

THE COMPLEX CONTROLLER APPLIED TO THE INDUCTION MOTOR CONTROL

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Abstract - This paper presents a design and tuning method of the complex gain controller, based on the three-phase induction motor mathematical model complex transfer function to be used in induction motor control when the machine operates at low speed which is a problem so far. The complex gain controller was applied in the direct rotor field orientation control method and also on the direct torque control method. It was designed and tuned using the frequency-response function of the closed loop system. The complex gain controller presents low complexity in the induction motor control implementation. Experimental results were carried out for the controller validation.

Keywords – Induction Motor Control, Complex gain controller, Complex Transfer Function, Field orientation, Direct torque control.

I. INTRODUCTION

The three-phase induction motor dynamics is traditionally represented by differential equations. The space-vector concept [1] is used in the mathematical representation of the induction machine state variables such as voltage, current, and flux.

The concept of complex transfer function derives from the application of the Laplace transform to the differential equations mentioned above in which the complex coefficients are in accordance with the spiral vector theory which has been presented by [2]. The complex transfer function concept of the three-phase induction motor mathematical model and the induction motor root locus were presented in [3]. Other procedures for modeling and simulating the three-phase induction motor dynamics using the complex transfer function concept are also presented in [4].

The induction machine high performance dynamics is achieved by the field orientation control (FOC) [5, 6]. The three-phase induction motor field orientation control using the complex transfer function concept to tune the PI controller by using the frequency-response function of the closed-loop complex transfer function of the controlled induction machine was presented in [7]. This strategy has satisfactory current response although stator currents had presented cross-coupling during the induction machine transients. An interesting solution was presented in [8] in which it was designed a stator-current controller using complex form. From this, the current controller structure employing single-complex zeros is synthesized with satisfactory high dynamic performance.

Some investigations by using deadbeat [9, 10], fuzzy [11], internal mode controller [12] and predictive controller [13] have been presented. These strategies have been presenting satisfactory current response although the complex transfer function of induction motor is not used and low-speed tests had not been shown.

An alternative for induction motor drive is the direct torque control (DTC), which consists of the direct control of the stator flux λ_1 and of the electromagnetic torque T_e . DTC controllers generate a stator voltage space vector that allows quick torque response with the smallest variation of the stator flux. The principles of the DTC using hysteresis controllers and variable switching frequency have been presented in [14] and [15]. It has disadvantages such as low speed operation [16].

To solve the variable switching frequency problem DTC strategies operating at constant switching frequency can be implemented by means of closed-loop schemes with PI controllers [17, 18], deadbeat [19, 20], sliding mode [21, 22], predictive [23] and neuro-fuzzy [24] controllers. The controllers calculate the required stator voltage vector, averaged over a sampling period. These strategies have satisfactory flux and torque response although the complex transfer function of induction motor is not used.

To overcome the low speed operation shortcomings, various approaches for FOC and DTC have been reported using indirect flux detection by on-line reactance measurement method [25], adaptive observer [26] and deadbeat control [27, 28]. These strategies aim the induction motor control operation at low speed. In this case, the complex transfer function and the complex gain controller are not used for such strategy.

This work proposes a design and tuning method for a complex gain control method, based on the three-phase induction motor mathematical model complex transfer function to be used in induction motor control when the machine operates at low speed which is a problem so far. The complex gain controller was applied on rotor field orientation control and on the direct torque control. It was designed and tuned by frequency-response function of the closed loop system. The complex gain controller presents a minor complexity for the induction motor control implementation when compared to the techniques mentioned above. Experimental results were carried out for the controller validation.

II. THE COMPLEX MATHEMATICAL MODEL OF THE THREE-PHASE INDUCTION MACHINE

The induction motor mathematical model is written in the synchronous reference frame (dq) and the state variables are stator current space vector \vec{i}_1 and stator flux space vector $\vec{\lambda}_1$

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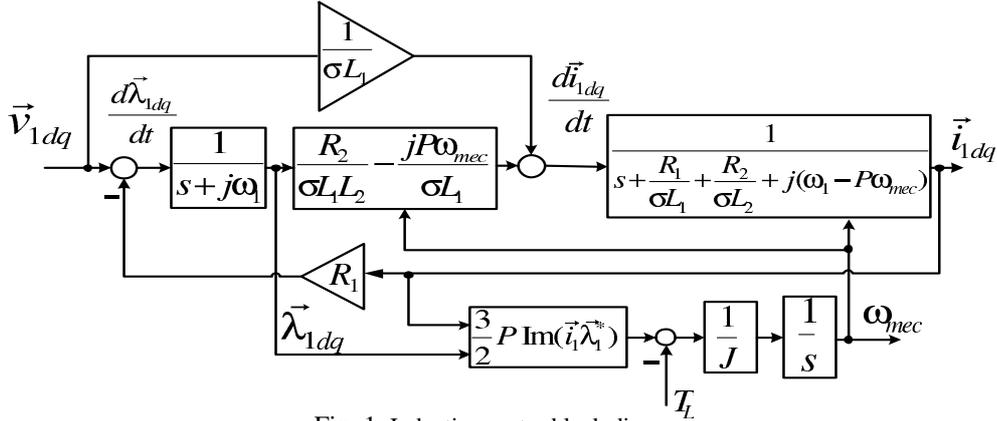


Fig. 1. Induction motor block diagram.

[29].

$$\begin{bmatrix} \dot{\vec{\lambda}}_{1dq} \\ \dot{\vec{i}}_{1dq} \end{bmatrix} = \begin{bmatrix} a_1 & a_2 \\ a_3 & a_4 \end{bmatrix} \begin{bmatrix} \vec{\lambda}_{1dq} \\ \vec{i}_{1dq} \end{bmatrix} + \begin{bmatrix} \vec{v}_{1dq} \\ \vec{v}_{1dq} / \sigma L_1 \end{bmatrix} \quad (1)$$

Where

$$a_1 = -j\omega_1 \quad (2)$$

$$a_2 = -R_1 \quad (3)$$

$$a_3 = \left(\frac{R_2}{\sigma L_1 L_2} - \frac{jP\omega_{mec}}{\sigma L_1} \right) \quad (4)$$

$$a_4 = - \left[\frac{R_1}{\sigma L_1} + \frac{R_2}{\sigma L_2} + j(\omega_1 - P\omega_{mec}) \right] \quad (5)$$

and σ is the leakage coefficient. The machine mechanical dynamics is given by

$$J \frac{d\omega_{mec}}{dt} = \frac{3}{2} P \text{Im}(\vec{i}_{1dq} \vec{\lambda}_{1dq}^*) - T_L \quad (6)$$

The subscripts 1, 2 and m represent the stator, rotor and magnetization parameters respectively, ω_1 is the synchronous speed, ω_{mec} is the machine mechanical speed, R_1 and R_2 are the per phase stator and rotor windings electrical resistance, respectively, L_1 , L_2 and L_m are the proper and mutual inductances of the stator and rotor windings, \vec{v} is the voltage vector, P is the machine number of pair of poles, J is the load and rotor inertia moment and T_L is the machine load torque.

In order to obtain the induction motor complex transfer function the Laplace transform is applied to (1) in accordance with the complex transfer function concept, it is assumed that the mechanical time constant of the motor is much larger than the transient electromagnetic time constants and the saturation effects is neglected. Thus, $\omega_{mec} = \text{constant}$ is a valid approximation [2, 8, 30]. The induction motor block diagram originated by use of (1) and (6) is shown in Figure 1.

When designing the DFOC or DTC control systems through the induction motor complex transfer function, \vec{v}_{1dq} is considered as the input and the \vec{i}_{1dq} is considered as the output. Therefore the induction machine complex transfer function

$H(s)$ is derived from the application of the Laplace transform to the equation (1) and it has the form given in 7.

$$H(s) = \frac{I_{1dq}}{V_{1dq}} = \frac{\left(\frac{s + j\omega_1}{\sigma L_1} \right) + a_3}{(s + j\omega_1)(s + a_4) + R_1 a_3} \quad (7)$$

Where $I_{1dq} = \mathcal{L}\{\vec{i}_{1dq}\}$ and $V_{1dq} = \mathcal{L}\{\vec{v}_{1dq}\}$.

III. DIRECT ROTOR FIELD ORIENTATION CONTROL

The direct rotor field orientation control (DFOC) allows to control the flux magnitude λ_2 and torque T_e through the stator currents [6] with the relationship given by

$$\lambda_2 = i_{1d} L_M \quad (8)$$

and

$$T_e = \frac{3L_M}{2L_2} P \lambda_2 i_{1q} \quad (9)$$

Using the complex notation in the direct rotor field orientation strategy where i_{1d} is the real component and i_{1q} is the imaginary component of the stator current. Thus, the traditional PI controller presented in [31] can be substituted by the complex gain $(a + jb)$. This controller generates the voltage reference by using the stator current space vector error $(\varepsilon_{i_{1d}} + j\varepsilon_{i_{1q}})$. The reference voltage space vector is given by

$$\vec{v}_{1dq_{ref}} = (a + jb) (\varepsilon_{i_{1d}} + j\varepsilon_{i_{1q}}) \quad (10)$$

which means

$$v_{1d_{ref}} = (a \varepsilon_{i_{1d}} - b \varepsilon_{i_{1q}}) \quad (11)$$

$$v_{1q_{ref}} = (b \varepsilon_{i_{1d}} + a \varepsilon_{i_{1q}}) \quad (12)$$

Where a is the real component of the complex gain, b is the imaginary component of the complex gain, $\varepsilon_{i_{1d}}$ is the error between $i_{1d_{ref}}$ and i_{1d} and $\varepsilon_{i_{1q}}$ is the error between $i_{1q_{ref}}$ and i_{1q} . The direct rotor field orientation strategy with the complex gain controller is shown in Figure 2.

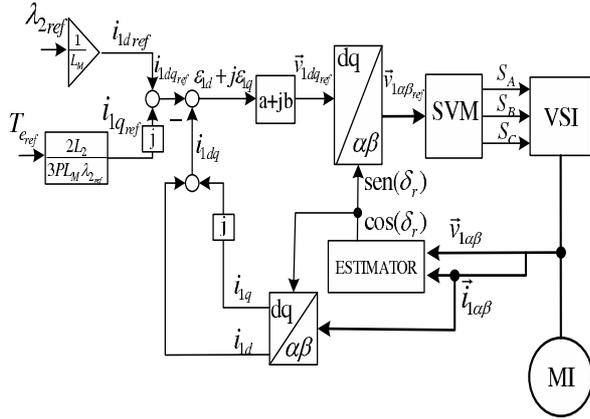


Fig. 2. DFOC strategy with complex gain controller.

The reference stator voltage vector \vec{v}_{1dqref} is transformed by using rotor flux angle δ_r to obtain the stator voltage in the stationary reference frame $\alpha\beta$ as it is shown in the next section.

A. Rotor Flux Estimation

The stator flux estimation is given by

$$\vec{\lambda}_{1\alpha\beta} = \int (\vec{v}_{1\alpha\beta} - R_1 \vec{i}_{1\alpha\beta}) dt \quad (13)$$

An interesting flux estimation at low speed using equation (13) is obtained by using the integration method given by [32]. The rotor flux estimation is done by

$$\vec{\lambda}_{2\alpha\beta} = \frac{L_2}{L_M} \vec{\lambda}_{1\alpha\beta} + \frac{L_M^2 - L_1 L_2}{L_M} \vec{i}_{1\alpha\beta} \quad (14)$$

Where the subscript $\alpha\beta$ is used to designate the stator stationary reference frame which is being used.

The rotor flux angle is estimated by using the trigonometric transfer function

$$\delta_r = \arctan \left(\frac{\lambda_{2\beta}}{\lambda_{2\alpha}} \right) \quad (15)$$

B. Design of the Complex Gain

The tuning operation of the complex gain controller requires the transfer function of the closed loop system, to obtain its frequency-response function. In the induction motor complex transfer function (7) \vec{v}_{1dq} is considered as the input and the \vec{i}_{1dq} is considered as the output.

The frequency-response function of the $H(s)$ given in (7) is shown in Figure 3. The low speed utilized in this paper is 125 rpm ($13rad/s$) that corresponds to 4.16Hz. As the variables at synchronous referential are constants the angle of the current \vec{i}_{1dq} is neglected.

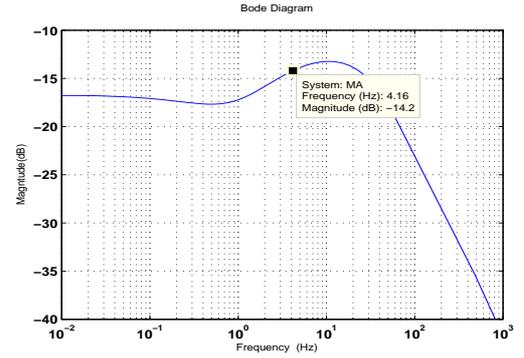


Fig. 3. Induction motor (7) frequency-response function.

In accordance with the control theory [33, 34] the control system block diagram used to obtain the closed loop transfer function is presented in Figure 4.

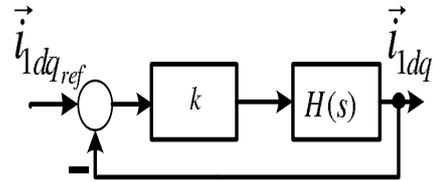


Fig. 4. System to design the complex gain.

Where $k = a + jb$. The transfer function of the system presented in Figure 4 is given by

$$C(s) = \frac{(a + jb)H(s)}{1 + (a + jb)H(s)} \quad (16)$$

For non-load induction motor it yields

$$C(s) = \frac{\left[\frac{(s + j\omega_1)}{\sigma L_1} + \frac{R_2}{\sigma L_1 L_2} - \frac{jP\omega_{mec}}{\sigma L_1} \right] (a + jb)}{m(s) - jP\omega_{mec} \frac{R_1}{\sigma L_1} + n(s) (a + jb)} \quad (17)$$

Where

$$m(s) = (s + j\omega_1) \left(s + \frac{R_1}{\sigma L_1} + \frac{R_2}{\sigma L_2} \right) + \frac{R_1 R_2}{\sigma L_1 L_2} \quad (18)$$

and

$$n(s) = \left[\frac{(s + j\omega_1)}{\sigma L_1} + \frac{R_2}{\sigma L_1 L_2} - \frac{jP\omega_{mec}}{\sigma L_1} \right]. \quad (19)$$

In the case of the 4.16Hz the complex gain to achieve 0 dB magnitude of $H(s)$ in the equation (17) is $75-25j$ as it can be shown in Figure 5.

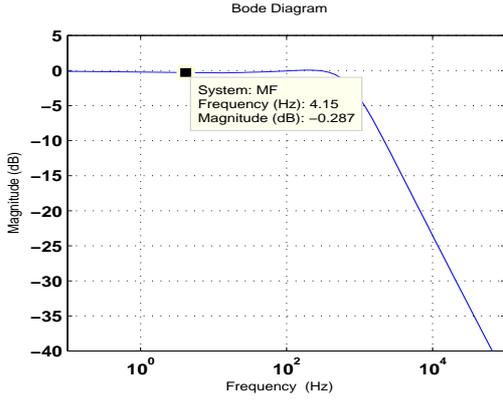


Fig. 5. Frequency-response function of the equation (17).

IV. DIRECT TORQUE CONTROL

By using stator field orientation in this present work, the torque and stator flux must become parts of a complex number, where the magnitude of the stator flux λ_1 is the real component and the torque T_e is the imaginary component. Hence, the reference signals and the error become a complex number. Thus, the traditional PI controller presented in [17] can be substituted by complex gain $(a + jb)$. This gain has the function to generate a voltage reference vector using the stator flux-torque vector error $(\varepsilon_\lambda + j\varepsilon_T)$. This way the stator voltage vector in this control strategy is given by

$$\vec{v}_{1dqref} = (\varepsilon_\lambda + j\varepsilon_T) (a + jb) \quad (20)$$

which means that the direct and quadrature axis of the voltage vector are

$$v_{1dref} = (-\varepsilon_T b + \varepsilon_\lambda a) \quad (21)$$

$$v_{1qref} = (\varepsilon_T a + \varepsilon_\lambda b) \quad (22)$$

Where a is the real component of complex gain, b is the imaginary component of complex gain, ε_λ is the flux error signal and ε_T is the torque error signal.

The block diagram of the strategy with the complex gain is presented in Figure 6.

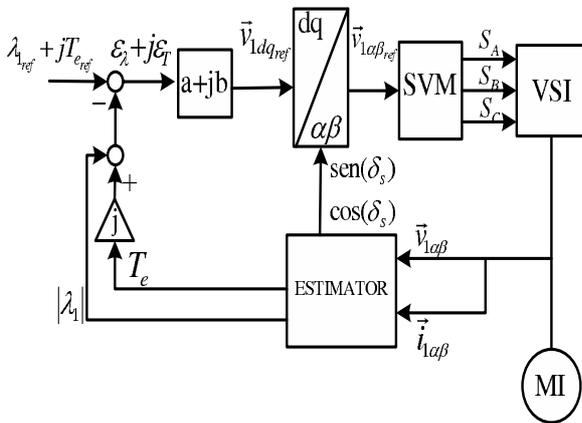


Fig. 6. DTC strategy with complex gain controller.

The reference stator voltage vector \vec{v}_{1dqref} is transformed by using stator flux angle δ_s to obtain the stator voltage at stationary reference frame $\alpha\beta$.

The stator flux estimation is obtained by using equation (13) in section III.A and its angle is estimated by using the trigonometric transfer function

$$\delta_s = \arctan \left(\frac{\lambda_{1\beta}}{\lambda_{1\alpha}} \right). \quad (23)$$

Design of the Complex Gain

The tuning operation for the complex gain determination requires the transfer function of the closed loop system to obtain its frequency-response function.

The frequency is chosen in accordance with the the induction motor desired speed operation to design the complex gain.

In accordance with the DTC control strategy the induction motor output has to be the stator flux magnitude λ_1 and the torque T_e . Therefore the $H(s)$ outputs have to become the stator flux magnitude λ_1 and the torque T_e according equation (7). The expression to obtain the stator flux by using the stator current i_{1d} is given by

$$\lambda_1 = \lambda_{1d} \cong G \sigma L_1 i_{1d} \quad (24)$$

and to obtain the electromagnetic torque in the dq reference frame one may use the expression:

$$T_e = \frac{3}{2} P \lambda_1 i_{1q} \quad (25)$$

As the stator flux magnitude λ_1 is assumed to be essentially constant, through of the equations (7), (24) and (25) the new transfer function is achieved with torque and flux as output and it is given by

$$\frac{X_{\lambda T}}{V_{1dq}} = H(s) \left(G \sigma L_1 + j P \frac{3}{2} \lambda_1 \right) \quad (26)$$

Where $X_{\lambda T} = \mathcal{L}\{\lambda_1 + jT_e\}$. The frequency-response function of equation (26) is presented in Figure 7 at frequency 4.16 Hz to tuning the complex gain.

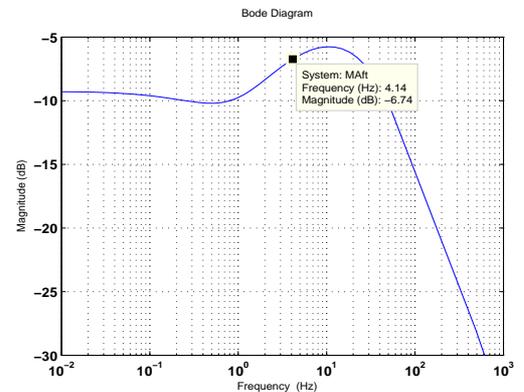


Fig. 7. Frequency-response function of the equation (26).

Then, from equations (20) and (26) one obtain the control system block diagram and it is presented in Figure 8.

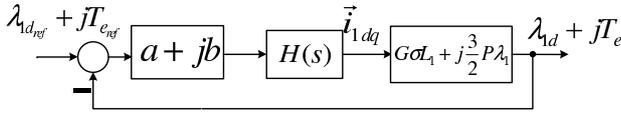


Fig. 8. System to design the complex gain.

The expression of the closed loop transfer function of the control system shown in Figure 8 is given by

$$\frac{X_{\lambda T}}{X_{\lambda T_{ref}}} = \frac{(a + jb) H(s) \left(G\sigma L_1 + jP \frac{3}{2} \lambda_1 \right)}{1 + (a + jb) H(s) \left(G\sigma L_1 + jP \frac{3}{2} \lambda_1 \right)} \quad (27)$$

Where $X_{\lambda T_{ref}} = \mathcal{L} \{ \lambda_{1ref} + jT_{e_{ref}} \}$. At the frequency of 4.16 Hz the complex gain was calculated by using simulations, considering slip approximately null and the 0 dB magnitude. Its value is $125 - j25$. The frequency-response function of equation (27) is shown in Figure 9 and its magnitude is near 0 dB.

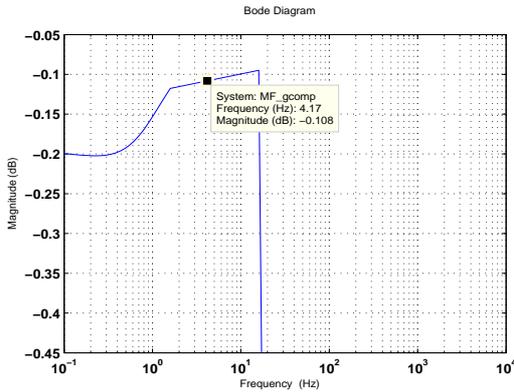


Fig. 9. Frequency-response function of the equation (27).

V. DIFFERENCE BETWEEN FORMULATIONS OF THE COMPLEX GAIN CONTROLLER AND OF THE PI CONTROLLER

The induction machine stator voltage was computed using the complex gain controller of the equation (10) for DFOC and of the equation (20) for DTC. In both cases the computed stator voltage v_{1dq} was the result of the product of two complex numbers: the complex error and the complex gain controller.

In the case of the PI controller use, in direct rotor field orientation control, the v_{1dq} is given by

$$v_{1d_{ref}} = k_{p1} \varepsilon_{1d} + \frac{k_{i1}}{s} \varepsilon_{1d} \quad (28)$$

$$v_{1q_{ref}} = k_{p2} \varepsilon_{1q} + \frac{k_{i2}}{s} \varepsilon_{1q} \quad (29)$$

for direct torque control the v_{1dq} is given by

$$v_{1d_{ref}} = k_{p1} \varepsilon_{\lambda} + \frac{k_{i1}}{s} \varepsilon_{\lambda} \quad (30)$$

$$v_{1q_{ref}} = k_{p2} \varepsilon_T + \frac{k_{i2}}{s} \varepsilon_T \quad (31)$$

Where k_{p1} , k_{p2} , k_{i1} and k_{i2} are the parameters of the PI controller. It can be seen in equations (28), (29), (30) and (31) that it is necessary one PI controller to compute each stator voltage component. Thus, the parameters of two PI controllers have to be tuned for induction motor control.

Comparing the equations (28) with (11), (30) with (21), (29) with (12) and (31) with (22) it can be seen that v_{1dq} was computed by a complex product between the complex error vector and the complex gain controller and in the case of the PI controller, each component of the v_{1dq} was computed by a product between a PI controller and the error of each component. Thus, using the complex gain controller the parameters number of the controller was reduced from four to two. Obviously this reduction decreases the difficult to tune the parameters of the controller.

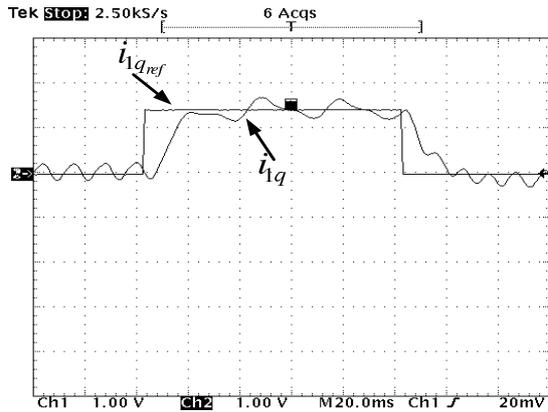
VI. EXPERIMENTAL RESULTS

The DFOC and DTC strategies were implemented using a Texas Instruments DSP TMS320F2812 platform. The system consists of a three-phase voltage source inverter with insulated-gate bipolar transistors (IGBTs) and the three-phase induction motor with electrical and mechanical characteristics are shown in the appendix. The stator voltage commands were modulated by using symmetrical space vector PWM with switching frequency equal to 2.5 kHz. The DC bus voltage of the inverter is 226 V and the stator voltages and currents was sampled in the frequency of 2.5 kHz. A conventional PI controller generates a torque reference by using the speed error. The flux and torque estimation, and the flux-torque complex regulator and speed controller have the same sampling frequency of 2.5 kHz. The encoder resolution is 1500 pulses per revolution.

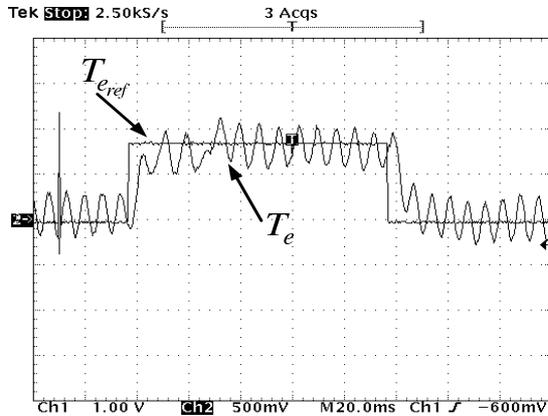
Three no-load induction motor tests were made. The first one was the response to a torque step of 12.2 Nm which is shown in Figure 10. The $i_{1q_{ref}}$ is calculated by equation (25). The response of the DTC with complex controller presented a slightly better performance during the transient and steady state when such response is compared with the response to the DTC with PI controller and DFOC with complex gain controller although the response of DFOC has less oscillations. These oscillations occur due to the natural lack of accuracy in the measurements of currents and voltages and induction motor parameters variations [35].

In the second test the speed varies in forward and reversal operation and the results are presented in Figure 11. The speed changes from 13 rad/s to -13 rad/s in 1 s and the complex gain is not changed during the test. This result confirms the satisfactory performance and the robustness of the controller due to the fact that the speed reaches the reference in several conditions. The responses of the DTC and DFOC with complex gain controller and of the DTC with PI controller have similar performance in the transient and in the steady

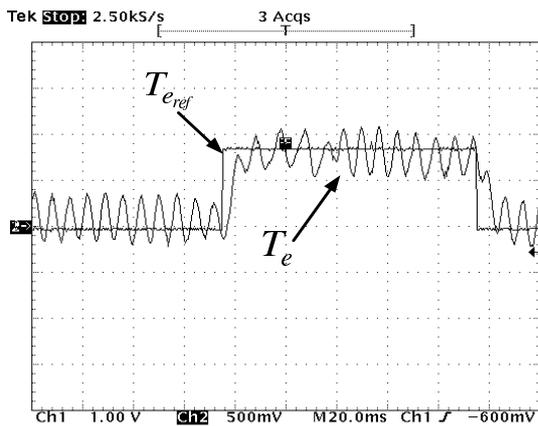
state. The small error occurs due to the natural lack of accuracy in the speed measurement.



(a) DFOC - i_{iqref} step function (4.2 A/div).



(b) DTC - T_{e_ref} step input function with complex controller (9 Nm/div).

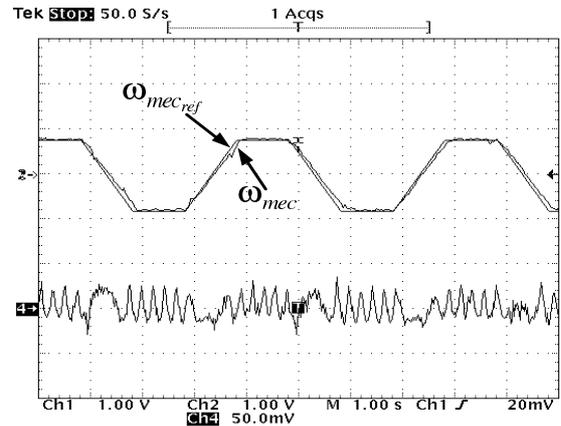


(c) DTC - T_{e_ref} step input function with PI controller (9 Nm/div).

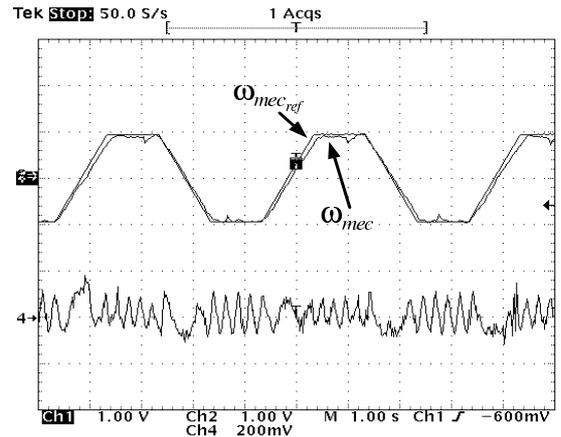
Fig. 10. Responses to step torque operation.

Figure 12 presents the speed response when the speed varies from 6.28 rad/s to -6.28 rad/s. The response of the DTC control with complex gain controller presented a slightly better performance in steady state when such response is compared with the response of DTC with PI controller and DFOC with

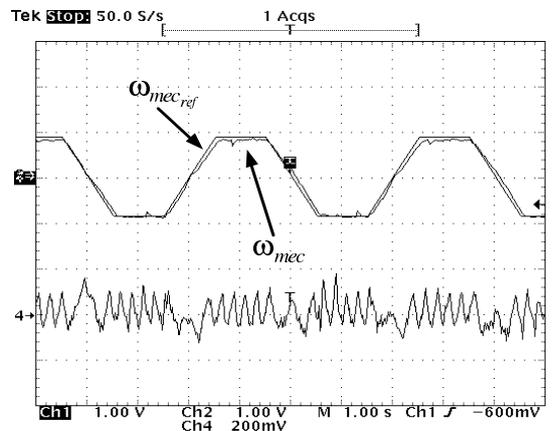
complex gain controller. The result confirms the satisfactory performance and the robustness of the complex controller again due to the fact that the speed reaches the reference value although the complex gain is designed for induction motor operation at 4.16 Hz.



(a) DFOC - speed (15.3 rad/s.div).



(b) DTC - speed with complex controller (13 rad/s.div).

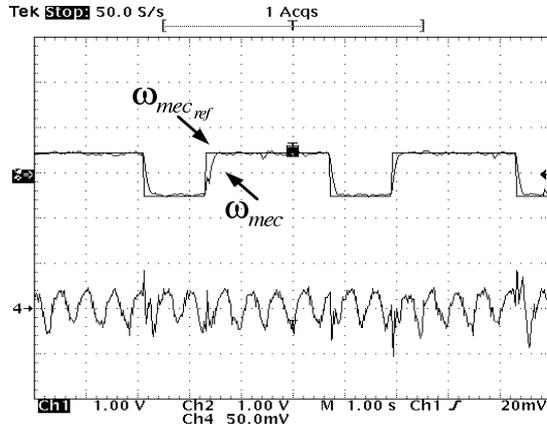


(c) DTC - speed with PI controller (13 rad/s.div).

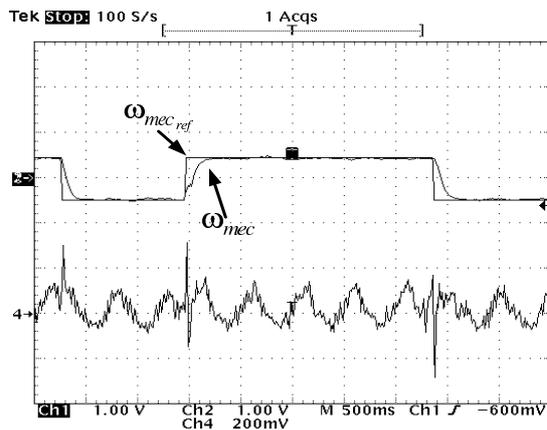
Fig. 11. Speed forward and reversal operation and current of phase a (10 A/div).

VII. CONCLUSION

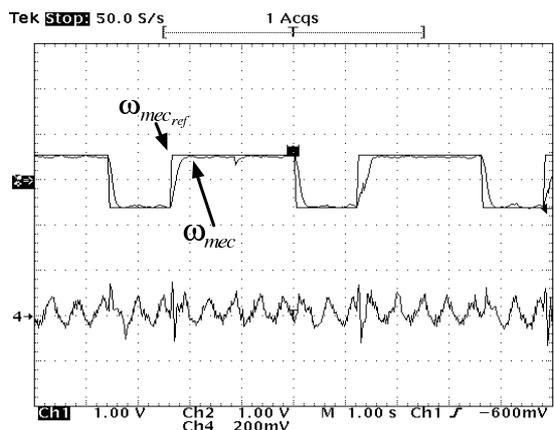
The concept of complex transfer function associated with the complex gain controller allows the design and tuning the complex controller by using the closed loop system frequency response function of the controlled induction motor.



(a) DFOC - speed (12.57 rad/s.div).



(b) DTC - speed with complex controller (12.57 rad/s.div).



(c) DTC - speed with PI controller (12.57 rad/s.div).

Fig. 12. Speed response to step operation and current of phase a (6 A/div).

The experimental results presented the satisfactory performance and the robustness of the complex controller due

to the fact that the speed reaches the reference value in several conditions although the complex gain had been designed for induction motor operation at 4.16 Hz. The complex gain controller has presented a slightly better performance in steady state when such response was compared with the response of the DFOC and to the DTC with PI [17] although the DFOC had less oscillation in the torque response [30]. Thus, the design of complex gain has an acceptable robustness although a detailed analysis considering parameters variations and other several speed operations has to be done. Due to the variable speed operation maybe it will be necessary to construct a table with complex gains designed for each speed desired speed or to an each speed range.

The complex gain controller overcomes the low speed operation shortcomings in the rotor flux orientation control and in the direct torque control with a minor complexity solution when compared with the solutions presented by indirect flux detection by on-line reactance measurement method [25], adaptive observer [26] or deadbeat [27]. Thus, the complex vector notation and the complex gain controller become an interesting tool for implementation induction motor drives.

APPENDIX

Three-phase induction motor variables and parameters: $P_N = 2.3kW$; $V_N = 220 V$; $Poles = 4$ $R_1 = 2.229 \Omega$; $R_2 = 1.522 \Omega$; $L_m = 0.238485 H$; $L_1 = 0.2470 H$; $L_2 = 0.2497 H$; $J = 0.0067 Kgm^2$.

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