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Nonlinear robust ADRC Control of Induction Machine

Abstract. *A robust control scheme is proposed for the induction motor speed tracking problem to overcome both parametric uncertainties and unmodelled dynamics. It is based on the augmentation of the well-known field oriented control (FOC) method of the induction machine controlled in the decoupled d-q model by an extended state observer which is designed to estimate the neglected uncertain terms combined with a PD Controller to ensure the convergence of the tracking errors. Furthermore, simulation results show that the proposed methodology is very efficient for both speed and flux tracking in the presence of uncertainties.*

Streszczenie. *Zaproponowano odporny schemat sterowania dla problemu śledzenia prędkości silnika indukcyjnego w celu przezwyciężenia zarówno niepewności parametrycznej jak i niemodelowanej dynamiki. Opiera się on na rozszerzeniu znanej metody sterowania zorientowanego na pole maszyny indukcyjnej sterowanej w odsprzężonym modelu d-q o rozszerzony obserwator stanu, który został zaprojektowany w celu oszacowania zaniedbanych niepewnych warunków w połączeniu z regulatorem PD w celu zapewnienia zbieżności błędów śledzenia. Ponadto, wyniki symulacji pokazują, że proponowana metodologia jest bardzo skuteczna zarówno dla śledzenia prędkości jak i strumienia w obecności niepewności.* ***(Nieliniowe wytrzymałe sterowanie ADRC maszyny indukcyjnej)***

Keywords: Induction motor, Active Disturbance Rejection Controller, Field Oriented Control, uncertain parameters, extended state observer, tracking error.
Słowa kluczowe: silnik indukcyjny, sterowanie nieliniowe, aktywny kontroler odrzucania zakłóceń, niepewne parametry, rozszerzony obserwator stanu.

Introduction

The induction motors are widely used in industry applications (electric railways, robots) to generate torque and rotate mechanical loads due to its ruggedness and low cost maintenance [1],[2]. However, controlling the induction machine to achieve highest performance is not straightforward due to high coupling and nonlinearities in the model and internal and external uncertainties acting on the machine. In the past two decades, trajectory tracking control of induction motor (IM) has been widely studied due to the requirements of high performance in the context of fast and accurate response, fast speed changes from disturbances and insensitivity to parameter variations. Consequently, equivalent performance characteristics of a DC motor can be obtained from the Induction Motor if the closed loop field oriented control (FOC) strategy is applied. This technique allows one to achieve fast, precise tracking of demanding trajectories using an IM, which makes the control task easier. However, the decoupling between speed and flux and the sensitivity to parameters changes present serious constraints for the FOC and affect highly the performance of these controllers for wide range of speed operations [3]. In the literature, nonlinear control theory has been applied extensively to control the induction machines such as nonlinear state feedback control and input-output linearization strategies [4], [5] and [6] based on the nonlinear coupled differential equations describing the machine dynamics. Feedback linearization controllers are among the most successful techniques to achieve input/output decoupling, high dynamic performance, and higher power efficiency [7]. Backstepping control is also applied for the control of induction machines trying to gain from the stabilizing nonlinear terms rather than eliminating them in feedback linearization [10],[11]. The main drawback of such methods is that they rely mainly on an accurate model of the machine with precise parameters. Robust nonlinear strategies have the ability to manage a nonlinear system despite the system uncertainties and the existence of external disturbances [9]. Sliding mode control (SMC) is also one of the best strategy to control of induction machines [12], [13] and [14]. However, the main disadvantage of using SMC in practical implementations is chattering problem which is caused by unmodelled dynamics, switching gain value and discontinuous sign function in classical sliding model control and this could damage the physical parts of the system. Several variants of the developed control laws try to relax the constraints of applying such methods to real systems where the uncertainty is structured and linear in the unknown parameters. The active disturbance rejection controller ADRC is a nonlinear control method proposed by J. Q. Han [15],[16] and Zhiqiang Gao [17] , recently applied ADRC to estimate and compensate the external disturbances, and it is independent of the model of the controlled system and insensitive to the variation of its parameters. Therefore, this technique has proven to be effective to solve the control problems of the nonlinear system . As one of the powerful control methods to deal with the uncertainty, the applications of ADRC in various control fields have shown its better adaptability and good robustness [18],[19]. A control method of ADRC for double closed-loop control is presented based on cascade systems [20], especially, in the field of the motor control [8],[22]. Few works based on ADRC have been reported some limited results to improve the robustness of the induction machine classical controllers. Due to the advantage of an ADRC control and its robustness to the external disturbances, it has been applied for a Five-Phase PMSM [23], ADRC has been implemented in order to ensure high dynamic performance of induction motors in [18]. A robust control strategy based on active disturbance rejection controllers is applied to the speed control of the IM drives [21]. But these studies neglected the study of the control robustness and the stability and convergence when the motor parameters vary with rapid varying loads. In this paper, a robust version of the feedback linearization technique is applied to induction motor based on ADRC in order to overcome unknown disturbances and nonlinearities for the flux and speed decoupled subsystems and tackle the limitations of feedback linearization. Furthermore, the robust controller based on the ADRC is capable to estimate unknown parameters of the induction motor and eliminate its effect in the control law in order to achieve a good tracking of the desired trajectory and to solve the problem of unknown fast variations in plant parameters and the load torque. The existing disturbances in the dynamics are estimated online through an Extended State Observer ESO. Then the ADRC control law is derived and its stability is proven using Lyapunov stability theory and the proposed controller efficiency and robustness are ensured through simulations. The rest of the paper is organized as follows: Section 2 is devoted to Induction motor modelling and transformation to the d-q frame and the conventional Field oriented control strategy. ADRC control is proposed to the flux and speed control design as separated second order decoupled subsystems in sections (3 to 5). The stability analysis is shown in section 6. Section 7 is devoted to simulation and result discussions. In the last section, we end up with concluding remarks.

Induction motor model

The induction motor is a complex multivariable nonlinear system. It has attracted the attention of control engineering academia where different control algorithms have been proposed to solve the speed and flux tracking problems. It is generally described by a fifth-order nonlinear differential equation with two inputs. Moreover, the control of an induction motor is very complex, because it is subject to unknown disturbances and the variation of motor parameters due to heating and magnetic saturation expressed in the $\alpha\beta$ fixed-frame and using the mechanical and electrical parameters of the machine, IM is represented by the model [4, 10].

$$(1)\quad \begin{cases} \frac{d\omega}{dt} = \mu\left(i_s\beta\psi_{ra} - i_{sa}\psi_r\beta\right) - \frac{f}{J}\omega - \frac{1}{J}\tau_L \\ \frac{\psi_{ra}}{dt} = -\eta_r\psi_{ra} + n_p\omega\psi_r\beta + \eta_r L_m i_{sa} \\ \frac{\psi_{r\beta}}{dt} = -\eta_r\psi_{r\beta} + n_p\omega\psi_{ra} + \eta_r L_m i_{s\beta} \\ \frac{di_{sa}}{dt} = \beta\left(\eta_r\psi_{ra} + n_p\omega\psi_{r\beta}\right) - \gamma i_{sa} + \frac{1}{\sigma L_s}u_{sa} \\ \frac{di_{s\beta}}{dt} = \beta\left(\eta_r\psi_{r\beta} + n_p\omega\psi_{ra}\right) - \gamma i_{s\beta} + \frac{1}{\sigma L_s}u_{s\beta} \end{cases}$$

In the above model, the angular speed of the rotor is denoted by ω, ψ_r is the flux in the stator reference frame, and i_S and u_S denote the stator currents and voltages, n_p is the number of pole pairs, R_s and R_r are the stator and rotor resistances, M is the mutual inductance, L_s and L_r are the stator and rotor inductances, and the two mechanical parameters: J is the inertia of the rotor and f is the load torque. The resistances R_s, R_r and the inductances L_s , L_r will be treated as uncertain parameters with R_{sn}, R_{rn} and L_{sn}, L_{rn} as their rated values, respectively. $\eta_{rn} = \frac{R_{rn}}{L_{rn}}, \quad \eta_{sn} = \frac{R_{sn}}{L_{sn}}$

$$\delta_{Rr} = \frac{(R_r - R_{rn})}{R_{rn}}, \quad \delta_{Rs} = \frac{(R_s - R_{sn})}{R_{sn}}$$

$$\delta_{Lr} = \frac{(L_r - L_{rn})}{L_{rn}}, \quad \delta_{Ls} = \frac{(L_s - L_{sn})}{L_{sn}}$$

The load torque τ_L will be treated as a bounded time varying disturbance with bounded derivative.

Our essential objective is to elaborate an ADRC output feedback controller that solve the tracking problem of $(\omega^* - \omega)$ and $(\psi^* - \psi)$ for the Induction Motor based on a modified FOC method in the presence of the highly unstructured uncertainties.

Conventional field oriented control

The main idea of FOC is to perform a change of variables to bring the equations into a new coordinates that will be simple to work with, where the currents regulating the flux and the speed are decoupled [24] . Thus, instead of working with (ψ_{ra},ψ_{rb}), one uses the polar coordinate representation (ρ,ψ_d) [4]

$$(2)\quad \rho = \arctan\left(\frac{R_{rn}}{L_{rn}}\right), \psi_d = \sqrt{\psi_{ra}^2 + \psi_{rb}^2}$$

The stator currents and voltages are then expressed in this new coordinates as follows [3]

$$(3)\quad \begin{array}{l} \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \begin{bmatrix} \cos(\rho) & \sin(\rho) \\ -\sin(\rho) & \cos(\rho) \end{bmatrix} \begin{bmatrix} i_{Sa} \\ i_{Sb} \end{bmatrix} \\ \begin{bmatrix} u_d \\ u_q \end{bmatrix} - \begin{bmatrix} \cos(\rho) & \sin(\rho) \\ -\sin(\rho) & \cos(\rho) \end{bmatrix} \begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} \end{array}$$

ψ_d and ρ are polar coordinates for the ordered pair (ψ_{ra}, ψ_{rb}), the word Field-Oriented refers to this new rotating coordinate system whose angular position is ρ so that $(\psi_q = 0)$ that is, this coordinate system has been aligned with the field flux whose position is ρ and magnitude is ψ_d. The rotation matrix used in (2) and (3) is referred to the direct-quadrature (d-q) transformation. The state space model of the IM in the (dq) stator reference frame can be developed yielding.

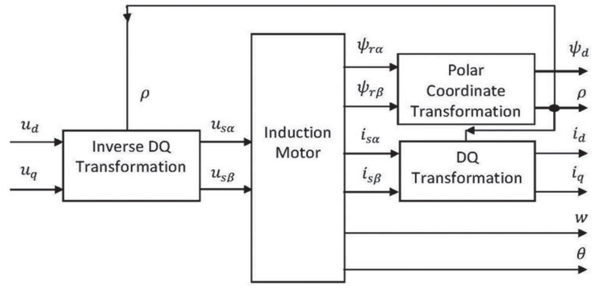

Fig. 1. Transformation to the field-oriented d-q coordinate system.

$$(4)\quad \begin{array}{l} \frac{d\omega}{dt} = \mu\psi_d i_q - \frac{f}{J}\omega - \frac{1}{J}\tau_L \\ \frac{\psi_d}{dt} = -\eta_{rn}\psi_d + \eta_{rn}L_m i_d \\ \frac{di_d}{dt} = -\gamma_n i_d + \eta r_n\beta\psi_d + n_p\omega i_d + \frac{\eta_{rn}L_m i_d^2}{\psi_d} \\ \qquad +f_1 + \frac{1}{\sigma L_s}u_d + \delta_1 \\ \frac{di_q}{dt} = -\gamma_n i_q - \beta n_p\omega\psi_d - n_p\omega i_d - \frac{\eta_{rn}L_m i_d i_q}{\psi_d} \\ \qquad +f_2 + \frac{1}{\sigma L_s}u_q + \delta_2 \\ \frac{d\rho}{dt} = n_p\omega + \frac{\eta_{rn}L_m i_q}{\psi_d} \\ \tau_e = J\mu\psi_d i_q \end{array}$$

$\delta_1 = \delta_R r g_1 + \delta_{Sr} g_2 + \delta_{Lr} g_3 + \delta_{Ls} g_4$
$\delta 2 = \delta_R r g_5 + \delta_{Sr} g_6 + \delta_{Lr} g_7 + \delta_{Ls} g_8$

Where f_1 and f_2 are continuous functions of δ_{Rr}, δ_{Rs}, δ_{Lr}, δ_{Ls} are continuous functions of $(\psi_d, i_d,\ i_q)$. The electromagnetic torque $\tau_e = J\mu\psi_d i_q$ is now just proportional to the product of tow state variables ψ_d and i_q. That is the first four equations of (4) may be written as two decoupled subsystems consisting of the flux subsystem model:

$$(5)\quad \begin{cases} \frac{d\psi_d}{dt} = -\eta_{rn}\psi_d + \eta_{rn}Mi_d \\ \frac{di_d}{dt} = F_d + K_d u_d \end{cases}$$

and the speed subsystem model:

$$(6)\quad \begin{cases} \frac{d\omega}{dt} = \mu i_q\psi_d - \frac{f}{J}\omega - \frac{\tau_L}{J} \\ \frac{di_q}{dt} = F_q + K_q u_q \end{cases}$$

Where $F_d = F_1 + \Delta F_1$
$F_1 = -\gamma_n i_d + \eta_{rn}\beta\psi_d + n_p\omega i_q + \frac{\eta_{rn}Mi_q^2}{\psi_d}$
$\Delta F_1 = f_1 + \delta_1$
and $F_q = F_2 + \Delta F_2$
$F_2 = -\gamma_n i_q - n_p\omega\beta\psi_d + n_p\omega i_q - \frac{\eta_{rn}i_d i_q}{\psi_d}$
$\Delta F_2 = f_2 + \delta_1$
$K_d = K_q = \frac{1}{\sigma L_s}, \gamma_n = \eta_{rn}\beta M + \frac{\eta_{sn}}{\sigma}$

ΔF_1 and ΔF_2 are perturbation terms. The field-oriented control consists of using u_d to force ψ_d to track the constant flux reference $\psi_{dn} = Mi_{dn}$ in the flux subsystem, and the control of speed in the subsystem is done through the input u_q. Consequently, the flux dynamics are now decoupled from the speed dynamics. However, the differential equations for $i_d$and i_q still contain quite complicated nonlinearities in both flux (5) and speed subsystems(6) . One possibility to simplify these current dynamics is to use Feedback cancellation (7).

$$(7)\quad \begin{cases} Ku_d = v_d - \gamma_n i_d + \eta_{rn}\beta\psi_d n_p\omega i_q + \frac{\eta_{rn} M i_q^2}{\psi_d} \\ Ku_q = v_q - \gamma_n i_q - n_p\omega\beta\psi_d - n_p\omega i_q - \frac{\eta_{rn} i_d i_q}{\psi_d} \end{cases}$$

From (5) and (7) it is clear after field-oriented control and nonlinear state feedback, the decoupled dynamics of the IM has a simpler structure. Moreover, the flux amplitude dynamics depend only on the direct current i_d and the direct voltage u_d. Thus, it can be regulated to achieve a given flux amplitude that can generate the desired electromagnetic torque τ. However, the robustness to parameter variation of field orientation and nonlinear state feedback control cannot be guaranteed since the designed controller relies entirely on the exact values of the induction motor parameters which are usually estimated so the unstructured uncertainties could cause system instability. Much work in the literature of robust control have been devoted to deal with such a problem but there is no a universal solution. To solve the problem of unknown variations in plant parameters and structure, in this paper a robust ADRC controller will be designed to eliminate the effect of unstructured uncertainties in each subsystem of the decoupled dynamics of flux and speed subsystems of the IM using ESO.

ADRC controller design

Noting that in applying the feedback (7), there is some uncertainty in the knowledge of the motor parameters and the state variables. Furthermore, the motor parameters R_r and R_s can vary significantly due to Ohmic heating while L_r and L_s can also vary due to magnetic saturation [4].

For that, assuming that all neglected terms for each subsystem as an error signal Δ_i (Δ_d and Δ_q) consequently, the outputs dynamics ($y_d = x_1 = \psi_d$) ($y_q = \xi_1 = \omega$). Defined by (5) and (6), respectively, will be expressed as :

Flux expressed:

$$(8)\quad \begin{cases} \dot{x}_1(t) = x_2(t) \\ \dot{x}_2(t) = f_d(t) + b_d u_d(t) \end{cases}$$

Speed expressed:

$$(9)\quad \begin{cases} \dot{\xi}_1(t) = \xi_2(t) \\ \dot{\xi}_2(t) = f_q(t) + b_\omega u(t) \end{cases}$$

Our goal is to develop a robust output feedback controller based on the ADRC technique that is capable to estimate unknown parameters of induction motor and eliminate its effect in the control law in order to achieve a good tracking of the desired trajectory and to solve the problem of unknown variations in plant parameters and the load torque. In designing a control law for speed tracking and torque load generation based on a given flux reference signal, we consider that the system (4) can be decoupled and divided to two subsystems (5),(6) and (7). For each subsystem, a simpler strategy is followed in this paper where two ADRC controls are used to overcome the effects of nonlinear uncertainties and neglected terms in the tracking accuracy.

Flux and speed tracking controller

The desired torque $\tau_e^* = \mu\psi_d^* i_q$ to be generated with the corresponding reference flux ψ_d^*. In order to the flux tracking, we introduce the state variables $(x_1 = \psi_d), (x_2 = \dot{x_1})$ and $(x_3 = f_d)$ the state space model (8) and for the tracking problem of the speed we define the new state vector$(\xi_1 = \omega)$, $\left(\xi_2 = \dot{\xi}_1\right)$ and $(\xi_3 = f_\omega)$ the speed subsystem dynamics given in (9) can be written as follows:

$$(10)\quad \begin{cases} \dot{x}_1(t) = x_2(t) \\ \dot{x}_2(t) = x_3(t) + b_d u(t) \\ \dot{x}_3(t) = \dot{f}_d(t) \end{cases}$$

Where

$x_3(t) = f_d(t) = \eta_{rn}^2\psi_d - \eta_{rn}^2 + \eta_{rn} M F_d, b_d = \frac{\eta_{rn} M}{\sigma L_s}$
and $u(t) = u_d(t)$

$$(11)\quad \begin{cases} \dot{\xi}_1(t) = \xi_2(t) \\ \dot{\xi}_2(t) = \xi_3(t) + b_q u(t) \\ \dot{\xi}_3(t) = \dot{f}_q(t) \end{cases}$$

Where

$b_q = \frac{\mu\psi_d}{\sigma L_s}$

$f_q(t) = \mu\psi_d\left(-\gamma_n i_q - \beta n_p\omega\varphi_d - n_p\omega i_d - \frac{\eta_{rn} L_m i_d i_q}{\varphi_d}\right)$
$- \frac{f}{J}\dot{\omega} - \frac{\tau_L}{J}$
and
$u(t) = u_q(t)$

Note that the objective of the ADRC technique is to obtain proper estimation, to minimize the total disturbance and to obtain a good trajectory tracking. The ADRC control law is given by (10) and (11)

$$(12)\quad u_d(t) = \frac{u_{d0}(t) - \hat{f}_d(t)}{b_d}$$

and

$$(13)\quad u_q(t) = \frac{u_{q0}(t) - \hat{f}_q(t)}{b_q}$$

where

$u_{d0}(t) = K_{p\psi}\left(x_1(t) - \psi_d^r\right) - K_{D\psi}\dot{\hat{x}}_2(t)$
and
$u_{q0}(t) = K_{p\omega}\left(\xi_1(t) - \omega^r\right) - K_{D\omega}\dot{\hat{\xi}}_2(t)$
are the baseline controller given from a linear feedback controller. The disturbances for the two subsystems(speed and flux) are grouped in $x_3 = \dot{f}_d$ and $\xi_3 = \dot{f}_q$. Using the extended observer they are estimated as $\hat{x}_3$ for $\dot{f}_d$ and $\hat{\xi}_3$ to estimate $\dot{f}_q$. Then, the estimation of the total disturbances f_q and f_d given by the ESO strategy are included in the ADRC augmented control law as shown in figure 2.

Note that the model (10) and (11) are extended dynamic models. This extended model is used to estimate the total disturbance $f_q$and f_d . Models (12) and (13) represent a different way to describe models (10) and (11) as well as the original models (5) and (6). This transformation to the extended model is required to control the linearized model of the induction motor using a base line classical feedback controller. The external disturbances are delt thanks to the ESO

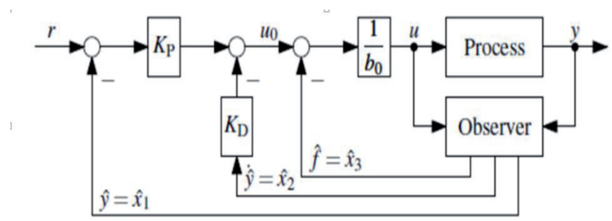

Fig. 2. ADRC principle

that actively estimates the uncertainties and the disturbances and cancel out them using ADRC in the control law as depicted in figure 3.

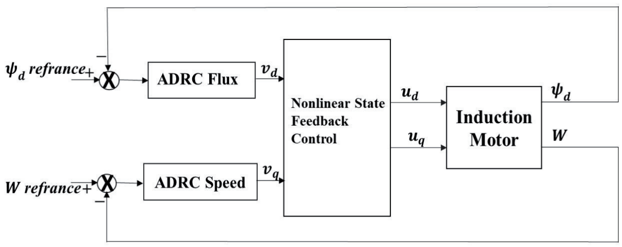

Fig. 3. Decoupled FOC with ADRC control of the IM system.

Stability of ESO

Consider a continuous Lyapunov function where z is an auxiliary variable that needs to be determined on the basis of the stability of the (redundant) output estimation error and of the disturbance Estimation error. The disturbance estimation error is defined as $\eta = f(t)–z$.

The output estimation error $e = (x_1 - \hat{x}_1)$ then evolves according to

$$\left\{\begin{aligned} \dot{e}_1 &= \dot{x}_1 - \dot{\hat{x}}_1 \\ &= x_2 - l_1(x_1 - \hat{x}_1) - \hat{x}_2 \\ &= e_2 - l_1 e_1 \\ \dot{e}_2 &= \dot{x}_2 - \dot{\hat{x}}_2 \\ &= f(t) - b_d u(t) - z - l_2(x_1 - \hat{x}_1) \\ &= \eta - l_2 e_1 \end{aligned}\right. \tag{14}$$

$$\left\{\begin{array}{l} \dot{e}_1 = e_2 - l_1 e_1 \\ \dot{e}_2 = \eta - l_2 e_1 \end{array}\right. \tag{15}$$

$$\dot{e} = Ae + B\eta \tag{16}$$

$$A = \left[\begin{array}{cc} -l_1 & 1 \\ -l_2 & 0 \end{array}\right], B = \left[\begin{array}{c} 0 \\ 1 \end{array}\right]$$

$$\mathrm{A^T P + PA + Q = 0} \tag{17}$$

Let γ be a strictly positive constant parameter and consider the following positive definite Lyapunov candidate in the $(e,)$ space:

$$V = \frac{1}{2}e^T Pe + \frac{1}{2\gamma}\eta^2 \tag{18}$$

The time derivative of the function $V(e,\eta)$ along the solutions of the error dynamics has the following form

$$V = \frac{1}{2}e^T Pe + \frac{1}{2\gamma}\eta^2$$

$$\dot{V} = \frac{1}{2}e^T P\dot{e} + \frac{1}{2}\dot{e}^T Pe + \frac{1}{2\gamma}\eta\dot{\eta}$$

$$\dot{V} = \frac{1}{2}e^T\left(PA + A^T P\right)e + \eta B^T Pe + \frac{1}{2\gamma}\eta\dot{\eta}$$

$$\dot{V} = -\frac{1}{2}e^T Qe + \eta\left(B^T Pe + \frac{1}{2\gamma}\dot{\eta}\right)$$

$$\dot{V} = -\frac{1}{2}e^T Qe + \eta\left(B^T Pe + \frac{1}{2\gamma}(\dot{f}(t) - \dot{z})\right)$$

It is clear that the choice

$\dot{z} = 2\gamma B^T Pe$

$$\dot{V} = -\frac{1}{2}e^T Qe + \eta\left(B^T Pe + \frac{1}{2\gamma}\left(\dot{f}(t) - 2\gamma B^T Pe\right)\right)$$

$$\dot{V} = -\frac{1}{2}e^T Qe + \frac{(f(\mathrm{t}) - z)}{2\gamma}\dot{f}(t) \tag{19}$$

choice for the z dynamics:

$$\dot{V} = -\frac{1}{2}e^T Qe + \frac{(f(\mathrm{t}) - z)}{2\gamma}\dot{f}(t) \leq -\frac{1}{2}e^T Qe + \frac{|f(\mathrm{t}) - z||\dot{f}(t)|}{2\gamma}$$

$$\leq -\frac{1}{2}e^T Qe + \frac{|f(\mathrm{t}) - z|K_1}{2\gamma}$$

Let $\delta_2 > 0$ be a positive scalar quantity. Consider the set $\eta \leq \delta_2$. Outside the set,

$$\left\{\begin{array}{l} \dot{e}_1 = e_2 - l_1 e_1 \\ \dot{e}_2 = f(t) - z - l_2 e_1 \\ \dot{z} = -l_3 e_1 \\ \dot{E} = A_E E + B_E(f(t) - z) \\ Y_E = C_E E \end{array}\right. \tag{20}$$

$$A_E = \left[\begin{array}{ccc} -l_1 & 1 & 0 \\ -l_2 & 0 & 1 \\ -l_2 & 0 & 0 \end{array}\right], B_E = \left[\begin{array}{c} 0 \\ 0 \\ -1 \end{array}\right], C_E = \left[\begin{array}{cc} 1 & 0 \end{array}\right]$$

Assuming that $\dot{\eta}$ is bounded $|\dot{\eta}| \leq \delta_2$ and A_E is a Hurwitz matrix, the error dynamics is exponentially stable [25]. Its characteristic polynomial is given by:

$$\ddot{e}_1 + l_1\ddot{e}_1 + l_2\dot{e}_1 + l_3 e_1 = \dot{f}(t) \tag{21}$$

$$s^3 + \left(\frac{2\zeta\omega_n + p}{\epsilon}\right)s^2 + \left(\frac{2\zeta\omega_n p + \omega_n^2}{\epsilon^2}\right)s + \left(\frac{\omega_n^2 p}{\epsilon^3}\right) = \dot{f}(t)$$

The choice of the observer design parameter set l_1 , l_2 ,l_3 may be settled in accordance with the characteristic polynomial of a desired asymptotically exponentially stable, unperturbed injected observation error dynamics. For suitable strictly positive parameters ξ,$\omega_n$and p, this polynomial may be of the form

$$l_1 = \frac{2\zeta\omega_n + p}{\epsilon}, l_2 = \frac{2\zeta\omega_n p + \omega_n^2}{\epsilon^2} \text{ and } l_3 = \frac{\omega_n^2 p}{\epsilon^3}$$

The extended observer of each subsystem of the induction motor can handle the uncertain terms efficiently and ensure the closed loop system stability and the convergence of the tracking errors.

Simulation Results and discussion

The performance of the FOC-ADRC control of the induction motor system is tested by a differential simulation model under Matlab / Simulink environment. The simulated motor is a 6-poles $(n_p = 3)$, 1/12 horsepower two-phase IM and the rated parameters of the motor were taken from [4] as $R_s = 1.7\Omega$, $R_r = 3.9\Omega$, $L_s = 0.0014H$, $L_r = 0.0014H$, $M = 0.0117H$, $J = 0.00011K.gm^2$, $f = 0.00014N.m/rad/sec$

The simulation of the controlled induction motor is undertaken to achieve an acceleration from rest to a fixed rated speed of $100rad/sec$ followed by a step change to $140rad/sec$ after 0.4 sec at the same time a varying torque load is demanded. Three scenarios are tested, in the first case the machine is controlled at rated parameters and no-uncertainty is present and in the second scenario parametric uncertainty is included by augmenting the motor parameters to 100 $\%$ and applying an external constant torque load. Under the above mentioned conditions the conventional field oriented control PD with and without ADRC augmentation are simulated.

The controller parameters are chosen to achieve good tracking and the settling time of less than 0.1 sec .

To overcome sadden step changes in the speed set-points a first order pre-filter is introduced for the speed reference signal which makes the derivative action of the PD controller smoother and limits the high amplitude of the control signals due to the sadden demand of speed variation.

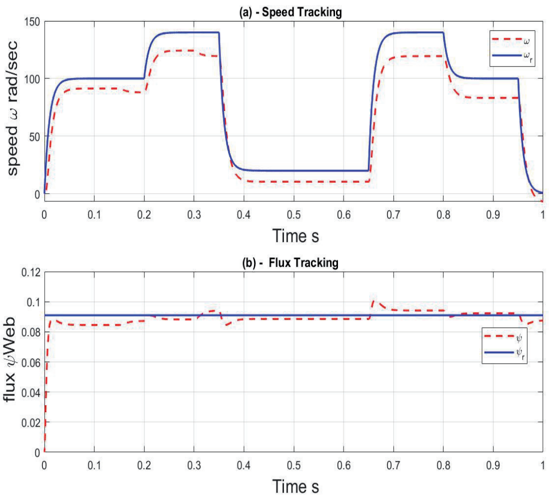

Fig. 4. :Simulation results with constant load and 100% parametric uncertainties of PD control without ADRC.

The speed and flux tracking performance of the PD controller are illustrated in Fig. 4 (a) and (b) respectively which shows that the PD controller is not robust to parametric uncertainties and external disturbance variation. When we changed the reference speed to $140rad/s$ and the torque load is kept constant with variations. The PD system control encountered a tracking problem to follow the reference speed and flux as illustrated in Fig. 4. Fig. 5 shows the achieved performance of ADRC for the perfect speed and flux tracking.

It is obviously clear that ADRC outperforms PD in terms of disturbance rejection and uncertainty in parameters which is an advantage of ADRC strategy.

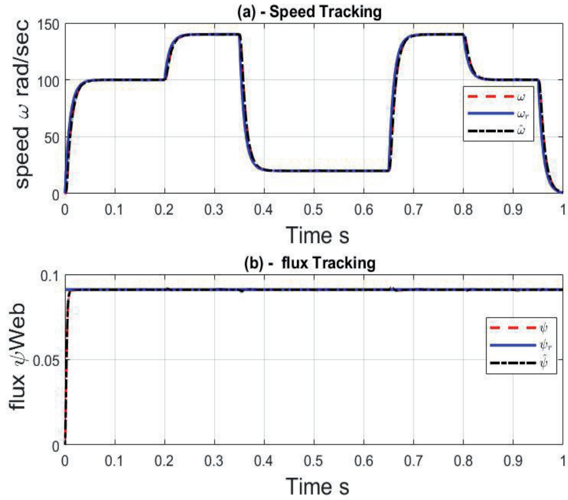

Fig. 5. :Simulation results with constant load and 100% parametric uncertainties with ADRC Controller.

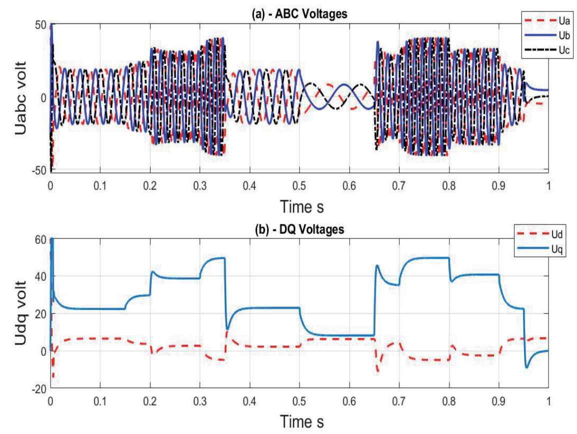

Fig. 6. :Control voltages (U_{ABC} and U_{dq}).

On the other hand, ADRC rejects efficiently the disturbance and compensates the system parameters variation to properly track their reference as illustrated in Fig. 5. Therefore, compared to PD systems, ADRC systems can achieve a smooth start without overshoot and are more resistant to load disturbances.

Fig. 6 illustrates the PD control system's $U_{ABC}$and d, q-axes control voltage waveform respectively, while Fig. 8 depicts the ADRC control system's d, q-axis voltage waveform respectively.

The applied load torque is shown in Fig. 6. It's obviously ADRC controllers in startup demand more voltage than PD controller that refer to the fast response of ADRC, moreover the absence of overshoots and oscillations in the ADRC control system speed response helps the voltage controllers to achieve their references almost without overshoots, whereas the PD control system does not.

For the last simulation scenario, a highly varying load

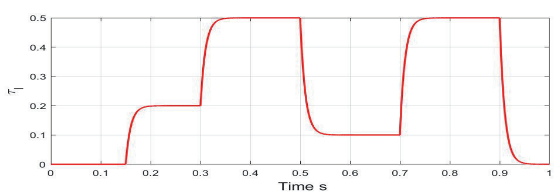

Fig. 7. :The applied varying load.

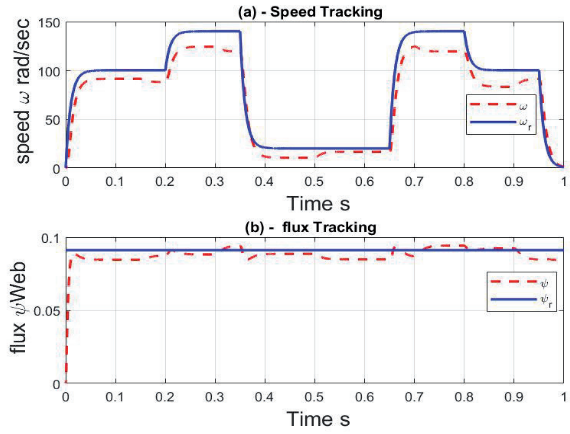

Fig. 8. :Simulation results with load variation and 100% parametric uncertainties of PD controller without ADRC.

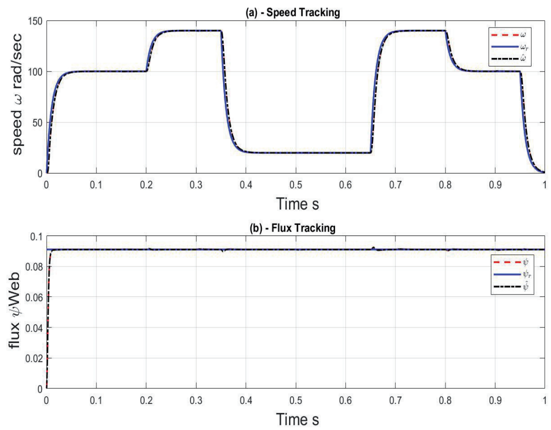

Fig. 9. :Simulation results with load variation and 100% parametric uncertainties with ADRC Controller .

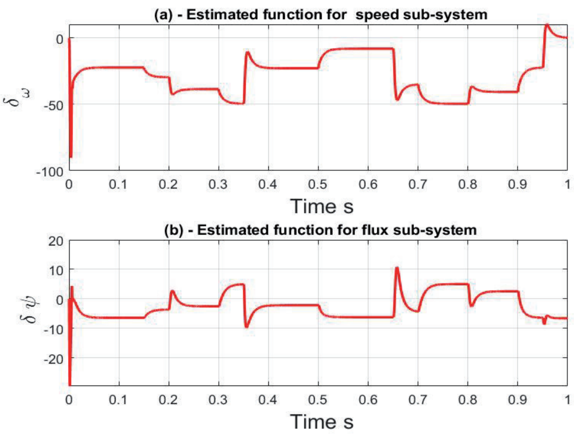

Fig. 10. :Simulation results with'sudden load variation and 100% parametric uncertainties with ADRC Controller .

torque is applied with 100% parametric uncertainty to show the power of ADRC to achieve perfect speed tracking for highly varying load.

Fig. 8 shows the deteriorated and bad tracking of the speed and flux due to the limitation of the PD controller whereas Fig. 9 shows the good tracking for flux and speed regulation of the ADRC controller in spite of the varying load torque. The sudden change in load torque simulation scenario demonstrates that the ADRC controller can estimate and reject the disturbance Fig. 10. The ADRC outperforms the PD controller and it has a faster speed reaction characteristic as well as a stronger disturbance rejection ability.

CONCLUSION

Based on the principle of active disturbance control, this paper analyzes and designs an induction motor modified FOC control composed of two second-order ADRC controllers for flux and speed subsystems. The performance of the ADRC controller and the conventional FOC PD controller are compared using simulations for a nominal and perturbed systems. Stability of the proposed controller is demonstrated using Lyapunov stability theory. According to the conducted simulation results, the ADRC controller improves the system's ability to overcome load disturbances and changes in motor parameters, as well as the tracking precision and robustness for severe varying torque load.

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