Electrical Parameter Estimation Procedures for Induction Machines during Selfcomissioning Phase

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Abstract— **This paper presents new procedures for electrical parameter estimation of induction motors based on its continuous-time model and by application of the recursive least-squares algorithm (RLS). The use of the continuous time model requires the calculation of the derivatives of current and voltage measured signals, which are determined using polynomial interpolation (spline-type) techniques. Three different linear regression forms are presented. The first one enables the simultaneous estimation of all the parameters of the model. The others are simplified models that are obtained from a priori knowledge of some parameters. For each procedure, all the electrical parameters are obtained considering the equivalent machine concept that will be defined. The estimation procedures are applied in the self-comissioning phase, prior to the machine start-up. In order to avoid any machine intervention, only no-torque signals are used in the standstill tests. The conditions to produce these no-torque voltages with the PWM inverter, avoiding ripple torque generation, are analyzed. The paper includes experimental results, which confirm the good performance of the proposed procedures.**

Keywords— **AC machines, adjustable speed drives, estimation techniques, field oriented control, induction motors, vector control.**

I. Introduction

Nowadays, higher quality industrial standards impose improved control performance that led to the development of advanced control schemes. The induction machine is one of the most widely used electrical apparatus in these modern power systems. In contrast with DC drives, AC motors present advantages such as robust construction and low cost. Nevertheless, it requires a complex control structure. While the independent control of torque and flux is inherent to DC machines design, in AC drives the torque and flux decoupling is usually obtained by two axis theory and field orientation concepts [1]. The knowledge of the induction motor parameters became an important step to the design and implementation of oriented field techniques and observers and controllers that compose these control systems.

The traditional measurement methods of the electrical parameters, the classical no-load and locked-rotor tests, are performed off-line and require extensive tests of the machine. In addition, these tests are performed under operating conditions that, usually, do not occur during vector control operation. For example, the locked-rotor test is usually performed with flux levels significantly lower than those used in normal operating conditions.

Many papers have been published on induction machine parameter estimation [2][3][4][5][6]. These researches are, normally, concerned on two different problems. The first is the parameter estimation that occurs before the motor start-up and is used for initial tuning of the control system, called self-commissioning phase [7]. A different problem concerns the continuous self-tuning during normal operation of the drive.

This paper deals with automatic identification procedures that enables the determination of all electrical parameters during the self-comissioning phase. To do this automatically, the induction machine should be fed directly by a PWM inverter. This drive is used to feed the machine with voltage signals that do not produce eletromagnetic torque. The best conditions to obtain these signals without ripple torque generation depends on the machine connection and will be investigated in this work.

The parameter estimation is performed by using the induction machine voltage-current continuous model and the recursive least squares algorithm [8]. Firstly, the complete model is used to estimate all the parameters simultaneously. Two others simplified models are deduced considering some a priori knowledge of the parameters. In all procedures the calculation of current and voltage derivatives of the measured signals are performed using a polynomial interpolation technique [9][10]. Experimental results were obtained under two different flux levels and demonstrated the good performance of the proposed estimation procedures.

II. Induction Machine Model

Usually, for estimation purposes, the electrical model of the machine is described into a stator reference frame. So, making the usual approximations, the model is defined by the following equations [11]:

$$\begin{aligned} \overrightarrow{v_s} &= R_s \overrightarrow{i_s} + L_s \frac{d}{dt}\overrightarrow{i_s} + L_m \frac{d}{dt}\overrightarrow{i_r} \\ 0 &= R_r \overrightarrow{i_r} + L_r \left(\frac{d}{dt}\overrightarrow{i_r} - jw_r \overrightarrow{i_r} \right) \\ &\quad + L_m \left(\frac{d}{dt}\overrightarrow{i_s} - jw_r \overrightarrow{i_s} \right) \end{aligned} \tag{1}$$

where subscripts s, r and m refer to the stator, rotor and mutual circuits, i and v are the current and voltage vector respectively, R is resistance, L is inductance and w_r is the rotor speed. From expression (1), the rotor current vector may be eliminated and the voltage-current model is obtained as:

$$\frac{d^2}{dt^2}\overrightarrow{i_s} = -A_1\frac{d}{dt}\overrightarrow{i_s} - A_0\overrightarrow{i_s} + B_1\frac{d}{dt}\overrightarrow{v_s} + B_0\overrightarrow{v_s} \quad (2)$$

with:

$$A_0 = R_s\frac{a}{\tau_r} - jaR_s w_r$$

$$A_1 = bR_r + aR_s - jw_r$$

$$B_0 = \frac{a}{\tau_r} - jaw_r$$

$$B_1 = a$$

where a and b are defined as $a = 1/\sigma L_s, b = 1/\sigma L_r$ with $\tau_r = L_r/R_r$ and $\sigma = \left(L_s L_r - L_m^2\right)/L_s L_r$.

The electromagnetic torque, invariant with the reference frame, is given by:

$$Te = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r}\,\mathrm{Im}\left(\overrightarrow{i_s}\ \overrightarrow{i_r^*}\right) \quad (3)$$

where P is the number of poles, Im indicates the complex imaginary part and $\overrightarrow{i_r^*}$ is the complex conjugate of rotor current vector.

A. Equivalent Machines

The determination of all the physical parameters present in equations (1) is not possible from the coefficients estimated with equation (2). The model is over-parameterized. Nevertheless, it is possible to overcome this drawback by defining electrical equivalent machines. Multiplying both sides of equation (1) by an arbitrary constant "m" and replacing the rotor current by $i_r' = i_r/m$, equation (4) is obtained.

$$\begin{aligned} \overrightarrow{v_s} &= R_s\overrightarrow{i_s} + L_s\frac{d}{dt}\overrightarrow{i_s} + mL_m\frac{d}{dt}\overrightarrow{i_r}' \quad (4)\\ 0 &= m^2R_r\overrightarrow{i_r}' + m^2L_r\left(\frac{d}{dt}\overrightarrow{i_r}' - jw_r\overrightarrow{i_r}'\right)\\ &\quad + mL_m\left(\frac{d}{dt}\overrightarrow{i_s} - jw_r\overrightarrow{i_s}\right) \end{aligned}$$

The expressions given by (4) define exactly the same electrical machine described by (1) for any value of m, $m \neq 0$ and $m \neq \infty$ [11]. The stator voltage and current, the stator parameters, the electromagnetic torque and rotor losses are invariant to the transformation of the rotor variables. So, a machine with parameters R_s, L_s, R_r, $L_r$and L_m is electrically equivalent to the machine with parameters R_s, L_s, m^2R_r, m^2L_r, and mL_m.

The choice of m corresponds exactly to use an arbitrary value for the relation between rotor and stator leakage inductances like in the classical tests, i.e. the standard no-load and the locked-rotor tests. For example, if m is set equal to $\sqrt{L_s/L_r}$, the stator self-inductance will be equal to the rotor self-inductance and $k = 1$. A relationship between L_{ls} and L_{lr} must be assumed. When design details are not available the ratio k must be assumed according to the design type [12]:

- $k = 1$ design A, D and wound rotor,
- $k = 0{,}67$ design B,
- $k = 0{,}43$ design C.

In the estimation process to be presented the following procedure will be adopted: Firstly, the k ratio is assumed equal to one that means $L_s = L_r$. With this assumption, all the physical parameters can be calculated from the estimated coefficients of equation (1). The parameters of design B and C motors can be found by an appropriate transformation. The value of m for this transformation is given by relation (5):

$$m = \frac{(k-1)\,L_m^1\sqrt{\left[(k-1)\,L_m^1\right]^2 + 4k\,(L_r^1)^2}}{2kL_r^1} \quad (5)$$

The electrical parameters are then determined by:

$$L_r^k = m^2L_r^k,\ L_m^k = L_m^1, R_r^k = m^2R_r^1 \quad (6)$$

In these equations the superscript means the actual value of k.

III. No-torque voltage signals

The machine model at standstill is obtained by making $w_r = 0$ at (2) that implies in the natural decoupling of the d and q circuits. By inspection of expression (3), the sufficient conditions to achieve zero torque are $v_{ds} = 0$ for any v_{qs}, or $v_{qs} = 0$ for any v_{ds}. In this section, the synthesis of such no-torque signals with a three-phase PWM voltage inverter is analyzed. Figure (1) shows a simplified representation of the inverter feeding a wye connected load. Table I shows the command signals, output phase voltages and the respective d,q components of each voltage vector and these voltage vectors are presented at figure (2).

From table(I), an expression relating the command signals pulse widths $\tau_1(k)$, $\tau_2(k)$, $\tau_3(k)$ and the mean value of the phase voltages $\bar{v}_{as}(k)$, $\bar{v}_{bs}(k)$, $\bar{v}_{cs}(k)$, during the kth PWM period is found as:

$$\begin{bmatrix} \bar{v}_{as}(k) \\ \bar{v}_{bs}(k) \\ \bar{v}_{cs}(k) \end{bmatrix} = \frac{E}{3T}\begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix}\begin{bmatrix} \tau_1(k) \\ \tau_2(k) \\ \tau_3(k) \end{bmatrix} \quad (7)$$

In all regular PWM methods, the reference voltage vector is sampled at each modulation period, T, and the

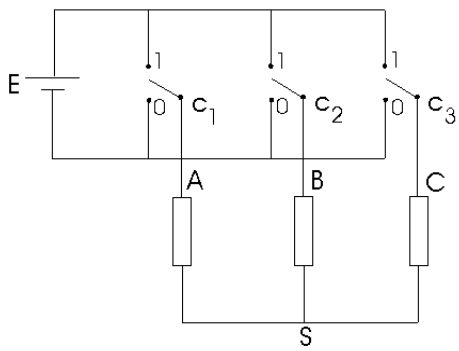

Fig. 1. PWM voltage converter

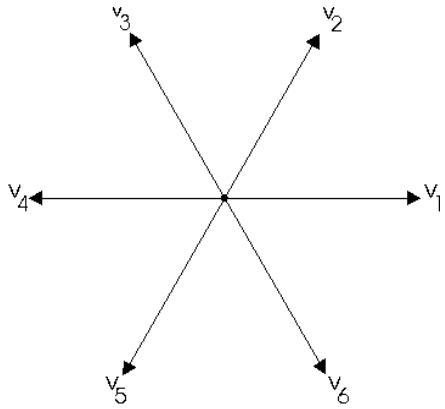

Fig. 2. Inverter voltage vectors

pulse widths are then calculated to make the mean voltage vector at the inverter output equal to this reference value. Computing the mean voltage vector $\bar{v}_{dqs}\ (k) = \frac{2}{3}\left(\bar{v}_{as}\ (k) + a\ \bar{v}_{bs}\ (k) + a^2\ \bar{v}_{cs}\ (k)\right)$ from equation (7), we obtain:

$$\tau_{dqs}\,(k) = \frac{T}{E}\ \bar{v}_{dqs}\ (k) \tag{8}$$

where $\tau_{dqs}\,(k) = \frac{2}{3}\left(\tau_1\,(k) + a\tau_2\,(k) + a^2\tau_3\,(k)\right)$ is defined as the pulse width command vector. While the voltage vector has null zero sequence component, the pulse width command vector must have:

$$\tau_0\,(k) = \frac{1}{3}\left(\tau_1\,(k) + \tau_2\,(k) + \tau_3\,(k)\right) \neq 0 \tag{9}$$

Applying the inverse transformation to the pulse width command vector, we finally obtain:

$$\begin{cases} \tau_1\,(k) = \frac{T}{E} v_{as}\,(k) + \tau_0 \\ \tau_2\,(k) = \frac{T}{E} v_{bs}\,(k) + \tau_0 \\ \tau_3\,(k) = \frac{T}{E} v_{cs}\,(k) + \tau_0 \end{cases} \tag{10}$$

where $0 \leq \tau_i\,(k) \leq T, \quad i = 1, 2, 3, \ldots$

This expression is general, valid for all regular sampled PWM methods. A particular PWM method can be obtained by the appropriate choice of $\tau_0\,(k)$.

TABLE I

COMMAND SIGNALS, OUTPUT VOLTAGES AND VECTORS IN PWM INVERTERS CONNECTED TO Y LOAD

c1	*c2*	*c3*	v_{as}	v_{bs}	v_{cs}	v_{ds}	v_{qs}	*vec*
0	0	0	0	0	0	0	0	$\overrightarrow{V_0}$
0	0	1	$-\frac{E}{3}$	$-\frac{E}{3}$	$\frac{2E}{3}$	$-\frac{E}{3}$	$-\frac{\sqrt{3}E}{3}$	$\overrightarrow{V_5}$
0	1	0	$-\frac{E}{3}$	$\frac{2E}{3}$	$-\frac{E}{3}$	$-\frac{E}{3}$	$\frac{\sqrt{3}E}{3}$	$\overrightarrow{V_3}$
0	1	1	$-\frac{2E}{3}$	$\frac{E}{3}$	$\frac{E}{3}$	$-\frac{2E}{3}$	0	$\overrightarrow{V_4}$
1	0	0	$\frac{2E}{3}$	$-\frac{E}{3}$	$-\frac{E}{3}$	$\frac{2E}{3}$	0	$\overrightarrow{V_1}$
1	0	1	$\frac{E}{3}$	$-\frac{2E}{3}$	$\frac{E}{3}$	$\frac{E}{3}$	$-\frac{\sqrt{3}E}{3}$	$\overrightarrow{V_6}$
1	1	0	$\frac{E}{3}$	$\frac{E}{3}$	$-\frac{2E}{3}$	$\frac{E}{3}$	$\frac{\sqrt{3}E}{3}$	$\overrightarrow{V_2}$
1	1	1	0	0	0	0	0	$\overrightarrow{V_7}$

We can now conclude our study of no-torque voltage signals. From table (I), the condition $v_{ds} = 0$ is fulfilled only by the application of zero voltage vectors $\overrightarrow{V}_0$ and $\overrightarrow{V}_7$. So, this condition can not be achieved with a wye connected charge in an instantaneous sense. Nevertheless, this condition can be accomplished in a mean sense by making:

$$\begin{aligned} \bar{v}_{ds}\ (k) &= 0\ and\ \bar{v}_{qs}\ (k) \neq 0 \\ &\Rightarrow \begin{cases} \bar{v}_{as}\ (k) = 0 \\ \bar{v}_{bs}\ (k) = -\ \bar{v}_{cs}\ (k) = \bar{v}\ (k) \end{cases} \end{aligned} \tag{11}$$

in a regular sampled PWM method as described by (10). $\bar{v}\ (k)$ is any desired voltage waveform.

The output mean torque will be equal to zero but some torque oscillations will exist. From (I), the condition $v_{qs} = 0$ is fulfilled by the application of the vectors $\overrightarrow{V}_0, \overrightarrow{V}_1, \overrightarrow{V}_4$ and $\overrightarrow{V}_7$. The condition $v_{qs} = 0$ implies:

$$\begin{aligned} \bar{v}_{qs}\ (k) &= 0\ and\ \bar{v}_{ds}\ (k) \neq 0 \\ &\Rightarrow \begin{cases} \bar{v}_{as}\ (k) = \bar{v}\ (k) \\ \bar{v}_{bs}\ (k) = \bar{v}_{cs}\ (k) = \frac{1}{2}\ \bar{v}\ (k) \end{cases} \end{aligned} \tag{12}$$

By substituting these phase voltage values at expression (10) we obtain $\tau_2\,(k) = \tau_3\,(k)$, and for a centered pulse PWM this also means $c_2\,(k) = c_3\,(k)$. Looking for this condition at table (I) we observe that the only vectors that will be used by the PWM converter are exactly $\overrightarrow{V}_0, \overrightarrow{V}_1, \overrightarrow{V}_4$ and $\overrightarrow{V}_7$. So, the PWM modulator assures instantaneously the condition $v_{qs} = 0$ with a wye connected charge. The same analysis can be carried out to a delta connected load, showing that in this case the condition $v_{ds} = 0$ is assured instantaneously by the PWM converter. Figures (3) and (4) show no-torque voltages obtained with $v_{ds} = 0$ to both wye and delta connections of the motor and confirm the previous analysis. The results were obtained by simulation.

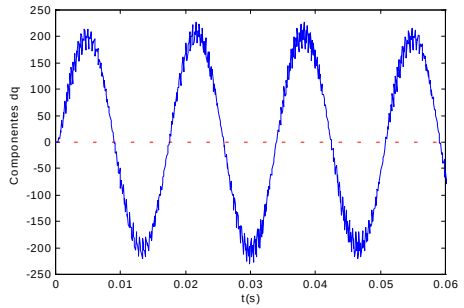

Fig. 3. Stator Voltages produced by PWM inverter for Delta load:(dashed) $v_{ds} = 0$ and (continuous)$v_{qs} = v(t) = 200\sin(377t)$

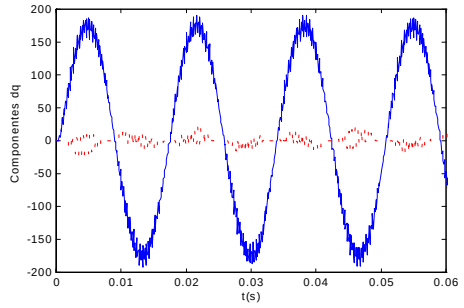

Fig. 4. Stator Voltages produced by PWM inverter for Wye load: (dashed) $v_{ds} = 0$ and (solid) $v_{qs} = v(t) = 200\sin(377t)$

IV. Estimation Procedures

The parameter estimation by recursive least square method is based on linear regression models [8], i.e.:

$$y(t|\Theta) = \Gamma(t)\Theta \tag{13}$$

where:

- $y(t|\Theta) \rightarrow$ output vector,
- $\Theta(t) \rightarrow$ regression matrix,
- $\Theta \rightarrow$ parameter vector.

The parameter vector to be estimated is defined as:

$$\hat{\Phi} = \arg\min_{\Theta} \; J \tag{14}$$

and is obtained from least-squares method which minimize a performance function of the form

$$J = \frac{1}{N}\sum_{i=1}^{N}\varepsilon^2 \tag{15}$$

where $\varepsilon = (y - \Gamma\hat{\Theta})$ is the prediction error.

A. Direct Estimation

The linear regression model of the induction machine at standstill is obtained by making $w_r = 0$ in (2). Since $v_{sd} = 0$, it can be written as:

$$\begin{aligned} y &= \ddot{i}_{qs} \\ \Gamma &= \left[\begin{array}{cccc} v_{qs} & \dot{v}_{qs} & -i_{qs} & -\dot{i}_{qs} \end{array}\right] \\ \hat{\theta} &= \left[\begin{array}{cccc} \hat{\theta}_1 & \hat{\theta}_2 & \hat{\theta}_3 & \hat{\theta}_4 \end{array}\right]^T = \\ & \left[\begin{array}{cccc} \frac{1}{\sigma\tau_r L_s} & \frac{1}{\sigma L_s} & \frac{1}{\sigma\tau_r\tau_s} & \frac{\tau_r+\tau_s}{\sigma\tau_r\tau_s} \end{array}\right]^T \end{aligned} \tag{16}$$

Considering the theory presented in the subsection (II-A) and taking the estimated coefficients from (16), all the parameters of the induction machine can be determined as:

$$\hat{\tau}_r^1 = \frac{\hat{\theta}_2}{\hat{\theta}_1} \tag{17}$$

$$\hat{R}_s = \frac{\hat{\theta}_3}{\hat{\theta}_1} \tag{18}$$

$$\hat{R}_r^1 = \frac{\hat{\theta}_4\hat{\theta}_1 - \hat{\theta}_3\hat{\theta}_2}{\hat{\theta}_2\hat{\theta}_1} \tag{19}$$

$$\hat{L}_s = \hat{L}_r^1 = \frac{\hat{\theta}_4\hat{\theta}_1 - \hat{\theta}_3\hat{\theta}_2}{\hat{\theta}_1^2} \tag{20}$$

$$\hat{L}_m^1 = \sqrt{\hat{L}_s\left(\hat{L}_s - \frac{1}{\hat{\theta}_2}\right)} \tag{21}$$

Applying (5) and (6) the values can be adapted to an especific design class of the induction machine.

B. Estimation with R_s known

If $v_{qs}(t)$ is a step voltage, the stator resistance may be determined from steady state values as:

$$\hat{R}_s = \left.\frac{v_{qs}(t)}{i_{qs}(t)}\right|_{t\rightarrow\infty} \tag{22}$$

The linear regression model (16) may be simplified to:

$$\begin{aligned} y &= \ddot{i}_{qs} \\ \Gamma &= \left[\begin{array}{ccc} \left(\dot{v}_{qs} - \hat{R}_s i_{qs}\right) & \left(v_{qs} - \hat{R}_s i_{qs}\right) & i_{qs} \end{array}\right] \\ \hat{\theta} &= \left[\begin{array}{ccc} \hat{\theta}_1 & \hat{\theta}_2 & \hat{\theta}_3 \end{array}\right]^T = \left[\begin{array}{ccc} \frac{1}{\sigma L_s} & \frac{1}{\sigma\tau_r\tau_s} & \frac{R_r}{\sigma L_s} \end{array}\right]^T \end{aligned} \tag{23}$$

Similarly, from (23) and item (II-A), it is possible to recover the remaining parameters of the machine. So,

$$\hat{\tau}_r^1 = \frac{\hat{\theta}_1}{\hat{\theta}_2} \tag{24}$$

$$\hat{R}_r^1 = \frac{\hat{\theta}_3}{\hat{\theta}_1} \tag{25}$$

$$\hat{L}_s = \hat{L}_r^1 = \frac{\hat{\theta}_3}{\hat{\theta}_2} \tag{26}$$

$$\hat{L}_m^1 = \sqrt{\hat{L}_s\left(\hat{L}_s - \frac{1}{\hat{\theta}_1}\right)} \tag{27}$$

C. Sequential Estimation

With R_s obtained from (22), it is possible to estimate the remaining parameters with a two steps procedure, avoiding the calculation of the current second derivative. The resultant models are:

$$\begin{aligned} y &= \dot{i}_{qs} \\ \Gamma &= \left[\begin{array}{cc} \left(v_{qs} - \hat{R}_s i_{qs}\right) & 1 \end{array} \right] \\ \hat{\theta} &= \left[\begin{array}{cc} \hat{\theta}_1 & \hat{\theta}_2 \end{array} \right]^T = \left[\begin{array}{cc} a & cR_r i_{qr} \end{array} \right]^T \end{aligned} \tag{28}$$

and,

$$\begin{aligned} y &= \left[\left(\dot{i}_{qs} - \hat{a}\dot{v}_{qs} + \hat{a}\hat{R}_s i_{qs}\right)\right] \\ \Gamma &= \left[\begin{array}{cc} \left(\hat{a}v_{qs} - \hat{a}\hat{R}_s i_{qs}\right) & -\hat{a}i_{qs} \end{array} \right] \\ \hat{\theta} &= \left[\begin{array}{cc} \hat{\theta}_3 & \hat{\theta}_4 \end{array} \right]^T = \left[\begin{array}{cc} \frac{1}{\tau_r} & R_r \end{array} \right]^T \end{aligned} \tag{29}$$

V. Derivative Estimation Procedure

Usually, the derivatives are estimated using filters or classical differential approaches. This work uses a polynomial interpolation method [10]. This method fits a polynomial to the data windows of m points using standard least square techniques. Such data correspond to the time signals of which one should estimate successive derivatives. The independent variable of the fitted polynomial is time. After fitting the polynomial function, one can determine the derivatives.

Assuming that the derivatives of the signal y(t) are required at sample k, the following window of data is chosen:

$$\left[\begin{array}{ccc} y_{k-m+1} & \cdots & y_k \end{array} \right] \tag{30}$$

where subscripts indicate the samples. Subsequently a polynomial of the type:

$$g\left(t\right) = \chi_n t^n + \chi_{n-1} t^{n-1} + \cdots + \chi_1 t + \chi_0 \tag{31}$$

is fitted to the data window. Finally, the derivatives at point y(k) are determined by differentiation of this polynomial. This method was applied to derivative calculation of the voltages and currents.

VI. Experimental Results

The experimental set-up is composed of a delta connected cage induction motor, a three-phase IGBT inverter and a Pentium/300 microcomputer. The test voltages were produced by a regular PWM method with a switching frequency $fs = 10$kHz. The sampling frequency was also 10kHz. Two currents and two voltages were measured using 12 bits resolution.

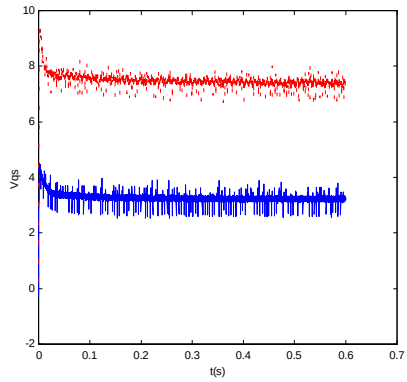

Fig. 5. input voltages (v_{qs}) at two different operating points.

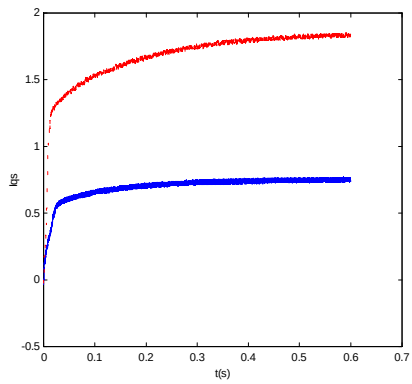

Fig. 6. Output current signals i_{qs} corresponding to $v_{qs} \approx 4V$ and $v_{qs} \approx 8V$.

A v_{sq} step voltage waveform were used in all tests. The step voltage level was varied to obtain two different flux conditions. For each operating point ten experiment realizations were made.Figures (5) and (6) show the two input voltage signals and the resultant i_{qs} motor currents. These signals were filtered with a 600Hz anti-aliasing filter.

Before calculating the derivatives, these signals were filtered again by a first order 10 Hz digital low-pass filter.

The estimated parameters presented on table (II) were obtained with $v_{sq} \approx$ 8V and represent the mean values obtained over all the 10 realizations. The first column presents the values obtained with the standard no-load and locked-rotor tests. The following columns present estimated parameters from the complete model (16) and the simplified models (23) and (28-29), respectively. The standard tests were obtained under the conditions specified by [12]. We can be observe the large differences between the standard values and estimated parameters.

The machine was simulated using the estimated parameters. Figures (7) and (8) show the results obtained with the regression model (16) considering the two different excitation levels. The estimated parameters used with v_{sq}= 4V are not shown on table (II). With v_{sq} = 8V, the machine was closer to the rated flux conditions while in the

TABLE II

ESTIMATED PARAMETERS WITH MACHINE AT STANDSTILL

Parameter	Std	Eq.(16)	Eq.(23)	Eq.(28 / 29)
$R_s\,(\Omega)$	3.41	3.898	3.898	3.898
$R_r\,(\Omega)$	3.64	2.379	2.292	3.571
$L_s\,(H)$	0.219	0.316	0.310	0.329
$L_r\,(H)$	0.224	0.328	0.321	0.341
$L_m\,(H)$	0.211	0.293	0.288	0.304
$\tau_r\,(s)$	0.062	0.138	0.140	0.095
$\sigma L_s\,(H)$	0.021	0.054	0.052	0.059

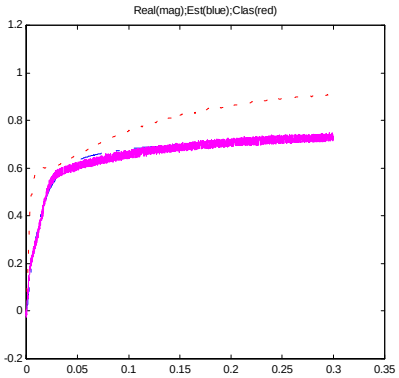

Fig. 7. Validation results ($v_{qs} \approx 4V$): (solid) actual current, (dashed) model using estimated parameters, (dotted) model using standard parameters

first it was underexcited.

Figure (9) shows the results from model (23) with v_{sq} = 8V. In all cases the transient and steady state responses obtained with the estimated parameters are closer to the actual current when compared with the simulation using the classical tests parameters.

Figure (10) presents the results obtained from models (28) and (29). In this case the actual and simulated responses with the estimated parameters are slightly different.

VII. CONCLUSIONS

The application of the continuous time model simplifies the recovering of the physical parameters of the induction motor. The presented procedures are adequate to the self-commissioning phase since no additional hardware or machine interventions are needed, enabling its inclusion into the drive system software.

The conditions to feed the machine without torque production were analyzed. It was demonstrated that the voltage vector component to be used is a function of the load connection. With this choice it is possible to minimize torque oscillations during standstill tests.

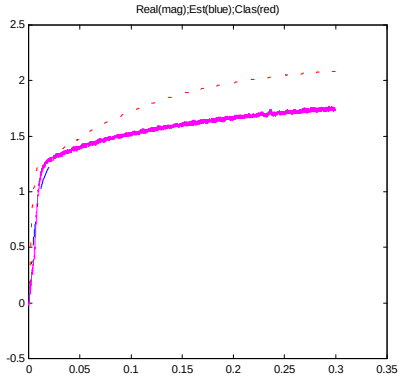

Fig. 8. Validation results ($v_{qs} \approx 8V$): (solid) actual current, (dashed) model using estimated parameters, (dotted) model using standard parameters.

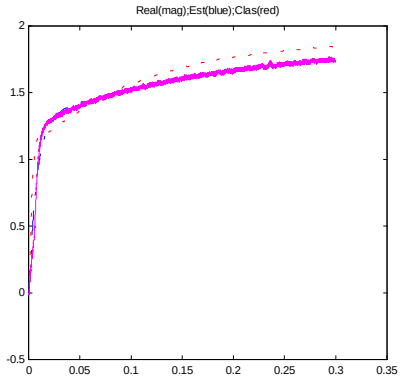

Fig. 9. Validation results ($v_{qs} \approx 8V$): (solid) actual current, (dashed) model using estimated parameters, (dotted) model using standard parameters.

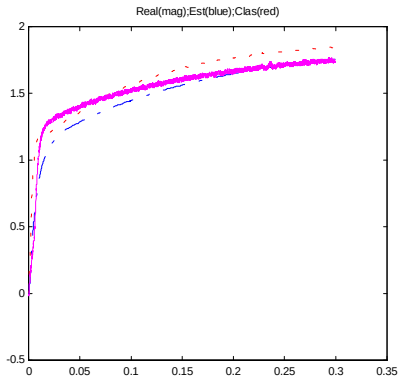

Fig. 10. Validation results ($v_{qs} \approx 8V$): (solid) actual current, (dashed) model using estimated parameters, (dotted) model using standard parameters

Through the definition of equivalent machines it was possible to recover all the machine parameters in all proposed procedures.

The validation results shown better results for models (16) and (23). The sequencial model was dependent on the quality of the a priori knowledge.

The proposed procedures are well suited to be used in a drive system with automatic self-comissioning.

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