Auto-Measurement of Induction Motor Parameters

Kyung-Seo Kim† and Sung-Hoon Byun*

Abstract - This paper presents the parameter measurement methods for high performance drive of induction motors, which are suitable for the self-commissioning function of commercial inverters. In this study, some factors that affect the accuracy of parameter measurement are examined. Measuring methods and conditions that are best fit to each parameter measurement procedure are then proposed. All the measurement procedures can be done without any auxiliary equipment, so that those can be easily adopted as self-commissioning functions of commercial inverters. To improve the measuring accuracy, least square approximation methods are adopted during the measurement procedure. The validity of the proposed methods are confirmed through experiments.

Keywords: auto-measurement, commissioning, induction motors, parameters

1. Introduction

Most high performance drives of induction motors including the vector control and the sensorless vector control make use of equivalent machine models. Therefore the performance of these kinds of drives highly depends on the accuracy of the machine parameters. One important request to the modern adjustable drive systems from the market is the auto-commissioning functions including auto-measurement of motor parameters. Many commercial drive systems have auto-measuring functions that are necessary for high performance drive systems.

Traditional methods for measuring motor parameters are the locked-rotor test and no-load test [1]. One drawback of these methods is that they require additional means for locking the rotor or rotating the rotor in synchronous speed. Another drawback is that the test condition of these methods is somewhat different from the real operating condition. During normal operation, for example, the slip frequency of induction motors is lesser than the rated slip. On the other hand, the slip frequency of induction motors during a locked-rotor test is equal to the frequency of the power source. Therefore, due to high slip frequency, the effect of deep-bar or double cage causes inaccuracy in measurement.

Recent publications for measurement of motor parameters include the method of test signal injection, MRAS, the single phase excitation, etc. [2-9]. Some are based on an assumption that motors are ideal linear systems and parameters are always constant regardless of measurement conditions. But there are many factors that affect the variation of machine parameters and accuracy of measurement results. Some hidden characteristics of machines can largely affect the measuring accuracy depending on which measurement method or measuring conditions are used. Also inverters have many nonlinear factors such as characteristics of switching devices, dead time effect, accuracy of sensors, etc.

In this study, some factors that affect the accuracy of parameter measurement are examined. Measuring methods and conditions that are best fit to each parameter are then proposed. All of the measurement procedures can be executed without any auxiliary equipment.

2. Measurements of Motor Parameters

2.1 Equivalent Circuit of Induction Motors

Fig. 1 shows the inverse-Γ type equivalent circuit used for this study. The inverse-gamma type is particularly convenient for calculating variables of equivalent circuit because the active component and the reactive component are decoupled in the rotor circuit.

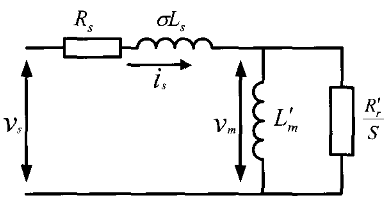

Fig. 1 Equivalent circuit of induction motors

The magnetizing inductance, rotor resistance and rotor time constant of Fig. 1 are newly defined from T type circuit parameters as follows.

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$$L'_m = L_m^2 / L_r = L_s(1-\sigma) \quad (1)$$

$$R'_r = R_r \frac{L_m^2}{L_r^2} \quad (2)$$

$$T_r = L_r / R_r = L'_m / R'_r \quad (3)$$

2.2 System Setup

All the measurement functions have been implemented on a commercial inverter system. The main control function of these inverters is the V/f control, and the indirect vector control and the sensorless vector control are optional. The main inverter processor is the TMS320F240, 16bit fixed point DSP. Power switching devices are IGBTs, and switching frequency is 10[kHz] nominal. Real phase current is measured using hole current sensors. The phase voltage is estimated from reference voltage considering the DC link variation and the dead time effect. Rotor speed is measured using an increment encoder.

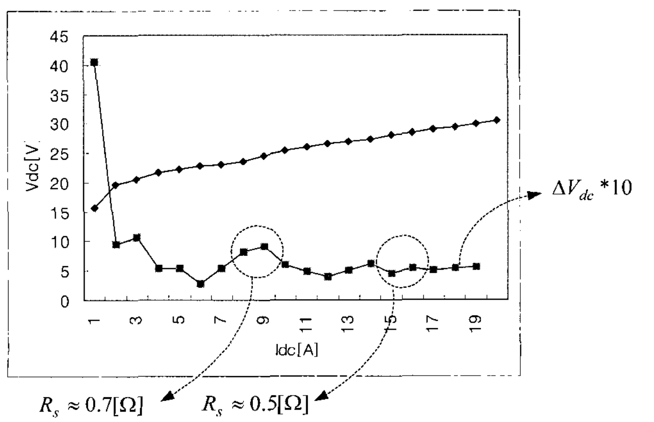
The test motor is the 7.5[kW] induction motor, and its rating is as shown in Table 1.

Table 1 Rating of test motor

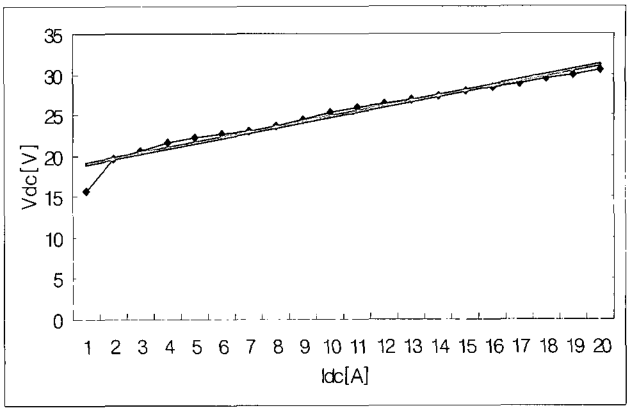
Type	3-phase induction motor
Capacity	7.5[kW]
Rated voltage	380[V]
Rated current	15.2[A]
Rated speed	1730[rpm]

2.3 Measurement of Stator Resistance

The most convenient method used to measure stator resistance using an inverter is to apply DC voltage to the motor terminals, then measure the motor current. The final value acquired through this method is the sum of the stator resistance and equivalent inverter resistance.



(a)



(b)

Fig. 2 Measurement of stator resistance (a) variation of stator resistance (b) approximation of stator resistance value

In a real system there exist nonlinear factors such as dead time effect, voltage drop of switching and sensor offset, etc.

These factors decrease the accuracy of measurement, so that they should be considered during measurement. One effective method to remove these factors is measuring the voltage and the current in two points and then calculating resistance by (4) [2, 3].

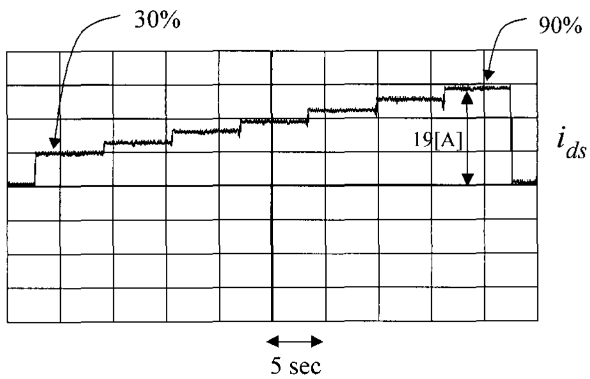


Fig. 3 Excitation current for measurement of stator resistance

$$R_S = \frac{(V_{s2} - V_{offset}) - (V_{s1} - V_{offset})}{I_{s2} - I_{s1}} = \frac{V_{s2} - V_{s1}}{I_{s2} - I_{s1}} \quad (4)$$

In (4), the offset value is canceled and does not affect the calculation of stator resistance. But (4) is true only if the offset value is constant through the entire operating range. In a real system, offset value cannot be considered as a constant value because inverter equivalent resistance changes according to the current level. Fig. 2(a) shows the variation of equivalent resistance according to the current level. If two points are selected in the vicinity of 50[%] of rated current, equivalent resistance value is calculated as

0.7[ohm]. If the measuring point moves to the vicinity of the rated current the equivalent resistance value decreases to 0.5[ohm]. How to select the measuring points causes almost 40[%] difference of measuring results in this case. To overcome this problem, multi-point measurement is adopted in this study. DC current is injected with several current levels, and then voltage values are measured at each current level. A straight line closest to the measured points is acquired through the least square approximation method [11], which is briefly summarized in the Appendix. The thick line in Fig. 2(b) is the approximation line, the slope of which is taken to be equivalent to the stator resistance.

The sequence for measurement of the stator resistance is as follows. At first 30[%] of rated DC current value is applied to the motor terminal and the magnitude of the DC current is increased to 90[%] through 7 steps. At the same time, voltage magnitude is measured at each step. Fig. 3 shows the measurement process of resistance. In each step, voltage is measured 40,000 times, and then the average voltage is calculated.

If the approximation line that is close to the measured points is represented by (5),

$$V = R_s I_i + V_{offset} \tag{5}$$

the equation to calculate equivalent resistance using the least square approximation is as follows.

$$R_s = l_0 V_0 + l_1 V_1 + \cdots + l_7 V_7 \tag{6}$$

where $l_0 \ldots l_7$ are pre-calculated constants using the Appendix, and $V_0 \ldots V_7$ are measured voltage values for each current level.

In the Appendix, if applied current is defined as follows and applied to (A5) with $M = 7, i = 1 \cdots M, n = 1$,

$$\begin{aligned}[x_1, x_2, \ldots, x_7] &= [I_1, I_2, \ldots, I_7] \\ &= [0.3I_{(rated)}, 0.4I_{(rated)}, \ldots, 0.9I_{(rated)}]\end{aligned} \tag{7}$$

Then, the slope of the approximation line can be calculated as in the following equation, which is the value of stator resistance R_s.

$$a_1 = R_s = \frac{-0.357}{I_{(rated)}}[3,2,1,0,-1,-2,-3][V_1, V_2, V_3, V_4, V_5, V_6, V_7]^T \tag{8}$$

2.4 Measurement of Leakage Inductance

If single-phase voltage is applied to the motor terminals, the motor is at a standstill, so that the slip frequency is equivalent to the frequency of applied voltage. If the frequency is high enough, the impedance of magnetizing inductance is much higher than the rotor resistance value, which means that the magnetizing circuit can be omitted in the equivalent circuit. Then, the equivalent circuit can be simplified as in Fig. 4. The difference of calculation results between Fig. 4 and Fig. 1 is under 0.5[%], which is negligible in terms of practical usage.

If single-phase sinusoidal voltage of (9) is applied to the motor terminal, lagging current flows in the motor phase. This phase current can be divided into the active current component and the reactive current component. The active current component has the same phase with the applied voltage, and the reactive current component lags 90°. The RMS value of the active current component $I_{P(rms)}$ and the reactive current component $I_{Q(rms)}$ can be calculated using (10) and (11).

$$v_s = V_s \sin(\omega_e t) \tag{9}$$

$$I_{P(rms)} = \frac{1}{\sqrt{2}\pi} \int_0^{2\pi} i_s \sin(\omega_e t) d\theta \tag{10}$$

$$I_{Q(rms)} = \frac{-1}{\sqrt{2}\pi} \int_0^{2\pi} i_s \cos(\omega_e t) d\theta \tag{11}$$

$$V_{s(rms)} = V_s / \sqrt{2} \tag{12}$$

$V_{s(rms)}$ is rms value of applied voltage.

Leakage inductance σL_s and total resistance R_r' can be calculated from (13) and (14), where P is active power and Q is reactive power.

$$\sigma L_s = \frac{Q}{\omega_e I_{s(rms)}{}^2} = \frac{I_{Q(rms)} V_{s(rms)}}{\omega_e (I_{P(rms)}^2 + I_{Q(rms)}^2)} \tag{13}$$

$$R_r' + R_s = \frac{P}{I_{s(rms)}^2} = \frac{I_{P(rms)} V_{s(rms)}}{(I_{P(rms)}^2 + I_{Q(rms)}^2)} \tag{14}$$

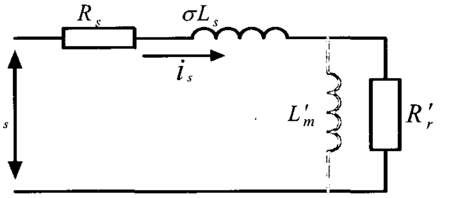


Fig. 4 Modified equivalent circuit with high frequency excitation

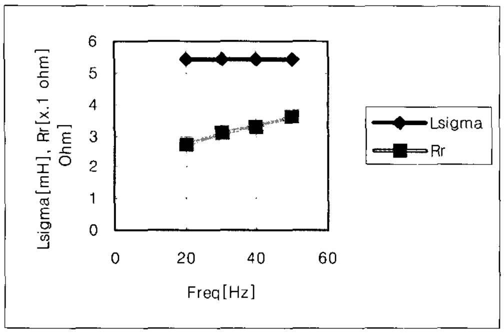


Fig. 5 Leakage inductance and rotor resistance with frequency change

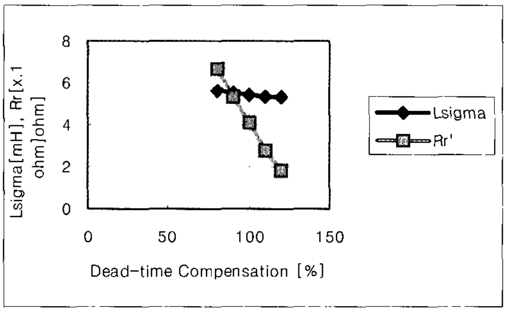


Fig. 6 Leakage inductance and rotor resistance with dead-time compensation voltage

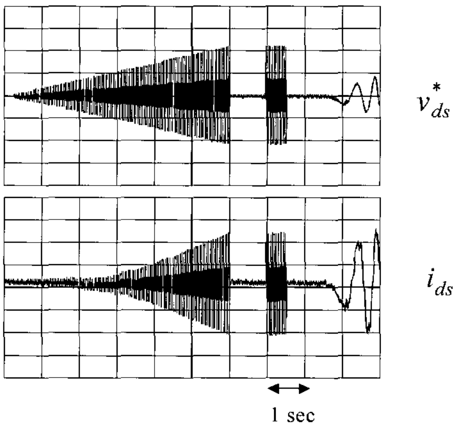


Fig. 7 Single phase voltage for leakage inductance measurement

If a high frequency signal is injected, slip frequency is much higher than rated slip because the rotor is at a standstill. This affects the variation of rotor resistance if the rotor has a deep bar or double cage structure. Fig. 5 illustrates this effect. Leakage inductance has no relation to frequency change. Conversely, rotor resistance value changes according to variation of frequency.

Dead-time effect also affects the measurement of rotor resistance. Fig. 6 is the variation of leakage inductance and rotor resistance when dead-time compensation voltage is varied from 80[%] to 120[%]. In Fig. 6, the variation of leakage inductance is not much compared with the variation of rotor resistance. The reason for this is that the compensation voltage has the same phase with the current, so that the dead time voltage only affects the variation of active power if only a fundamental component is considered.

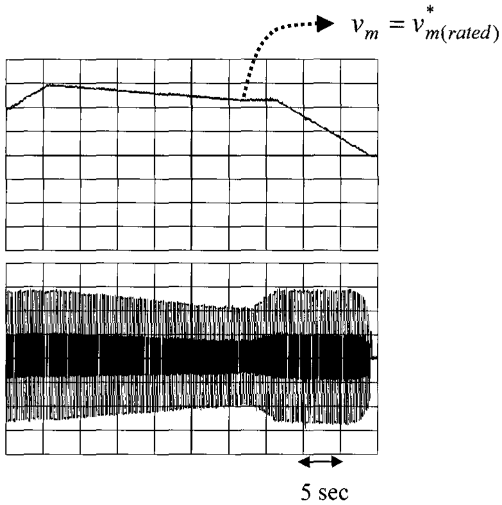


Fig. 8 The voltage magnitude (upper) and the phase current (lower) during the measurement of magnetizing inductance

From the above results it can be concluded that only leakage induction can be measured correctly by the single-phase excitation method. Fig. 7 presents the single-phase voltage and current waveforms during the leakage inductance measurement. The voltage is gradually increased until the current reaches the desired value. After that, the reactive power is measured from the current and the voltage.

2.5 Determination of Magnetizing Current and Measurement of Magnetizing Inductance

Most high performance drives contain current control loops. Therefore it should be guaranteed that the peak value of the terminal voltage is always lower than the dc link voltage even under the rated speed and the rated load condition in order to assure the voltage margin for current control. If the magnetizing current is determined through no-load test by applying rated frequency and rated voltage, the current may be uncontrollable under rated speed and rated load condition because there is insufficient voltage margin required for the current to be controllable. Another problem of the no-load test is that it requires equipment that enables test motors to run in synchronous speed.

This is not a practical method for the auto-measurement function of a commercial inverter. It may be possible to calculate air-gap voltage in rated load condition by using nameplate values of test motors. Assuming that the stator resistance and leakage inductance are already known, the rated airgap voltage $v_{m(rat)}$ of Fig. 1 under rated load can be calculated as follows.

$$\vec{v}_{m(rat)} = v_{s(rat)} - (i_{s(rat)}\cos\varphi + ji_{s(rat)}\sin\varphi)(R_s + j\omega_{e(rat)}\sigma L_s) \quad (15)$$

Where $v_{m(rat)}$ is the voltage of magnetizing circuit under rated speed $\omega_{e(rat)}$, rated voltage $v_{s(rat)}$ and rated current $i_{s(rat)}$. If $v_m = v_{m(rat)}$, magnetizing current I_m and rotor flux is always constant regardless of load variation and v_s is always lower than $v_{s(rat)}$ until full load condition.

Measurement begins with the acceleration of the motor until the motor speed reaches the rated speed. Once the motor speed reaches the rated speed and the supply voltage becomes the rated value, supply voltage v_s decreases until v_m equal to $v_{m(rat)}$ of (15). v_m is calculated by (16).

$$\begin{aligned}\vec{v}_m &= v_{m(real)} + jv_{m(imag)} \\ &= v_s - (i_{s(real)} + ji_{s(imag)})(R_s + j\omega_{e(rat)}\sigma L_s)\end{aligned} \quad (16)$$

When $v_m = v_{m(rat)}$, rated magnetizing current can be calculated as in the following equation.

$$i_{m(rated)} = i_m = \frac{v_{m(real)}i_{s(imag)} - v_{m(imag)}i_{s(real)}}{v_m} \quad (17)$$

Magnetizing inductance is calculated at the same time when $i_m = i_{m(rat)}$ as follows.

$$L'_m = \frac{v_{m(rat)}}{i_{m(rat)}\omega_{e(rat)}} \quad (18)$$

Fig. 8 indicates voltage magnitude and phase current during measurement of magnetizing inductance. By the proposed method, proper magnetizing current and magnetizing inductance can be measured precisely regardless of load.

2.6 Measurement of Rotor Time Constant

For measuring rotor time constant, methods that compare the real variables and the reference variables from the reference model are used. Among various reference models, the d-axis voltage model is most sensitive to the variation of rotor time constant [10]. Equation (19) is d-axis reference voltage under field oriented condition.

$$v_{ds}^{e*} = R_s i_{ds}^e - \omega_e L_\sigma i_{qs}^e \quad (19)$$

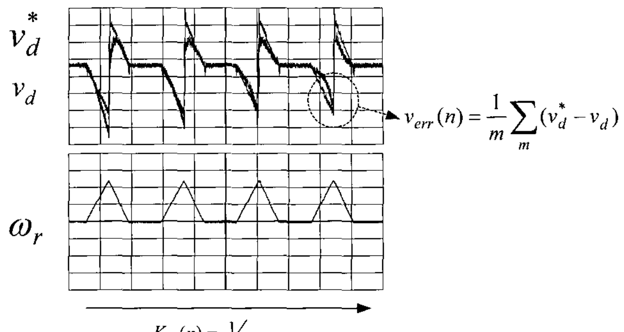


Fig. 9 D-axis voltage and rotor speed during slip constant measurement

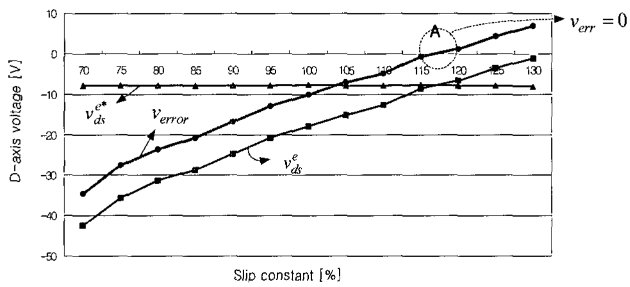


Fig. 10 Variation of d-axis voltage with increment of slip constant

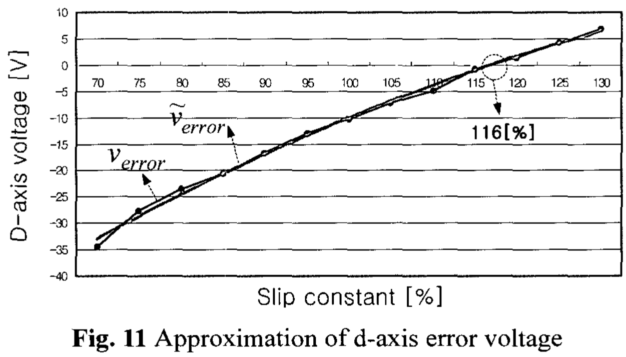


Fig. 11 Approximation of d-axis error voltage

To measure the rotor time constant, the indirect vector control is used as a control method. If rotor time constant is exactly known so that the slip gain has an exact value, the measured d-axis voltage coincides with the d-axis reference voltage of (19). There needs to be a small amount of load for measuring the rotor time constant. Because, if the slip is too small, there is no difference between the d-axis voltage with correct slip gain and the d-axis voltage with incorrect slip gain. Loading the test motor is not so easy in practical application. To make the loaded condition, rotor inertia is used as load by accelerating the test motor for a short interval.

The measuring sequence is as follows. Before starting, the rated magnetizing current is applied, after which the torque reference is applied as a reference value. During acceleration the error voltage is measured from the measured d-axis voltage and the reference voltage. If the motor has achieved maximum speed, it then decelerates to stop. This sequence repeats a number of times. The rotor time constant can be identified by means of a model reference adaptive system (MRAS) [10]. But, in order to guarantee the convergence, the parameters of the MRAS block should be tuned properly. Another difficulty of adaptive methods is that convergent time is not predictable.

In this study, the direct measurement method is used. After applying the torque reference, the error voltage is measured during acceleration, after which the motor decelerates to stop. This acceleration and deceleration cycle repeats a number of times with the increment of slip gain in each cycle. Fig. 9 shows the measuring sequence mentioned above. Fig. 10 indicates the variation of d-axis voltage with the increment of slip constant. In Fig. 10, slip constant begins with 70[%] of initial value and increases until 130[%] with 5[%] increments. In each step, the error voltage is measured every 10[msec], then the average value is calculated from these data.

At the beginning the real d-axis voltage v_{ds}^{e} is lower than the reference d-axis voltage v_{ds}^{e*}. v_{ds}^{e} increases as the cycle advances and becomes close to reference d-axis voltage in point 'A' of Fig. 10, where slip constant is close to the exact value. Exact slip constant lies on the point where error becomes zero.

To find out the exact point where the real d-axis voltage equals the reference d-axis voltage, the measured points are approximated to 2-nd order curve by means of least square approximation in the Appendix. The equation to calculate slip constant using least square approximation is as follows.

$$V_{err} = a_0 + a_1 K_{s(n)} + a_2 K_{s(n)}^2 \quad ,n = 70,75...130 \tag{20}$$

$$a_0 = [l_0, l_1, ..., l_{13}][V_{err}(0), V_{err}(1), ..., V_{err}(13)]^T$$

$$a_1 = [m_0, m_1, ..., m_{13}][V_{err}(0), V_{err}(1), ..., V_{err}(13)]^T$$

$$a_2 = [n_0, n_1, ..., n_{13}][V_{err}(0), V_{err}(1), ..., V_{err}(13)]^T$$

Where $[l_0, l_1, ..., l_{13}]$, $[m_0, m_1, ..., m_{13}]$, $[n_0, n_1, ..., n_{13}]$ is pre-calculated constants using the method in reference [11], and $[V_{err}(0), V_{err}(1), ..., V_{err}(13)]$ is the voltage error measured quantity and reference value in each cycle.

The thick gray line of Fig. 11 is the approximation curve, and the exact value is on cross point of $\tilde{v}_{error}$ and x-axis. If slip constant K_s is measured, rotor time constant T_r can be calculated as follows.

$$T_r = 1 / K_s \tag{21}$$

In the proposed method, the time for measurement is always 13 cycles and the exact value can always be found if it lies within ±25[%] of the initial value. If the exact value is out of this range, the initial value is increased or decreased 50[%], after which the measuring sequence is tried again.

2.7 Experimental Result

Table 2 presents the measurement result for the motor of Table 1. The values acquired through the no-load test, the torque linearity test and the torque response comparison, etc. are summarized in Table 2 as reference values. It was hard to know the true values of motor parameters, so that the 'reference values' of Table 2 are only used as references. Most of the parameters measured by the proposed methods are close to the reference values. However, the difference of the magnetizing current is about 17[%] since the voltage margin for current control was not considered during the no-load. All the measuring sequences and corresponding waveforms are as shown in preceding sections.

These parameters were applied to actual drive systems that have the sensorless vector control function and the indirect vector control function. Performance of this system showed that the proposed measuring methods are satisfactory in real operating situations.

Table 2 Experimental results

Motor Parameters	Measured Values	Reference Values
R_s	0.522 [ohm]	0.518 [ohm]
σL_s	10.1 [mH]	11.5 [mH]
$I_{m(rms)}$	5.4 [A]	6.34 [A]
L_s	92.3 [mH]	91.2 [mH]
T_r	0.288 [sec]	0.264 [sec]

3. Conclusion

This paper presents the parameter measurement methods for the high performance drive of induction motors. The purpose of this study is to develop a parameter measurement method for the self-commissioning function of commercial inverters, which need to be reliable and could be done by inverter alone for the self-commissioning function. For this, an adequate method for each parameter is developed that considers the factors that affect the accuracy of parameter measurement. Then, the measuring methods and the conditions that are best fit to each parameter measurement procedure are proposed. All of the measurement procedures can be done without any auxiliary equipment, so that the proposed methods may be easily adopted as self-commissioning functions of commercial inverters.

Appendix

Curve fitting by least square approximation [11]:

If the n-th order approximation curve is (A1) and measured data is $\{[x_0, x_1, \ldots, x_M], [Y_0, Y_1, \ldots, Y_M]\}$,

$$y = a_0 + a_1 x + a_2 x^2 + \ldots + a_n x^n \quad \text{(A1)}$$

then error between the approximation curve and measured data is (A2) and sum of the error square is (A3).

$$err_i = Y_i - y_i = Y_i - a_0 - a_1 x_i - a_2 x_i^2 \ldots - a_n x_i^n \quad ,i = 0 \ldots M \quad \text{(A2)}$$

$$S = \sum_{i=1}^{M} err_i^2 = \sum_{i=1}^{M} (Y_i - a_0 - a_1 x_i \ldots - a_n x_i^n)^2 \quad \text{(A3)}$$

In an error minimum point, the following equations are satisfied.

$$\frac{\partial S}{\partial a_0} = \frac{\partial S}{\partial a_1} = \ldots = \frac{\partial S}{\partial a_n} = 0 \quad \text{(A4)}$$

Simultaneous equations (A5) can be obtained from (A4).

$$\begin{bmatrix} M & \sum x_i & \ldots & \sum x_i^n \\ \sum x_i & \sum x_i^2 & \ldots & \sum x_i^{n+1} \\ \vdots & \vdots & & \vdots \\ \sum x_i^n & \sum x_i^{n+1} & \ldots & \sum x_i^{2n} \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ \vdots \\ a_n \end{bmatrix} = \begin{bmatrix} \sum Y_i \\ \sum x_i Y_i \\ \vdots \\ \sum x_i^n Y_i \end{bmatrix} \quad \text{(A5)}$$

By solving (A5), the coefficient for (A1) can be obtained.

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