and rotor resistance at a steady-state operating point. The effectiveness of the proposed method for rotor resistance identification has been verified by digital simulation and experimentation.

Key Words

Induction motor, MRAS, rotor field control, saturation

1. Introduction

Induction motors traditionally have been used in open-loop drives for various reasons, including cost, size, reliability, simplicity, efficiency and ease of manufacturing. With developments in digital signal processors and power electronics, induction motors can now be used in high-performance variable-speed drives. Some of these drives are based on the rotor field oriented (RFO) method [1]. There are two types of RFO method: direct RFO and indirect RFO (IRFO) control systems. The direct RFO control system measures or estimates the rotor flux position. Measuring the flux has the disadvantage that coils must be inserted into the air gap of the motor. Estimating the flux from the stator voltages and currents results in poor performance over a low-speed region. To make the RFO control systems generally applicable, the IRFO control system is used. Here, the rotor flux space vector is estimated using the induction motor model. But, the estimation of the rotor flux space vector depends on the mutual inductance and rotor resistance. The mutual inductance varies with magnetic saturation and the rotor resistance is a thermally dependent parameter. Therefore, a method for mutual inductance identification is developed, and their result is utilized for rotor resistance identification in real time. The mutual inductance identification is based on the iterative method described in Section 3. Rotor resistance identification at a constant magnetizing level of induction motor is described in [2]. In this paper, the method is improved from the viewpoint of the variable magnetizing level of induction motor. This is important from the viewpoint of the effectiveness of the induction motor [3] because very often, induction motor drives work below the nominal torque most of the time. Also, an induction motor operating in the field-weakening region has variable magnetizing levels. The effectiveness of the proposed method for rotor resistance identification has been verified by digital simulation and experimentation.

In this paper, an original analysis of the IRFO system stability by taking small deviations of mutual inductance and rotor resistance at a steady-state operating point is presented. By plotting trajectories of the poles and zeros of the IRFO system transfer function, it is possible to draw conclusions on system stability.

2. Rotor Field Oriented Control of Induction Motor

Fig. 1 shows the IRFO control system of induction motor with mutual inductance and rotor resistance identification.

An induction motor can be described by the following equations in a synchronously rotating (d,q) reference frame [1]:

\[ u_s = i_s R_s + \frac{d\psi_s}{dt} + j\omega_s \psi_s \]  

(1)
Taking into account (5), (2) becomes

\[ 0 = \frac{1}{T_r} \frac{d\psi_r}{dt} + \frac{L_m}{L_r} i_s + s\psi_r + j(\omega_e - \omega_r)\psi_r \]  (6)

where \( T_r \) is the rotor time constant, and \( s \) is the Laplace operator \( \left( \frac{d}{dt} \right) \).

The substitution of (5) into (4), and then (4) into (1) yields

\[ u_s = (R_s + (s + j\omega_e)\sigma L_s)i_s + (s + j\omega_e)\frac{L_m}{L_r} \psi_r \]  (7)

Equation (6) is well known as the current model and (7) as the voltage model of an induction motor with rotor speed, stator currents and rotor fluxes as the state variable of the system.

Rewriting (7) to give the stator voltage in the \( d,q \) reference frame yields

\[ u_{sd} = R_s i_{sd} + s\sigma L_s i_{sd} - \omega_e \sigma L_s i_{sq} + s\frac{L_m}{L_r} \psi_{rd} - \omega_e \frac{L_m}{L_r} \psi_{rq} \]  (8)

\[ u_{sq} = R_s i_{sq} + s\sigma L_s i_{sq} + \omega_e \sigma L_s i_{sd} + s\frac{L_m}{L_r} \psi_{rq} + \omega_e \frac{L_m}{L_r} \psi_{rd} \]  (9)

Considering the rotor flux orientation conditions, \( \psi_{rq} = 0 \) and steady-state conditions, (8) and (9) can be simplified to

\[ R_s i_{sd} + s\sigma L_s i_{sd} = u_{sd} + \omega_e \sigma L_s i_{sq} \]  (10)

\[ R_s i_{sq} + s\sigma L_s i_{sq} = u_{sq} - \omega_e \left( \sigma L_s i_{sd} + \frac{L_m}{L_r} \psi_r \right) \]  (11)

It can be seen that the \( d \)- and \( q \)-axes voltage equations are coupled by the terms \( \omega_e \sigma L_s i_{sq} \) and \( -\omega_e \sigma L_s i_{sd} \). These terms can be eliminated by the use of a decoupler in the control part, requiring accurate knowledge of \( \sigma L_s \). The term \( \sigma L_s \) mainly depends on knowledge of the mutual inductance because stator and rotor leakage inductances are constant in normal operating conditions of an IRFO control system. Therefore, mutual inductance identification in real time is one of the goals of this paper.
The two PI current controllers (Fig. 1) act to produce the decoupled voltages $u_{sd1}$ and $u_{sq1}$

$$R_s i_{sd}^* + s\sigma L_s i_{sd}^* = u_{sd1}$$  \hspace{1cm} (12)

$$R_s i_{sq}^* + s\sigma L_s i_{sq}^* = u_{sq1}$$  \hspace{1cm} (13)

With regards to (10), (11), (12) and (13), the correct reference voltages are

$$u_{sd1}^* = u_{sd1} - \omega_e \hat{\sigma} \hat{L}_s i_{sq}^*$$  \hspace{1cm} (14)

$$u_{sq1}^* = u_{sq1} + \omega_e \left( \hat{\sigma} \hat{L}_s i_{sq}^* + \frac{\hat{L}_m}{L_r} \psi_r^* \right)$$  \hspace{1cm} (15)

Reference voltages $u_{sd1}^*$ and $u_{sq1}^*$ ensure decoupled two-axes control of the induction motor.

For the IRFO control system, the rotor flux linkage vector has only the real component, which is assumed to be constant in the steady state (from (6)), i.e.,

$$\psi_{rd} = \psi_r^* = \hat{L}_m i_{sd}$$  \hspace{1cm} (16)

and the slip frequency is

$$\omega_s = \omega_e - \omega_r = \frac{i_{sq}}{T_{rsd}}$$  \hspace{1cm} (17)

Fast power electronic transistors ensure an ideal current control, i.e., the actual currents $i_{sd}$ and $i_{sq}$ in (16) and (17) are replaced by the references $i_{sd}^*$ and $i_{sq}^*$.

3. Identification of Magnetic Saturation

Inductances of induction motor are saturable parameters. Mutual inductance saturation is a consequence of variable magnetizing levels of induction motor. This is very important from the viewpoint of all types of field oriented control systems. In the past, many methods for mutual inductance identification are developed. Their brief review is referenced in [4, 5]. Paper [6] has been the first publication to incorporate the effects of main flux saturation in the space-vector equations of induction machines.

Saturation impact of leakage inductances on induction motor characteristics appears in conditions with increased motor currents (e.g., as the induction motor starts in mains-supplied induction motor drives), an effect not found in induction motor field oriented control systems. Hence, the identification of magnetic saturation in this paper implies mutual inductance saturation only. As a consequence of magnetizing circuit saturation, the mutual inductance becomes a current dependent parameter, i.e., $L_m(i_m)$ [7].

The calculation of the mutual inductance saturation is based on measured currents from the traditional locked rotor and no-load tests, and an output value is a saturation factor of the magnetizing circuit. In the steady-state conditions mentioned earlier, the calculation is performed by iterative changes of the state variables (the linkage stator and rotor fluxes) in the induction motor equations for steady state. The input data are an unsaturated mutual inductance ($L_m^i$) and corresponding curve of the saturation factor ($k_{zm}$). The originality of this paper is in the improvement of the method referenced in [8] by iterative changes of state variables (stator and rotor linkage fluxes). In [8], the authors compute saturation factors using an analog computer.

Appendix A shows the parameters of the squirrel cage induction motor used in this work.

The saturated mutual inductance is [7,8]

$$L_m^z = L_m^i (1 - k_{zm})$$  \hspace{1cm} (18)

The mutual inductance $L_m^z$ as a function of the no-load $I_0$ current is shown in Fig. 2. This result has been obtained by measuring the no-load and locked rotor currents. In this paper, the saturation factor $k_{zm}$ is expressed as $k_{zm}(i_m)$. This factor is approximated by analytical expressions given in Appendix B.

![Figure 2](image-url)
The linkage space vectors of stator and rotor (under steady-state conditions) in the synchronously rotating \((d,q)\) reference frame (from (1) and (2)) are

\[
\psi_s = -\frac{j}{\omega_e} (u_s - i_s R_s)
\]
\[
\psi_r = \frac{j}{\omega_s} i_r R_r
\]

The stator and rotor linkage space vectors expressed as a sum of the leakage flux and the magnetizing flux are

\[
\psi_s = \psi_{sl} + \psi^z_m
\]
\[
\psi_r = \psi_{rl} + \psi^z_m
\]

where

\[
\psi_{sl} = L_{sl} i_s
\]
\[
\psi_{rl} = L_{rl} i_r
\]

The space vector of the main magnetizing flux is the following when only the saturated value exists:

\[
\psi^z_m = L^z_m (i_s + i_r)
\]

All space vectors of the induction motor are expressed in the synchronously rotating \((d,q)\) reference frame, where the space vector of stator voltages \(u_s\) is aligned to the \(d\)-axis.

Substituting (21) into (20), the space vectors of stator \(i_s\) and rotor \(i_r\) currents become

\[
i_s = \frac{1}{L_{sl}} (\psi_s - \psi^z_m)
\]
\[
i_r = \frac{1}{L_{rl}} (\psi_r - \psi^z_m)
\]

When (23) is substituted into (19) we obtain

\[
\psi_s = \frac{1 - j \omega_e T_s}{1 + \omega_e^2 T_s^2} (T_s u_s + \psi^z_m)
\]
\[
\psi_r = \frac{1 - j \omega_e T_r}{1 + \omega_e^2 T_r^2} \psi^z_m
\]

where \(T_s = L_{sl}/R_s, T_r = L_{rl}/R_r\).

From (6) and (7), in steady state, we derive the space vector of the stator voltages, which ensures constant rotor flux:

\[
u_s = \psi^* \omega_e \frac{\sigma z L_z L^z_m}{R_s L^z_m} \times \left( \frac{R_r}{\sigma^2 L^z_e} + \frac{\omega_s R_s}{\omega_e \sigma^2 L^z_e} \right)^2 + \left( \frac{\psi_s - \frac{R_s R_r}{\omega_e \sigma^2 L^z_e}}{\omega_e \sigma^2 L^z_e} \right)^2 + j0
\]

where \(\psi^*\) is a set reference rotor flux command.

Here, superscripts \(z\) and \(n\) denote the saturated and unsaturated portion of the corresponding parameter or variable, respectively.

The saturated magnetizing flux space vector \(\psi^z_m\) in (24) is

\[
\psi^z_m = \psi^z_m (1 - k_z m) = \psi^n_m - \Delta \psi_m
\]

The unsaturated space vector \(\psi^n_m\) in (26) is

\[
\psi^n_m = L^n_m (i_s + i_r)
\]

According to (26), this can be written as

\[
\Delta \psi_m = k_z m \psi^n_m
\]

The substitution of (26) into (23) and then (23) into (27) gives

\[
\psi^n_m = \frac{L^*_{m}}{L_{sl}} (\psi_s + \Delta \psi_m) + \frac{L^*_{m}}{L_{rl}} (\psi_r + \Delta \psi_m)
\]

where \(L^*_{m} = L^n_m L_{sl} L_{rl} /(L^n_m (L_{sl} + L_{rl}) + L_{sl} L_{rl})\).

Equations (19)–(21) present a mathematical model for the calculation of the saturated mutual inductance and corresponding state variables.

Determination of the saturation factor is based on the so-called iterative method assuming a sufficiently low difference \(\varepsilon\) (usually less than \(10^{-3}\)) of the stator and rotor currents between the \((k + 1)\)th and \(k\)th sampling time.

The computation of the induction motor steady-state characteristics with magnetic saturation is described by the flowchart shown in Fig. 3. The initial rotor speed \(\omega_{\text{start}}\) and the final rotor speed \(\omega_{\text{end}}\) determine the first and the last rotor speed through the developed method.

The iterative method ensures accurate and fast convergence for stator frequencies up to 1 Hz. Therefore, the number of iterations is fewer than 1,000 for each set of the input parameters. Sensitivity of the mutual inductance identification with respect to the stator and rotor resistance variations is negligible. This fact is consistent with the mutual inductance estimation described in [4].

Using this iterative method, it is shown that mutual inductance identification mainly depends on the rotor flux level \(\psi^*\) over a wide range of rotor speeds and load torques. The rotor speed range is taken to be 0–157 rad/s, and the range of the load torque is 0–25 Nm. Obtained mean values of the mutual inductance and their standard deviations for different rotor flux levels over the above-mentioned load torque and rotor speed range are shown in Table 1.

Based on results shown in Table 1, the mean value of the mutual inductance is almost constant for fixed rotor flux levels. Variations of the mean value are described by the standard deviation \(\sigma\) which is negligible with respect to the mean value of the mutual inductance.

The highest difference between this mean value and the actual mutual inductance is less than 3% over the complete range of the load and rotor speed changes and over the complete rotor flux range. This analysis justifies setting
the constant mutual inductance in RFO control system at a constant rotor flux level. Based on the above-mentioned analysis, we conclude that the mutual inductance depends on the rotor flux level only.

Dependence of the mutual inductance on the rotor flux level is utilized using a look-up table for online mutual inductance updates in the IRFO control system.

4. Identification of the Inverse Rotor Time Constant

There are many methods for estimation of the (inverse) rotor time constant [5]. The one group of online rotor time constant adaptation methods is based upon principles of model reference adaptive system (MRAS). This is the approach with relatively simple implementation requirements. The basic idea is that one quantity (vector or scalar) can be calculated in two different ways. One of the two obtained quantities is independent of the rotor time constant. The difference between the two quantities is an error signal, whose existence is assigned to the error in the rotor time constant. The error signal is applied to drive an adaptive mechanism (PI or I), which provides a correction to the rotor time constant.

There are three common features that all MRAS methods of this group share [5]. First, rotor resistance adaptation is usually satisfactory in steady states only and it is disabled during transients. Second, stator voltages are required for calculation of the adaptive quantity and they have to be either measured or reconstructed from the inverter driving signals and measured DC link voltage. Third, in most cases, identification does not work at zero speed and at zero load torque.

This paper utilizes the identification of the inverse rotor time constant described in [2]. In that paper, the identification is applied in real time with the sample time equal to the sample time of the control system. As mentioned earlier, an identification of the rotor time constant is satisfied in steady states only. Hence, there is no need for online identification, which can even deteriorate control characteristics during transients. Because of this, the method is applied where the time constant of the PI mechanism is set to 100 ms. Additionally, the method is upgraded with the mutual inductance identification. Taking into account unchangeable stator and rotor leakage inductances, rotor resistance identification is ensured.

Equations (6) and (7) described in the stationary (α, β) reference frame become:

\[
0 = \frac{1}{T_r} \dot{\psi}_r - \frac{\dot{L}_m}{T_r} i_s + s \dot{\psi}_r - j\omega_r \psi_r \tag{30}
\]

\[
u_s = \frac{R_s + s\sigma \dot{L}_s}{T_r} i_s + \frac{L_m}{T_r} \psi_r \tag{31}
\]

Equation (30) gives an estimation for the rotor flux space vector upon easily measured stator currents and rotor speed. This estimation depends mainly on the accuracy of the mutual inductance and the inverse rotor time constant identification. Note that mutual inductance identification has been carried out upon the method described in sequence 3. Equation (30) presents a so-called adaptive model for the rotor flux estimation. However, (31) gives an estimation for the rotor flux space vector based

<table>
<thead>
<tr>
<th>(\psi_r^*/\psi_{rn})</th>
<th>0.7</th>
<th>0.8</th>
<th>0.9</th>
<th>1.0</th>
<th>1.1</th>
<th>1.2</th>
<th>1.3</th>
</tr>
</thead>
<tbody>
<tr>
<td>(L_m) (H)</td>
<td>0.4058</td>
<td>0.4055</td>
<td>0.3765</td>
<td>0.3269</td>
<td>0.2588</td>
<td>0.2206</td>
<td>0.1920</td>
</tr>
<tr>
<td>(\sigma) (H)</td>
<td>0</td>
<td>6.3 \cdot 10^{-4}</td>
<td>0.0030</td>
<td>0.0074</td>
<td>0.0028</td>
<td>0.0019</td>
<td>0.0011</td>
</tr>
</tbody>
</table>

Table 1

Mean Value of Mutual Inductance and their Standard Deviations at Fixed Rotor Flux Levels
upon measured stator currents and reconstructed voltage space vector from the measured DC link voltage and the inverter driving signals. Equation (31) is independent of the inverse rotor time constant and, accordingly, can be used as the reference model for the rotor flux space vector. The error signal of the rotor flux magnitude of the two estimators is applied to drive an adaptive mechanism (PI), which provides a correction to the inverse rotor time constant. Taking into account constant leakage rotor inductance and mutual inductance identification (Section 3), the rotor resistance can be identified by the equation

$$\hat{R}_r = \frac{1}{T_r} (\hat{L}_m + L_{rl}) \quad (32)$$

The overall identification procedure of the rotor resistance is shown in Fig. 4.

![Figure 4. Rotor resistance identification.](image)

In our case, the rotor resistance identification is valid beyond a rotor speed that is approximately 10% of the rated speed and beyond a load torque that is 20% of the rated torque.

### 5. Stability of the IRFO Control System

In this section the analysis of the proposed system shown in Fig. 1 is presented. By taking a small perturbation at a steady-state operating point, a linear model of the proposed system can be deduced [9, 10]. To examine the effect of stator resistance variation, the stator resistance identifier is not included in the following linear model. For simplifying the stability analysis, the following assumptions are considered.

1. The motor parameters are constant, except for the mutual inductance and rotor resistances.
2. The motor actual voltages are equal to the voltages’ references, i.e.,

   $\begin{align*}
   u_{sd} &= u^*_{sd} \\
   u_{sq} &= u^*_{sq}
   \end{align*}$

   (33)

To produce a transfer function $G(s) = \frac{\Delta \omega_r}{\Delta \omega^*_r}$, all of the differential equations are linearized using the small-signal analysis technique.

In the following analysis, the corresponding matrices are given in Appendix C.

Using (1)–(5), the linear model of the motor can be written as follows:

$$s \Delta x = A \Delta x + B \Delta \omega^*_r + B_T \Delta t_l \quad (34)$$

where $\Delta x = [\Delta i_{sd} \; \Delta i_{sq} \; \Delta \psi_{rd} \; \Delta \psi_{rq} \; \Delta \omega_r]^T$, $\Delta u_s = [\Delta u_{sd} \; \Delta u_{sq} \; \Delta \omega^*_r]^T$ and $t_l$ is the load torque.

Equation (34) presents a linearized model of the induction motor over a range of small perturbations at a steady-state operating point. To analyze the stability of the overall IRFO control system, linearized equations of the control part could be derived.

The linear model of the control part is expressed as:

$$s \Delta z = A_z \Delta z + B_r \Delta \omega^*_r \quad (35)$$

$$\Delta u_s = F_z \Delta x + F_r \Delta \omega^*_r \quad (36)$$

where $\Delta z = [\Delta e_s \; \Delta e_{vd} \; \Delta e_{vq}]^T$.

Here, $e_s$, $e_{vd}$ and $e_{vq}$ are the output signals of the PI controllers of the rotor speed, the d-axis stator current component and the q-axis stator current component, respectively.

The substitution of (36) into (34) and taking into account (35) yields:

$$s \Delta x = A \Delta x + B \Delta \omega^*_r + B_T \Delta t_l \quad (37)$$

where $\Delta x = \begin{bmatrix} \Delta x \\ \Delta z \end{bmatrix}$, $A = \begin{bmatrix} A_s & B_z F_z \\ A_x & A_z \end{bmatrix}$, $B = \begin{bmatrix} B_z F_z \\ B_r \end{bmatrix}$, $B_T = \begin{bmatrix} B_r \\ 0 \end{bmatrix}$.

Taking into account the constant load torque, i.e., $\Delta t_l = 0$ and $\Delta \omega_r = C\Delta x$, we can derive from (37) the following transfer function:

$$G(s) = \frac{\Delta \omega_r}{\Delta \omega^*_r} = C(sI - A)^{-1}B \quad (38)$$

The transfer function $G(s)$ consists of six zeros and eight poles. Analyzing the placement of the poles and zeros on the s plane, it is possible to make conclusions about the stability of the RFO control system. Usually, poles and zeros with negative real parts ensure stability. For the stability analysis, the estimation errors in the mutual inductance and rotor resistance are taken into consideration.

The estimation errors of the rotor resistance of two dominant poles and zeros of the transfer function $G$ for $\omega^* = 28.26 \text{ rad/s}$, $t_l = 0.5 t_n$ and $\psi^* = 0.5 \psi_{yn}$.

It can be seen that zero $z_1$ in Fig. 5(b) has a positive real part, causing oscillations in the rotor speed response, as shown in Fig. 6. Fig. 6 shows the impact of the rotor resistance estimation errors on the transient response of the rotor speed step change between 0–28.26 rad/s at time 0 s, and 28.26–48.26 rad/s at time 2 s.
inductance estimation errors always have negative real parts. This is a typical situation in the region of a weakened field over a wide range of rotor speeds and load torques.

6. Rotor Resistance Identification at Variable Magnetizing Levels

To verify the rotor resistance identification, numerous experiments and simulations are performed over a wide range of rotor speeds, magnetizing levels and load torques. To verify simulations, a laboratory experiment is performed. The control algorithm is executed on a dSpace DS1104 board. The induction motor is loaded with a DC generator. The voltage fed PWM inverter with IGBT transistors (SKM100GB123D) is used to drive the induction motor, and an encoder with 1,800 pulses measures the rotor speed. The frequency of space vector PWM modulation is 4 kHz, and the sampling frequency is 8 kHz. The gains of the PI speed controller and the PI current controller are $K_2 = 0.2 \text{ As}$ and $K_i = 2 \text{ V/A}$, respectively. The time constants of the PI speed controller and the PI current controller are $T_2 = 0.1 \text{ s}$ and $T_1 = 0.01 \text{ s}$, respectively. The bandwidths of the speed and current control loops in the low speed region are 56 rad/s and 44 rad/s, respectively. The rotor speed is filtered by the first-order low-pass filter with the cut-off frequency of 100 rad/s.

This paper deals with the IRFO control system at a nominal rotor flux and rotor fluxes lower than the nominal one. The rotor flux being lower than the nominal one plays an important role in induction motor drives operating below the nominal torque most of the time. The rotor flux level being higher than the nominal will cause higher motor losses and, consequently, does not have any practical importance.

Figs. 7 and 8 show the transient response of the rotor speed change between 26.92 and 36.92 rad/s, both with the load torque of 0.5 $\psi_{rn}$, $\psi^*_r = \psi_{rn}$ and $\psi^*_r = 0.7 \psi_{rn}$, respectively.

It has been shown that rotor resistance identification is satisfied at variable magnetizing levels. Both operating conditions give a step response in speed that follows the input command closely.

7. Conclusion

This paper proposes a rotor resistance identification method in an IRFO control system of a saturated induction motor. In our case, a saturated induction motor implies an induction motor with saturable mutual inductance. To achieve rotor resistance identification at variable magnetizing levels, mutual inductance identification in real time is performed. Rotor resistance identification is based on the MRAS theory; the rotor flux magnitude is estimated in the stationary reference frame by the voltage model (reference model) and by the current model (adaptive model). The error signal of the rotor flux magnitude of the two estimators is applied to drive a PI mechanism that provides the identification of the inverse rotor time constant. Therefore, the leakage inductances and stator resistance are constant. The simulations and experiments...
Figure 7. The transient response of the speed step change between 26.92 and 36.92 rad/s and rotor resistance identification, $\psi_r^* = \psi_{rn}$, $t_l = 0.5t_n$; (a) experiment, (b) simulation.

Figure 8. The transient responses of the speed step change between 26.92 and 36.92 rad/s and rotor resistance identification, $\psi_r^* = 0.7\psi_{rn}$, $t_l = 0.5t_n$; (a) experiment, (b) simulation.

show that the rotor resistance identification is valid under variable magnetizing levels.

To analyze IRFO system stability, the system is completely linearized using small deviations of the mutual inductance and the rotor resistance at a steady-state operating point. By plotting trajectories of poles and zeros of the IRFO system transfer function, it is possible to make conclusions on system stability. The estimation errors of the rotor resistance and mutual inductance of ±30% are taken into account. An oscillatory speed step response occurs in a field-weakening region, where the estimated rotor resistance is higher than the actual resistance. As the magnetizing level becomes lower, sensitivity to the rotor
resistance becomes higher. This is particularly pronounced over a lower speed region (below 20% of the rated speed) and at load torques higher than 30% of the rated torque.

References


Appendix A

\[ P_n = 1.5 \text{ kW}, \quad U_n = 380 \text{ V}, \quad \text{four-pole, star}, \quad I_n = 3.81 \text{ A}, \quad n_n = 1.391 \text{ r/min}, \quad L_m = 0.4058 \Omega, \quad L_{sl} = 0.01823 \Omega, \quad L_{cl} = 0.02185 \Omega, \quad R_s = 4.293 \Omega, \quad R_L = 3.866 \Omega \quad \text{(at 20°C)}, \]
\[ t_n = 10.5 \text{ Nm}, \quad R_L = 0, \quad J = 0.0071 \text{ kg m}^2, \quad \psi_{kn} = 0.845 \text{ Wb}. \]

Appendix B

\[ k_{zm} = 0.2139 I_m - 0.369 \quad \text{for} \quad I_m \in [1.725 \text{ A}, \quad 3.08 \text{ A}], \quad k_{zm} = 0.8386 \tanh \left( \frac{0.3179 I_m}{0.0258 I_m} \right) + 0.04208 \quad \text{for} \quad I_m > 3.08 \text{ A}. \]

Appendix C

\[ A_s = \begin{bmatrix} a_3 & \omega_s & \frac{1 - \frac{\sigma L_s}{\sigma L_m}}{I_r} & a_1 L_m & a_1 \omega_s & a_1 L_m \psi_q & -a_2 \psi_q & -a_2 \psi_q \end{bmatrix} \]
\[ B_s = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 & 0 & -\frac{L_m \psi_q^*}{\sigma L_s} & 0 & 0 & -\psi_r \omega_s \end{bmatrix}, \quad A_x = \begin{bmatrix} 0 & 0 & 0 & K_2 & 0 & 0 & 0 \end{bmatrix}, \quad \frac{K_1}{T_1} \quad 0 \quad 0 \quad 0 \quad \frac{K_1 K_2}{T_1} \quad 0 \quad 0 \quad 0 \quad \frac{K_1 K_2}{T_1} \]
\[ \frac{K_2}{T_2} \quad 0 \quad -K_1 \quad 0 \quad -K_1 K_2 \quad 0 \quad 0 \quad 0 \quad 1 - \frac{K_2 \dot{L}_m}{T_r \psi_q^*} \]
\[ F_r = \begin{bmatrix} 0 \end{bmatrix}, \quad F_s = \begin{bmatrix} 0 & 1 & 0 \end{bmatrix}, \quad C = \begin{bmatrix} 0 & 0 & 0 & 1 & 0 \end{bmatrix}, \quad a_1 = 1 \quad \frac{P^2}{\sigma L_s L_r}, \quad a_2 = 3 \quad \frac{P^2}{2 \sigma J L_r}, \quad a_3 = -a_1 \left( R_s L_r + \frac{L_m^2}{T_r} \right). \]

Biographies

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