Robust Nonlinear Control for High Performance Induction Motor Drives Based on Adaptive Disturbance Compensation

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***Abstract*—This paper proposes a new active disturbance rejection control (ADRC) for induction motor (IM) drives. In particular, differently from a classic Extended State Observer (ESO), here, a high-gain unknown input observer is used, with a driving term that is a function of the tracking error. This approach allows total robustness, as explained in the paper. The proposed control technique has been experimentally verified on a suitably devised test set-up.**

***Index Terms*—Induction motor, disturbance compensation, unknown input observer.**

I. Introduction

This paper deals with the theoretical development and experimental application of the Active Disturbance Rejection Control (ADRC) to Induction Motor (IM) drives. The industrial standard for high-performance control of IM drives is the so-called field-oriented control (FOC), proposed and rearranged in various forms in the scientific literature [1], [2]. FOC guarantees optimal dynamic performance and full decoupling of the speed and flux loops in constant flux operation. Whenever flux variation is required, for example, when an electrical losses minimization technique (ELMT) is integrated into the drive control, FOC does not permit a full decoupling of the speed and flux loops, with consequent reduction of the dynamic performance [3]. To overcome this limit of FOC, nonlinear control techniques must be adopted. Control theory offers manifold nonlinear control methodologies to deal with the significant nonlinearities of IM drives, among which the most common is feedback linearization (FL) control [4], [3], [5], [6]. Although FL is, in principle, applicable, once the dynamic model of the plant is known, it still presents some problems to be faced up, either related to rotating or linear induction motors. A typical problem arises when there is no precise knowledge of the mathematical model or when there is uncertainty about parameters or unmodeled dynamics. To address this issue, in [7], the linear controller has been substituted with a suitable controller designed to be robust to the variations of the main parameters of the induction motor, like stator and rotor resistances, and the three-phase magnetizing inductance. In some works, more complex models have been adopted [8], [9]. Alternatively, the FL has been integrated with a suitable online parameter estimation method [10], [11], [12]. A further approach, adopted in [13], [14]-[15], [16] proposes the adoption of the ADRC. In [17], [18], the problem of three-phase asynchronous motor regulating energy saving is analyzed using ADRC. In the electrical drives field, ADRC has been applied several times to control Permanent Magnet Synchronous Motors (PMSM) [19], [20], [21], [22], [23]. In [19], a novel parallel structure to improve dynamic responses, which replaces the traditional cascade structure of position and speed loops has been proposed, [20] proposes an enhanced linear active disturbance rejection controller-based rotor position sensorless field-oriented control scheme, [21] investigates a class of linear-nonlinear switching ADRC to design speed controllers and current controllers for PMSM in servo systems, which aims at enhancing the ability of disturbance rejection of speed and current controllers. In contrast, [22] deals with performance deterioration due to DC and AC disturbances. It proposes a discrete-time repetitive control-based ADRC for the current loop, and finally, [23] proposes a linear ADRC with a variable gain load torque sliding mode observer to reduce the effects of the load torque disturbance of interior PMSMs. The ADRC method is a robust adaptive extension of the input-output feedback linearization control. It performs the exact linearization of the IM model by a suitable nonlinear state transformation based on the online estimation of the corrective term by the so-called Extended State Observers (ESO). Consequently, any unmodelled dynamics or uncertainty of the parameters are properly addressed. Parameter variations and errors in estimating the total disturbance cannot be included in the endogenous disturbance and cannot be estimated by the ESO, so that these problems may deteriorate the performance of the ADRC method. In [24], an original solution was proposed to partially solve these critical aspects and achieve robustness; in particular, an advanced ADRC controller was developed by adding a sliding mode (SM) component. However, as it is well known, the SM contribution induces chattering and current ripples that, in turn, imply torque oscillations in the input voltage.

In this paper, a new control structure is proposed, where in place of an ESO, a high-gain Unknown Input Observer (UIO) is implemented, with a driving term that is a function of the tracking error. This approach permits overcoming all mentioned problems by achieving total robustness, even against exogenous disturbances coming from the ESO input (See [24, Section IV]). More in detail, in the classical ADRC, there is a cascade of the controller and the ESO, and the controller does not influence the ESO. On the contrary, in the proposed approach, there is an interconnection between the observer and the controller due to the driving term, so the observer can be considered embedded in the controller. As a result, the controller and the UIO influence each other. Moreover, the proposed observer structure does not contain the input u applied to the motor (as for the ESO in ADRC); in this way, all uncertainties associated with inverter nonlinearities, delays, and parameter variation of the input gain are automatically eliminated. Finally, since no SM component is applied in the input voltage as in [24], problems due to chattering and current ripples are avoided. Conversely, the design of the control parameters and the stability proof is more complicated. All these aspects will be better explained in the following parts of the paper. The proposed controller, suitably derived to control induction motor drives, has been tested experimentally on a suitably devised test set-up.

II. Dynamic model of the Induction Motor

The differential equations describing the continuous-time mathematical model of the induction motor in the rotating rotor flux reference frame can be written as follow:

$$\dot{\boldsymbol{x}} = \begin{bmatrix} \dot{i}_{sx} \\ \dot{i}_{sy} \\ \dot{\psi}_{rx} \\ \dot{\omega} \end{bmatrix} = f(\boldsymbol{x}, u_{sx}, u_{sy})$$
$$= \begin{bmatrix} -a_{11}\, i_{sx} + \left(\omega + a_{21}\frac{i_{sy}}{\psi_{rx}}\right) i_{sy} + a_{12}\,\psi_{rx} + c_1 u_{sx} \\ -a_{11}\, i_{sy} - \left(\omega + a_{21}\frac{i_{sy}}{\psi_{rx}}\right) i_{sx} - c_1\,\omega\,\psi_{rx} + c_1\, u_{sy} \\ a_{21}\, i_{sx} - a_{22}\,\psi_{rx} \\ -a_m\,\omega + b_m\left(\frac{2}{3}\,p\, i_{sy}\,\psi_{rx} - t_l\right) \end{bmatrix}, \quad (1)$$

where:

$$a_{11} = \frac{1}{L_e}\left(R_s + \frac{L_s - L_e}{\tau_r}\right), \quad a_{12} = \frac{1}{\tau_r\, L_e}, \quad a_{21} = \frac{L_s - L_e}{\tau_r},$$
$$a_{22} = \frac{1}{\tau_r}, \quad c_1 = \frac{1}{L_e}, \quad a_m = \frac{\rho_v}{J_m}, \quad b_m = \frac{1}{J_m}.$$

i_{sx}, i_{sy}, u_{sx}, u_{sy} and ψ_{rx}, ψ_{sy} are the stator current, the stator voltage and the rotor flux components along x–axis and y–axis of the rotor flux reference frame, ω is the rotor speed, R_s and L_s are the stator resistance and the stator inductance, ρ_v is the viscous friction coefficient, t_l is the load torque, J_m is the inertia coefficient, $\tau_r = \frac{L_r}{R_r}$ is the rotor time constant, where L_r and R_r are the rotor inductance and the rotor resistance, $L_e = L_s - \frac{L_m^2}{L_r}$ is the stator transient inductance, where L_m is mutual inductance and finally p represents the pole pairs.

III. Control Design

The main goal of this work is to design an adaptive control law based on active disturbance compensation employing UIOs. The previous model (II) is useful for FOC techniques since the rotor flux control is decoupled from the speed control. It has been demonstrated in [3], [10] that the decoupling between the speed and rotor flux loop is guaranteed by FOC only in constant flux operation. In contrast, such loops maintain a coupling in variable flux operation (e.g., under ELMT). It is known from the scientific literature that a technique to overcome this drawback is the FL approach (cf. [5], [6], [3]). However, the FL technique has various critical aspects highlighted in the Introduction, and therefore, the ADRC method was proposed.

This work proposes an adaptive control law inspired by the ADRC structure, where a high-gain unknown input observer is considered instead of an ESO. The proposed controller allows us to obtain a more robust system by overcoming problems related to standard FL and ADRC, as highlighted in Section I.

In order to derive the proposed robust control law, the first step is to obtain the flux model (in canonical form), which describes the dynamics of the rotor flux, and the speed model (in canonical form as well), which describes the speed dynamics.

Considering model (II), defining, $x_{\psi_1} = \psi_{rx}$ and $x_{\psi_2} = \dot{\psi}_{rx}$, the flux model can be written as:

$$\dot{x}_{\psi_1} = x_{\psi_2}, \quad (2a)$$
$$\dot{x}_{\psi_2} = h_\psi(\boldsymbol{x}) + b_\psi u_{sx}, \quad (2b)$$

where $b_\psi = a_{21}c_1$ and f is the total disturbance given by:

$$h_\psi(\boldsymbol{x}) = \left(a_{22}^2 + a_{21}a_{12}\right)\psi_{rx} - a_{21}\left(a_{22} + a_{11}\right) i_{sx} + a_{21}\left(\omega + a_{21}\frac{i_{sy}}{\psi_{rx}}\right) i_{sy}.$$

The same procedure is used in order to obtain the speed model. Considering model (II) the mechanical acceleration equation is:

$$\ddot{\omega} = -a_m\dot{\omega} - \frac{2}{3}\,p\,b_m\Big(\left(a_{11} + a_{22}\right) i_{sy} + \omega\left(i_{sx} + c_1\psi_{rx}\right)\Big)\psi_{rx} - b_m\dot{t}_l + \frac{2}{3}\,p\,c_1 b_m \psi_{rx} u_{sy}. \quad (3)$$

By defining $x_{\omega 1} = \omega$ and $x_{\omega 2} = \dot{\omega}$, the speed model is obtained:

$$\dot{x}_{\omega 1} = x_{\omega 2}, \quad (4a)$$
$$\dot{x}_{\omega 2} = h_\omega(\boldsymbol{x}) + b_\omega(\boldsymbol{x}) u_{sy}, \quad (4b)$$

where $b_\omega(\boldsymbol{x}) = \frac{2}{3}\,p\,c_1 b_m \psi_{rx}$, and $h_\omega(\boldsymbol{x})$ is the total disturbance and is given by:

$$h_\omega(\boldsymbol{x}) = -a_m\dot{\omega} - \frac{2}{3}\,p\,b_m\Big(\left(a_{11} + a_{22}\right) i_{sy} + \omega\left(i_{sx} + c_1\psi_{rx}\right)\Big)\psi_{rx} - b_m\dot{t}_l.$$

It is important to note that flux and speed models have the same structure, given by:

$$\dot{x}_1 = x_2, \quad (5a)$$
$$\dot{x}_2 = h(\boldsymbol{x}) + bu. \quad (5b)$$

Below we will focus on the general model (5) as it is more convenient in order to derive the robust control law. By considering system (5), the control variable structure is chosen as follow:

$$u = \frac{1}{b}\left(-\hat{h}(\boldsymbol{x}) + \nu\right), \quad (6)$$

where $\hat{h}(\boldsymbol{x})$ is an estimate of the disturbance $h(\boldsymbol{x})$ and ν is the auxiliary control variable which can be designed so that the speed and rotor flux can evolve according to the desired dynamics. The control law (6) leads to the following model:

$$\dot{x}_1 = x_2, \quad (7a)$$
$$\dot{x}_2 = \left(h(\boldsymbol{x}) - \hat{h}(\boldsymbol{x})\right) + \nu. \quad (7b)$$

Since model (7) is observable and reachable, in order to obtain steady-state null errors, a state feedback control law based on the assignment of the eigenvalues can be derived; however, this technique does not allow to obtain steady-state null errors. So, the best way, in order to achieve a perfect tracking of a constant reference is to add a third variable to the model (7), whose dynamics is described as follows:

$$\dot{r} = x_1^* - x_1, \tag{8}$$

where x_1^* is the desired value of x_1. The model (7) becomes:

$$\dot{x}_1 = x_2, \tag{9a}$$
$$\dot{x}_2 = \big(h(\boldsymbol{x}) - \hat{h}(\boldsymbol{x})\big) + \nu, \tag{9b}$$
$$\dot{r} = x_1^* - x_1. \tag{9c}$$

To assign the internal dynamics of the system, the auxiliary control variable ν is chosen as follow:

$$\nu = -\boldsymbol{k}\left[x_1 \ x_2 \ r\right]^\top, \tag{10}$$

where $\boldsymbol{k} = [k_1 \ k_2 \ k_3]$.

Starting from model (9)-(10), the dynamics is suitably extended in order to estimate the external disturbance $h(x)$, but differently from ADRC, where a classical ESO sourced by the input u is used, a high-gain UIO is considered, with a driving term that is a function of the tracking error. That being stated, the proposed controlled system dynamics is described by the following equations:

$$\dot{x}_1 = x_2, \tag{11a}$$
$$\dot{x}_2 = \big(h(\boldsymbol{x}) - z_3\big) - k_1x_1 - k_2x_2 - k_3r, \tag{11b}$$
$$\dot{r} = x_1^* - x_1, \tag{11c}$$
$$\dot{z}_1 = z_2 + \epsilon^{-1}g_1\left(x_1 - z_1\right), \tag{11d}$$
$$\dot{z}_2 = z_3 + \epsilon^{-2}g_2\left(x_1 - z_1\right) + \rho\text{sign}(x_1^* - x_1), \tag{11e}$$
$$\dot{z}_3 = \epsilon^{-3}g_3\left(x_1 - z_1\right), \tag{11f}$$

where z_1, z_2 and z_3 represent an estimate of x_1, x_2 and $h(\boldsymbol{x})$ respectively, while ϵ, ρ, g_i, i=1, 2, 3, and k_i, i=1, 2, 3, are constant parameters. They will be chosen to ensure the stability of the closed-loop system (11) with sufficiently high margins.

Remark 1: Note that system (11) does not contain the input applied to the motor u for the estimate; in this way, all uncertainties associated with inverter nonlinearities, delays, and parameters variation of the input gain are automatically eliminated by solving the problem highlighted in [24].

To show the stability of systems (11), the model is written in error coordinates by defining an extended error space as follows:

$$e_1 = -r, \tag{12a}$$
$$e_2 = x_1^* - x_1, \tag{12b}$$
$$e_3 = -x_2, \tag{12c}$$
$$e_4 = x_1 - z_1, \tag{12d}$$
$$e_5 = x_2 - z_2, \tag{12e}$$
$$e_6 = h(x) - z_3. \tag{12f}$$

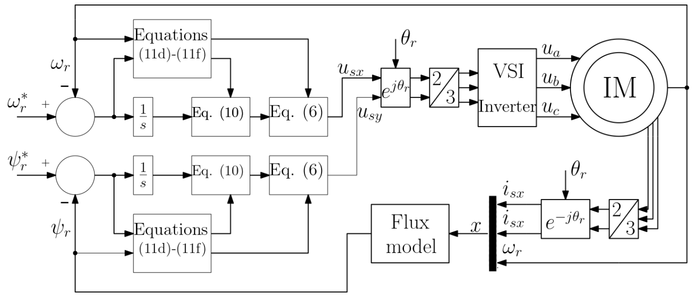


Fig. 1. Block diagram of the proposed control algorithm.

The errors dynamics is computed by means of equations (11) and are given by:

$$\dot{e}_1 = -e_2, \tag{13a}$$
$$\dot{e}_2 = \dot{x}_1^* + e_3, \tag{13b}$$
$$\dot{e}_3 = -k_3e_1 - k_1e_2 - k_2e_3 - e_6 + k_1x_1^*, \tag{13c}$$
$$\dot{e}_4 = -\epsilon^{-1}g_1(e_4) + e_5, \tag{13d}$$
$$\begin{aligned}\dot{e}_5 = {} & k_3e_1 + k_1e_2 + k_2e_3 - \epsilon^{-2}g_2(e_4) + 2e_6 - \rho\text{sign}(e_2)\\ & - k_1x_1^* - h(x),\end{aligned} \tag{13e}$$
$$\dot{e}_6 = \dot{h}(x) - \epsilon^{-3}g_3(e_4), \tag{13f}$$

that can be rewritten in a suitably compact form as follows:

$$\dot{e} = Fe + B_1\text{sign}(e_2) + B_2\mu \tag{14}$$

where

$$F = \begin{bmatrix} 0 & -1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ -k_3 & -k_1 & -k_2 & 0 & 0 & -1 \\ 0 & 0 & 0 & -\epsilon^{-1}g_1 & 1 & 0 \\ k_3 & k_1 & k_2 & -\epsilon^{-2}g_2 & 0 & 2 \\ 0 & 0 & 0 & -\epsilon^{-3}g_3 & 0 & 0 \end{bmatrix}, \tag{15}$$

$$B_1 = \begin{bmatrix} 0 & 0 & 0 & 0 & -\rho & 0 \end{bmatrix}^T, \quad B_2 = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & 0 & k_1 & 0 \\ 0 & 0 & 0 & 0 \\ -1 & 0 & -k_1 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix},$$

and

$$\mu = \begin{bmatrix} h(x) & \dot{h}(x) & x_1^* & \dot{x}_1^* \end{bmatrix}^T.$$

The stability of the estimation error can be inferred by choosing parameters of matrix F such that his eigenvalues are with negative real part.

IV. Test Set-up

A test setup has been suitably built to validate the proposed control technique. The machine under test is a 2.2 kW IM SEIMEC model HF 100LA 4 B5 equipped with an incremental encoder. The employed test setup consists of the following:

- a three-phase 2.2 kW induction motor,
- a frequency converter which consists of a three-phase diode rectifier and a 7.5 kVA, three-phase VSI,
- a dSPACE card (DS1103) with a PowerPC 604e at 400 MHz and a floating-point DSP TMS320F240.

The test set-up is also equipped with a torque-controlled PMSM (Permanent Magnets Synchronous Motor) model Emerson Unimotor FM mechanically coupled to the IM to implement an active load for the IM. A torque-meter model Himmelstein 59003V(4-2)-N-F-N-L-K measures the electromagnetic torque on the shaft. The whole system is processed at 12 kHz.

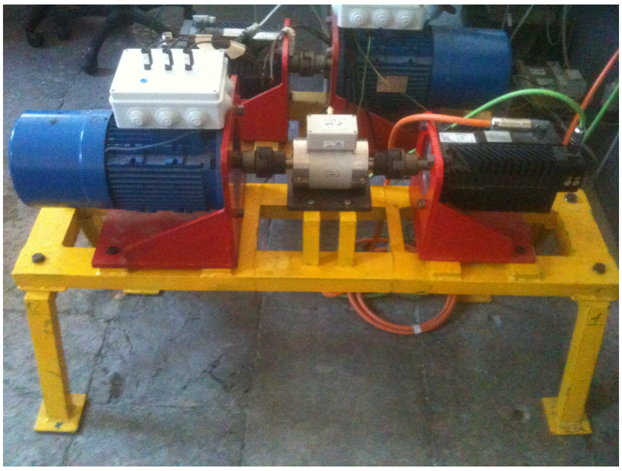


Fig. 2. Photograph of the experimental test set-up.

TABLE I
MOTOR PARAMETERS.

PARAMETER	VALUE
L_s	$0.2030\,\mathrm{H}$
σL_s	$0.017\,98\,\mathrm{H}$
R_s	$2.9\,\Omega$
τ_r	$0.135\,\mathrm{s}$
f_v	$0.0023\,\mathrm{N\,m\,s}$
J_m	$0.0088\,\mathrm{Nms}^2$
p	2

V. EXPERIMENTAL RESULTS

The proposed innovative version of the ADRC, suitably devised for the IM drives, has been experimentally tested on the test set-up described in section IV. Two kinds of tests have been performed (in the full version of the paper, additional experimental results will be provided). Both tests have been performed in constant flux, while in the full paper also, the variable flux operation will be investigated. The first test is a transient response. A constant reference flux equal to 0.8 Wb, corresponding to the rated flux of the IM, has been provided to the IM drive. A set of speed step references of the type 15-30-50 rad/s has been given at no load. Fig.s 4 shows clearly that the measured speed properly tracks its reference with high dynamic performance and null steady-state error. Even the estimated rotor flux amplitude tracks its constant reference of 0.8 Wb with negligible oscillations. The electromagnetic torque presents a step-like waveform, with peaks occurring at each reference speed change, as expected. The direct component of the stator current has a waveform proportional to the rotor flux. In contrast, as expected, the quadrature component has a waveform proportional to the electromagnetic torque. The torque response to the step speed variation is rapid, as desired. The second test is a load rejection test at a constant speed of 50 rad/s. A load step torque of 10 Nm (close to the rated one of the IM) is firstly applied and afterward released (adopting the torque-controlled PMSM exploited as active load). Fig.s 3 presents the same waveforms shown in the former test. It can be seen that the speed controller quickly reacts to the load leading the measured speed to the reference one after the application/release of the step load torque. The rotor flux amplitude and direct component of the stator current are maintained constant, as desired, while the electromagnetic torque and the quadrature component of the stator current increase as soon as the step load torque is applied. The above results show the good dynamic performance achievable with the proposed ADRC.

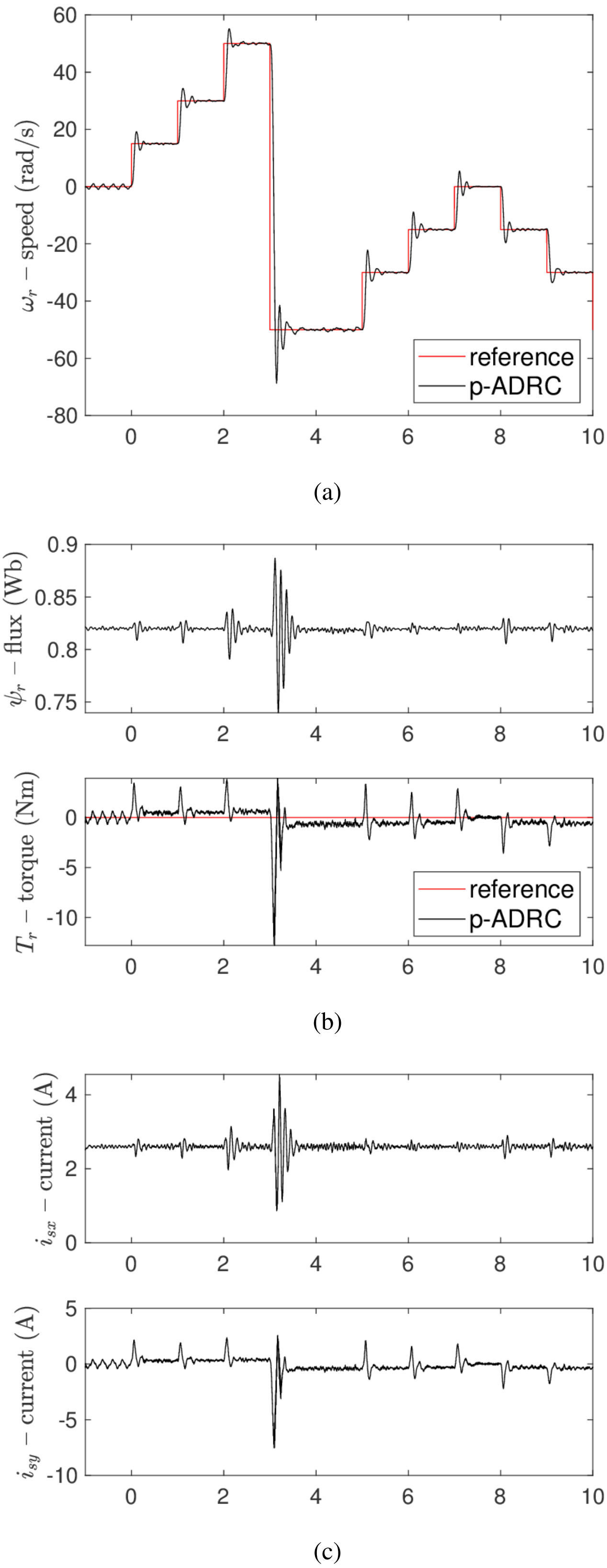
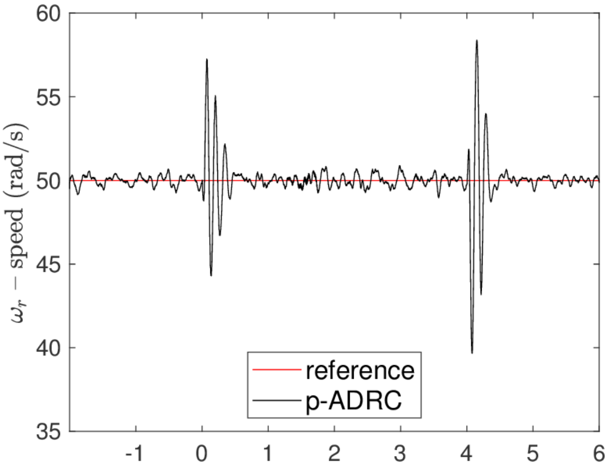
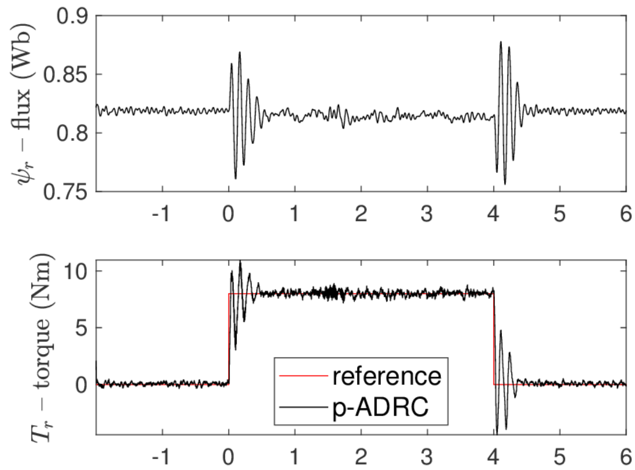


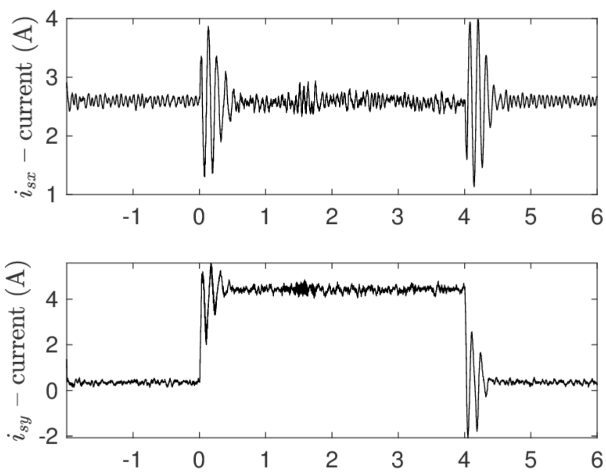
Fig. 3. Reference and measured speed (a), rotor flux amplitude and electromagnetic torque (b) and direct and quadrature component of the stator current (c) in the field-oriented reference frame for a transient response test with a constant reference flux equal to 0.8 Wb and a set of speed step references of the type 15-30- 50 rad/s at no load.



(a)



(b)



(c)

Fig. 4. Reference and measured speed (a), rotor flux amplitude and electromagnetic torque (b) and direct and quadrature component of the stator current (c) in the field-oriented reference frame for a load rejection test at a constant speed of 50 rad/s.

VI. Conclusion

This paper presents a new active ADRC for induction motor drives. In the proposed approach, a high-gain unknown input observer is implemented, with a driving term that is a function of the tracking error, differently from a classic ESO. This approach allows all the classic ADRC problems to be overcome, achieving total robustness. The proposed ADRC has been experimentally verified on a suitably devised test set-up.

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