Standstill Parameter Identification of Vector-Controlled Induction Motor Using Frequency Characteristics of Rotor Bars

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Abstract— This paper suggests a current injection-based estimator to accurately identify standstill induction motor (IM) parameters necessary for the vector control. A mathematical model that faithfully represents the general deep bar effect is introduced. Then, two exciting signals with a different frequency are sequentially injected to track the parameters based on the frequency function of the rotor bar. Little knowledge of the unknown motor allows the proposed methodology to employ a closed-loop control of an injected current, rather than open-loop voltage injection approaches commonly used in sensorless control schemes. Subsequently, this control scheme proactively prevents electrical accidents resulting from an inadequate open-loop voltage injection. We develop a specialized offline commissioning test to compensate the phase delay resulting from the drive, which significantly affects the estimation precision. The effectiveness of the identification technique is validated by means of experiments performed on the three different IMs. The developed algorithm is scheduled to be fully applied to the IM drive system in rolling mill plants of Pohang Steel Company (POSCO) by 2009.

Keywords - Closed-loop control of an injected current, current-injection-based estimator, frequency function of the rotor bar, standstill induction motor (IM) parameters, vector control.

I. INTRODUCTION

Practical vector-controlled induction motor (IM) drive systems require an intensive and time-consuming effort for the tuning of their electrical parameters in order to achieve a satisfactory performance [1]. In most cases, the motors in industry plants originate from different manufacturers, and the parameters are not accurately known prior to startup. At first, it is not easy to control the motor motion in accordance with the given speed/torque profile because the system dynamics is subjected to initial load perturbations. In order to find appropriate parameters, the iterative tuning tests should be performed in the factory. The tuning process is mainly performed by experienced personnel based on field observations and by experimenting with different combinations of parameter values [2].

Modern rolling mill plants, paper winding processes, or hoist crane systems, for example, transfer the motor torque to loads through complex mechanical connections that produce initial load disturbances on the whole system [3-4]. Improper tension control due to incorrect IM parameters often cause mechanical and electrical problems in the commissioning stage or tuning process. It is widely recognized that this process may appear to be an intimidating and time-consuming task even to skilled engineers. Therefore, the motor must be kept at standstill during the parameter identification process to avoid potential damage of the system and to minimize a scheduled downtime. In an attempt to overcome these problems, some previous works have been developed to identify the vector-controlled IM parameters at standstill [5-8].

The time and frequency domain standstill methods [5-6] require special test equipments and expensive procedures. In [7], a single-phase sinusoidal current with a constant frequency was applied to estimate the rotor resistance. The success of this approach, however, hinges on the selection of the injection frequency because the rotor resistance estimate is subject to the skin effect. An alternative standstill estimation method has been reported in [8], where the model is obtained as a linear least-squares fit to a special voltage waveform. For the adequate voltage waveform generation, it is necessary to design an additional pulse width modulation strategy.

This paper suggests a current injection-based estimator to accurately identify standstill IM parameters necessary for the vector control. A mathematical model that faithfully represents the general frequency characteristics of the rotor bar is introduced. Then, two exciting currents with a different frequency are sequentially injected to track the parameters from the frequency function of the rotor bar. Little knowledge of the unknown motor allows the proposed methodology to employ a closed-loop injected current control, rather than open-loop voltage injection approaches commonly used in sensorless control schemes. Subsequently, this control scheme proactively prevents electrical accidents resulting from an inadequate open-loop voltage injection. We develop a specialized offline commissioning test to compensate the phase delay of the drive, which significantly affects the precision of the IM parameters. The effectiveness of the identification technique is validated by means of experiments performed on the three different induction motors. The developed algorithm is scheduled to be fully applied to the IM drive system in rolling mill plants of Pohang Steel Company (POSCO) in Korea by 2009.

II. FREQUENCY CHARACTERISTICS OF ROTOR BAR

The current distribution in the shorted rotor bars may vary significantly with frequency, giving rise to significant
variations in the rotor resistance and inductance. For the simple rectangular bar with a depth \(d\), the effective impedance of the bar can be represented as [9]

\[
Z_{bar} = R_{r-hf} + jX_{r-hf} = \alpha dR_{r-dc} \left[ \frac{\sinh 2\alpha d + \sin 2\alpha d}{\cosh 2\alpha d - \cos 2\alpha d} + j \frac{\sinh 2\alpha d - \sin 2\alpha d}{\cosh 2\alpha d - \cos 2\alpha d} \right]
\]

(1)

where

\[
\alpha = \sqrt{\frac{2\pi f \mu_0}{\rho}}
\]

and \(f\) represents the frequency, \(\mu_0\) indicates the permeability of air, \(\rho\) is the resistivity of the conducting bar, and \(R_{r-dc}\) denotes the dc rotor resistance. Fig. 1 shows the plot of the hyperbolic function of (1).

![Figure 1. Plot of the hyperbolic function of (1).](image1)

It can be noticed from Fig. 1 and (1) that the effective resistance and leakage reactivity approach equality as the frequency or bar depth increases. Thus, if \(\alpha d > 2\),

\[
R_{r-hf} \approx X_{r-hf}.
\]

(2)

### III. PROPOSED IM PARAMETER IDENTIFICATION

The \(d\)-axis equivalent circuit of an IM at standstill can be drawn as shown in Fig. 2. For the identification, we adopt a method of injecting an alternating \(d\)-axis current with a dc bias while letting the \(q\)-axis current be equal to zero.

\[
i_{ds} = I_{ds} + i_{dh} = I_{ds} + I_m \cos \omega_h t
\]

(3)

where \(I_m\) is the amplitude of the ac current and the dc bias \(I_{ds}\) is determined as the nominal current value on a name plate of the motor [10]. When the injection frequency \(\omega_h\) is high enough, most of the high-frequency (HF) current flows through the rotor branch.

In this paper, two exciting signals with a different frequency are sequentially injected to determine the parameters. First, the HF current is injected to estimate the stator resistance and the stator leakage inductance. Then, a low-frequency (LF) current test is performed to find the value of the rotor leakage inductance and the rotor resistance at the rated slip frequency.

In this paper, the subscript \(lf\) or \(hf\) of each parameter indicates that the corresponding parameter is related to the LF or the HF test process, respectively.

![Figure 2. \(d\)-axis equivalent circuit of an IM at standstill.](image2)

#### A. HF Current Test

From Fig. 2, the \(d\)-axis voltage equation at steady-state is derived as

\[
V_{ds} = R_s I_{ds} + (R_s + R_f) i_{dh} - \omega_h L_{eq} I_{m} \sin(\omega_h t)
\]

(4)

where \(L_{eq} = L_{ls} + L_{lr}\). A low-pass filter (LPF) is used to obtain the equivalent stator resistance including the voltage drop of power devices and the inverter-motor cables as

\[
\hat{R}_s = \frac{\text{LPF}(V_{ds})}{I_{ds}} = \frac{\text{LPF}(V_{ds})}{\text{LPF}(i_{dh})}.
\]

(5)

For a practical application, it is assumed that the skin effect of stator parameters is negligible in the frequency range of interest (<250 Hz) compared to that of rotor parameters [11]. Then, the equivalent resistance and inductance under the HF injection condition can be identified from the product of the ac voltage and current component as followings:

\[
v_{dh} \cdot i_{dh} = R_{eq-hf} i_{dh}^2 - \omega_h L_{eq-hf} I_{m}^2 \sin(\omega_h t) \cos(\omega_h t)
\]

\[
= \frac{1}{2} R_{eq-hf} I_{m}^2 + \frac{1}{2} R_{eq-hf} I_{m}^2 \cos(2\omega_h t) - \frac{1}{2} \omega_h L_{eq-hf} I_{m}^2 \sin(2\omega_h t)
\]

(6)

where

\[
R_{eq-hf} = R_s + R_{r-hf}.
\]

(7a)

\[
L_{eq-hf} = L_{ls} + L_{lr-hf}.
\]

(7b)

From (6), the equivalent resistance and inductance can be estimated as

\[
\hat{R}_{eq-hf} = \frac{\text{LPF}(v_{dh} \cdot i_{dh})}{\text{LPF}(i_{dh})}
\]

(8a)

\[
\hat{L}_{eq-hf} = \frac{\sqrt{\text{LPF} \left( \left( v_{dh} - \hat{R}_{eq-hf} i_{dh} \right)^2 \right)}}{\omega_h \sqrt{\text{LPF}(i_{dh})^2}}.
\]

(8b)
By combining (5) and (7a), the HF rotor resistance \( R_{r\_bf} \) becomes
\[
\hat{R}_{r\_bf} = \hat{R}_{eq\_bf} - \hat{R}_s .
\] (9)

The HF rotor leakage inductance \( L_{lr\_bf} \) can be described using (2) and (9) as
\[
\hat{L}_{lr\_bf} = \frac{X_{lr\_bf}}{\omega_h} = \frac{\hat{R}_{r\_bf}}{\omega_h} .
\] (10)

Hence, the stator leakage inductance can be identified as
\[
\hat{L}_{ls} = \hat{L}_{eq\_bf} - \hat{L}_{lr\_bf} .
\] (11)

From (1) and (10), the HF rotor leakage inductance can be rewritten as
\[
\hat{L}_{lr\_bf} = \frac{1}{\sqrt{jh}} \frac{3}{2} \frac{A_{lr\_bf}}{d} \frac{L_{lr\_dc}}{\sqrt{\alpha}} .
\] (12)

where \( L_{lr\_dc} \) denotes the dc rotor leakage inductance and the constant \( K_{lr} \) represents a proportional coefficient with respect to the inverse square root of the frequency.

**B. LF Current Test**

The same test with a LF current is performed again to find the rotor leakage inductance and resistance at the rated slip frequency. Using the same approach for the equivalent inductance can give
\[
\hat{L}_{eq\_bf} = \omega_l \hat{L}_{PF} \left\{ v_{al} - \hat{R}_{eq\_bf} \cdot i_{al} \right\}^2
\] (13)

where \( v_{al} \) and \( i_{al} \) represent the ac voltage and current component of the LF current test, and \( \omega_l \) is the injection current frequency in the LF current test.

Then, the LF rotor leakage inductance \( L_{lr\_bf} \) can be described as
\[
\hat{L}_{eq\_bf} - \hat{L}_{ls} = \hat{L}_{lr\_bf}
\] (14)

where \( \alpha_l = \frac{\pi f_l \mu_r}{\rho} \).

Then, the value of \( \alpha_l d \) is obtained by solving (14)
\[
\alpha_l d = \frac{\pi \mu_r}{\rho} - d \sqrt{f_l} = K_{ad} \sqrt{f_l} .
\] (15)

Using \( K_{ad} \) in (15) and \( \hat{R}_{r\_bf} \) gives
\[
\hat{R}_{r\_dc} = \frac{\hat{R}_{r\_bf}}{K_{ad} \sqrt{f_h}} .
\] (16)

and
\[
\hat{L}_{lr\_dc} = \frac{2}{3} K_{ad} \sqrt{f_h} \hat{L}_{lr\_bf} .
\] (17)

Finally, the rotor resistance and leakage inductance at the rated slip frequency can be uniquely determined as
\[
\hat{R}_r = K_{ad} \sqrt{f_{slip}} \frac{\sin(2K_{ad} \sqrt{f_{slip}}) + \sin(2K_{ad} \sqrt{f_{slip}})}{\cosh(2K_{ad} \sqrt{f_{slip}}) - \cos(2K_{ad} \sqrt{f_{slip}})}
\] (18)

\[
\hat{L}_{lr} = \frac{3}{2K_{ad} \sqrt{f_{slip}}} \left[ \frac{\sin(2K_{ad} \sqrt{f_{slip}}) - \sin(2K_{ad} \sqrt{f_{slip}})}{\cosh(2K_{ad} \sqrt{f_{slip}}) - \cos(2K_{ad} \sqrt{f_{slip}})} \right]
\] (19)

where \( f_{slip} \) is the rated slip frequency. These results explain that the proposed identification method can provide a deep-bar-effect curve of the effective rotor resistance and leakage inductance over the whole frequency range.

**IV. SELECTION OF INJECTION FREQUENCY AND CURRENT CONTROLLER DESIGN**

**A. Selection of Injection Frequency**

Selection of \( f_h \) in the HF injection test involves a tradeoff between the estimated accuracy due to a better signal-to-noise ratio and the maximum sampling frequency of the phase current. In the HF current injection test, \( f_h \) should be chosen such that \( \alpha d > 2 \) and it depends on the bar depth. Thus, the minimum available frequency can be obtained as
\[
f_{h\_min} = 2 \sqrt{\frac{\pi \mu_r}{\rho}} \times d
\] (20)

where \( \rho_{(Al)} \) denotes the resistivity of aluminum. Fig. 3 shows a plot of \( f_{h\_min} \) with respect to the bar depth of commercial IMs with the power rating ranging from 5 kW to 100 kW.

It can be observed from this example that all values of \( f_{h\_min} \) are below 200 Hz. This implies that the value of 200 Hz is sufficient to satisfy the condition of \( \alpha d > 2 \) down to the power range of 5 kW.

In the LF test, \( f_l \) is chosen so that
\[
\frac{|j\omega_l f_l|}{\sqrt{f_r + j\omega_l f_r}} \geq 9 \text{.}
\] (21a)
5.1 \alpha d < 1.5. \tag{21b}

The motor parameters in (21a) denote the values obtained from the name plate data [10]. Equation (21a) comes from the condition that more than 90% of the ac current should flow through the rotor branch.

The maximum available frequency can be obtained as

\[
f_{\text{f}_{\text{max}}} = \left( \frac{4.5r_r}{\pi \sqrt{R_m^2 - (9f_h)^2}} \right) \leq f_1 < f_{\text{f}_{\text{max}}}. \tag{22}\]

A plot of \( f_{f_{\text{max}}} \) as a function of \( d \) is given in Fig. 4. Thus, \( f_1 \) is approximately determined from (21a) and (22) as

\[
\frac{4.5r_r}{\pi \sqrt{R_m^2 - (9f_h)^2}} \leq f_1 < f_{f_{\text{max}}}. \tag{23}\]

B. Offline Commissioning for Phase Delay Compensation

In the HF test, a phase delay of the feedback current and a PWM generation delay can give a detrimental effect on the proposed identification accuracy. In this work, we propose an offline commissioning process to find the delay effect coming from the drive prior to start-up. The HF current and voltage with the delay factor can be considered as

\[
i_{dh} = I_m \cos(\omega_h t - \delta) \tag{24a}
\]
\[
v_{dh} = V_m \cos(\omega_h t + \phi + \delta_{PWM}) \tag{24b}
\]

where \( \delta \) denotes the phase delay of the sampled current, \( V_m \) is the amplitude of the ac voltage, \( \phi \) represents the phase angle, and \( \delta_{PWM} \) is the corresponding angle of the PWM generation delay. Applying (24a) to (6) gives

\[
\tilde{R}_{eq-hf} = \frac{L \tilde{v}_{dh} \cdot \tilde{i}_{dh}}{V_m \sin(\phi + \delta_i + \delta_{PWM})}. \tag{25a}
\]
\[
\tilde{L}_{eq-hf} = \frac{V_m}{\omega_h I_m} \cos(\phi + \delta_i + \delta_{PWM}). \tag{25b}
\]

Since \( \phi \) almost reaches 90 degrees in this test, the estimation error more severely affects \( \tilde{R}_{eq-hf} \) than \( \tilde{L}_{eq-hf} \).

An accurate estimation of the phase delay is critical for the success of the proposed identifier design. Here, we define a new variable “Phase Delay Compensation Index (PDCI)” \( K_\delta \) using (1) and (9) as

\[
K_\delta = \frac{\tilde{R}_{r_{\text{dc}}}}{\tilde{f_r}} = \frac{\tilde{R}_{eq-hf} - \tilde{R}_s}{\tilde{f_r}} = \sqrt{\frac{\pi d_c}{\rho}} \cdot d \cdot R_{r_{\text{dc}}}. \tag{26}\]

Equation (26) indicates that \( K_\delta \) must maintain a constant value irrespective of \( f_r \) if no phase delay is involved in the identification process. Fig. 5 shows the offline test result for the testing drive of IM #1 with unknown phase delay.

This plot reveals a trajectory that varies with injection frequency in the case of the incorrect delay compensation. The inaccurate state is clearly noticeable in the trajectory of \( K_\delta \), giving a high sensitivity to detuning. From these observations, it can be concluded that the correct phase delay is about 138 \( \mu s \) for the testing drive.

C. Current Controller Design

For the single-phase (d-axis) HF current control, in this paper, we introduce a simple proportional-integral plus resonant (PI+R) controller which achieves infinite gain at the resonant frequency of concern. The main objective here is that the dc bias current control is performed through the PI action and the resonant controller is responsible for the HF current regulation. In combination with both controllers, this approach can effectively deal with the closed-loop control of the single-phase current with a HF component. The PI+R design method leads to an advantage of easy implementation of the proposed
method. The transfer function of the proposed PI+R controller can be designed as in (27).

\[
G_{AC}(s) = K_p + \frac{K_L}{s} + \frac{K_R\omega_{cut}s}{s^2 + \omega_{cut}s + \omega_1^2}
\]  

(27)

where \(\omega_{cut}\) and \(k_L\) represent the 3dB cut-off frequency and the gain of the resonant controller.

Unfortunately, designing the proposed PI+R controller and the LPF is not trivial as it does not work well at low frequencies. Therefore, the LF current injection test is performed by a simple linear observer instead of the LPF-based scheme of (13). For the purpose of designing an observer, a state-space model can be proposed as

\[
\begin{align*}
\dot{x}_1 &= v_{dl} \sin(\omega_t) \\
\dot{x}_2 &= \frac{1}{2} \hat{R}_{eq, yf} I_m \sin(2\omega_t) + \frac{1}{2} \omega_t \hat{L}_{eq, yf} I_m \cos(2\omega_t) \\
\dot{x}_3 &= \frac{1}{2} \hat{R}_{eq, yf} I_m \cos(2\omega_t) - \frac{1}{2} \omega_t \hat{L}_{eq, yf} I_m \sin(2\omega_t)
\end{align*}
\]  

(28)

where \(v_{dl}\) can be simply obtained by subtracting \(\hat{R}_s I_{ds}\) from \(V_{dc}\).

Here, we choose the observer which is designed as (29).

\[
\dot{x} = Ax + LC(x - \hat{x})
\]

\[
y = \hat{x}_2 = \begin{bmatrix} 0 & 0 & 2\omega_t \\ 0 & -2\omega_t & 0 \end{bmatrix} \begin{bmatrix} l_1 \\ l_2 \\ l_3 \end{bmatrix} \]

where \(L = [l_1, l_2, l_3]^T\) is a gain matrix. The observer gain selection and the phase delay error effect are not at all critical in this test since the state dynamics is much slower compared to the sampling frequency. Then, \(\hat{L}_{eq, yf}\) is obtained as

\[
\hat{L}_{eq, yf} = -\frac{2}{\omega_t} \hat{I}_m = (\hat{x}_1 - \hat{x}_2)
\]  

(30)

and \(\hat{L}_{lr, yf}\) can be computed by combining (14) and (30).

Fig. 6 shows the overall block diagram of the proposed strategy.

V. EXPERIMENTAL RESULTS

The proposed algorithm is implemented on three different inverters to drive three different IMs described in Table I. In the table, the equivalent stator resistance \(R_s\) including a voltage drop of power devices and the inverter-motor cables was extracted from the laboratory dc test [7] and \(R_r\) is compared to \(\hat{R}_r\), tuned by considering the indirect vector control performance. The errors of the stator resistance and the rotor resistance are within \(\pm 5\%\) and \(\pm 20\%\), respectively. Such errors are reasonably tolerable to perform the practical vector control, even with the name-plate-based \(l_m\).

TABLE I

<table>
<thead>
<tr>
<th></th>
<th>Motor IM #1 (1.5kW)</th>
<th>Motor IM #2 (5.5kW)</th>
<th>Motor IM #3 (17.5kW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_s/R_r[\Omega])</td>
<td>2.37/2.47</td>
<td>0.934/0.902</td>
<td>0.202/0.197</td>
</tr>
<tr>
<td>(\Delta R_s[%])</td>
<td>-4.1</td>
<td>+3.5</td>
<td>+2.5</td>
</tr>
<tr>
<td>(\hat{L}_{di}[mH])</td>
<td>10.5</td>
<td>11.6</td>
<td>4.9</td>
</tr>
<tr>
<td>(\hat{L}_{lr}[mH])</td>
<td>2.15</td>
<td>3.4</td>
<td>1.8</td>
</tr>
<tr>
<td>(\hat{R}_r/R_r[\Omega])</td>
<td>0.66/0.7</td>
<td>0.434/0.522</td>
<td>0.126/0.135</td>
</tr>
<tr>
<td>(\Delta R_r[%])</td>
<td>-5.8</td>
<td>-16.8</td>
<td>-6.7</td>
</tr>
<tr>
<td>(l_m/l_m[mH])</td>
<td>192/134</td>
<td>138/86.8</td>
<td>54.8/41.9</td>
</tr>
</tbody>
</table>

TABLE II

<table>
<thead>
<tr>
<th></th>
<th>Motor IM #1 (1.5kW)</th>
<th>Motor IM #2 (5.5kW)</th>
<th>Motor IM #3 (17.5kW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sampling Frequency [kHz]</td>
<td>10</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>(f_s[Hz])</td>
<td>250</td>
<td>200</td>
<td>200</td>
</tr>
<tr>
<td>(f_l[Hz])</td>
<td>30</td>
<td>30</td>
<td>20</td>
</tr>
<tr>
<td>Phase Delay [\mu s]</td>
<td>138</td>
<td>319</td>
<td>358</td>
</tr>
</tbody>
</table>

Figure 6. Proposed identification and current control scheme. (a) Block diagram of HF current test. (b) Block diagram of LF current test.
Once the motor is driven in obedience to a specified command, the magnetizing inductance $L_m$ or the rotor time constant can be accurately estimated [12]. The current sampling frequency of each inverter and test conditions are summarized in Table II. In all tests, $I_m$ is set to 9% of $I_{ds}$. At commissioning stage the deadtime effect of the drive was rejected from a prescribed compensation scheme [13].

In all tests, the gain of the resonant controller $k_r$ and $\alpha_{cut}$ are fixed to 6720 and 15.7 rad/s, respectively. In the testing system, the bandwidth of the PI controller and the observer are given as 2000 and $4\omega_f$ rad/s, respectively.

The time-domain responses of the current regulation performance in the HF and LF test are plotted in Fig. 7 and 8. At the top, an overlay plot of the $d$-axis current command and the controlled current is plotted. The bottom plot shows the $dq$-axis voltage command at the same time. With the use of the PI+R control and the proposed observer, it can be observed that the current controller provides complete control over the LF and HF injection.

A HF current injection test was chosen for the first step of the identification of IM #1 as shown in Fig. 9. At instant $t = 0.6$ s, the identification algorithm started while the rotor speed was kept zero. After 0.2 s, all the estimated parameters were stabilized and resulted in a close fit of their known values.

At the end of the first estimation process, a LF current injection test started at instant $t = 0.4$ s as shown in Fig. 10. The estimated parameter quickly converges the known value and it is observed that the algorithm is stable at steady-state. The total time taken for both tests is within 1 s. This makes the proposed algorithm very promising for automated self-commissioning of modern industry plants combined with a huge number of IMs.

Combining (14) and (15) yields the bar depth of IM #1 as $d=1.6$ cm. This value and obtained rotor parameters are used to compute deep-bar-effect curves of IM #1 as plotted in Fig. 11. Note that resulting curves over 250 Hz indicate a perfect match between rotor resistance and leakage reactance as given in (2). This implies that the proposed method can provide frequency characteristics of rotor bars over the whole frequency range. Thus, the main idea is expected to lead to a wide cross-section
of exciting ongoing researches such as fault diagnosis and the condition monitoring of IMs.

In order to reconfirm how close \( \hat{R}_r \) matches “true” one, we performed the load test using another motor coupled to the testing IM as shown in Fig. 12. An SS-201 torque sensor from Ono Sokki with a maximum level of 200 Nm and ±0.2% accuracy is connected to read the actual load torque. In this test, the accurate magnetizing inductance is employed just to evaluate the accuracy of \( \hat{R}_r \). Fig. 13 shows the load test result for the indirect vector-controlled IM #3 (17.5 kW). Here, we observe that the torque error is within 20%. The result shows that the estimated rotor resistance is reasonably accurate for performing the practical vector control.

VI. CONCLUSION

This paper presents a standstill parameter identification method for vector-controlled IMs by means of two simple current injection tests. A deep-bar model has been presented and a proper closed-loop current controller design has been incorporated in the process as an injection technique. Using the offline phase delay correction procedure, the accurate parameter identification is made available. The proposed algorithm is implemented in the three different IM systems and verified to be appropriate for the vector control of unknown induction motors. Since the proposed identification method can provide a deep-bar-effect curve over the whole frequency range, the main idea can be extended to other applications such as the fault diagnosis/detection and the condition monitoring of IMs.

REFERENCES