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A Nonlinear Control Method for Speed Control of the Induction Motor

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Abstract: *This paper presents a novel nonlinear speed control method for the induction motor. The control scheme is derived in the rotor field coordinates and employs an appropriate estimator for the estimation of the rotor flux angle, flux magnitude, and their derivatives. This nonlinear control strategy resolves the dependency problem of the model parameter deviation and measurement errors. However, in this control strategy, the speed, and flux controller are not sensitive to the IM parameters when it is changed during long-term operation or inaccuracies determined. In addition, these controllers also independent of the load torque, the moment of inertia, and coefficients of the mechanical structure. Therefore, the speed and flux responses are fast and accurate as required.*

Keywords: *Field Oriented Control (FOC), High Performance, Dead-beat.*

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I. INTRODUCTION

Nowadays, an induction motor (IM) drive system with a structure of semiconductor inverter – field-oriented control (FOC) IM has been researched and widely used in the industry. Since FOC has the advantage of controlling IM similar to the independent excitation motors, it has received the most attention ever [1-4]. The FOC structure implemented by a cascade control (the inner loop is torque control; the outer loop is the magnetic and speed control) together with the advanced controllers and available hardware has been opening up the possibility of researching high-quality driven systems that fully meet dynamic and sustainable requirements in fluctuating load environments. It can be realized that the FOC structure can apply linear design methods, or more recently nonlinear methods to the stator, speed, and flux controllers to ensure that the drive quality meets technical requirements.

For the torque controller using the stator current, different control methods have been successfully designed. For example, the deadbeat linear controller for stator current regulators has achieved fast, accurate, and coupling compensation. When this controller is combined with a voltage source inverter, the induction motor fed by a current-source inverter ensures to provide two current components, isd and isq [5-7]. In other studies, the exact linearization method has been successfully designed for the stator current controller in which the state feedback control is designed to separate the flux and torque controllers [1], [8,9]. In this method, the dynamic response of the stator is controllable. Another example is the flatness-based controller, in which fast and smooth stator response in dynamic mode is addressed to limit the current and voltage [1], [10-12].

For the speed controller, the fast response problem was solved by a flatness-based nonlinear control. Besides, different speed trajectories are built such as the trajectory for the control object with differentiable order of the model, and the special trajectories of the variable input systems. Especially, the rest-to-rest trajectory is the desired output [1]. Backstepping control has solved the problem of global stability according to the Lyapunov standard.

Also, the quality of the drive system decreases when the parameters of the motor are incorrect. To address this drawback, an observer that estimates motor parameters and incorporates adaptive controllers is used widely today. Besides, in practical applications, the load is diverse and abundant, such as reactive load and potential load, etc. Therefore, when evaluating the speed controllers for IM driven systems, it is difficult to verify the effectiveness of the proposed control strategies.

II. STATE MODEL OF INDUCTION MOTOR IN THE dq COORDINATE SYSTEM

To implement speed and flux controllers using exact linearization combined with the state-derivative feedback algorithm, a model of the IM with 5th order in the *dq*-coordinate system is used as below [1], [3]:

$$\begin{cases} \dfrac{di_{sd}}{dt} = -\left(\dfrac{1}{\sigma T_s} + \dfrac{1-\sigma}{\sigma T_r}\right) i_{sd} + \omega i_{sq} + \dfrac{L_m i^2{}_{sq}}{T_r \psi_{rd}} + \dfrac{1-\sigma}{\sigma T_r}\dfrac{\psi_{rd}}{L_m} + \dfrac{1}{\sigma L_s} u_{sd} \\ \dfrac{di_{sq}}{dt} = -\omega i_{sd} - \dfrac{L_m}{T_r}\dfrac{i_{sd} i_{sq}}{\psi_{rd}} - \left(\dfrac{1}{\sigma T_s} + \dfrac{1-\sigma}{\sigma T_r}\right) i_{sq} - \dfrac{1-\sigma}{\sigma}\omega \dfrac{\psi_{rd}}{L_m} + \dfrac{1}{\sigma L_s} u_{sq} \\ \dfrac{d\psi_{rd}}{dt} = \dfrac{1}{T_r}\psi_{rd} + \dfrac{1}{T_r} L_m i_{sd} \\ \dfrac{d\theta_{rf}}{dt} = z_p \omega + \dfrac{1}{T_r \psi_{rd}} L_m i_{sq} \\ \dfrac{d\omega}{dt} = \dfrac{3}{2} z_p \dfrac{L_m}{J T_r} \psi_{rd} i_{sq} - \dfrac{z_p}{J} m_L - \dfrac{d}{J}\omega \end{cases} \tag{1}$$

This is a general mathematical model for the IM, which fully describes the electric and mechanical properties of the motor.

In the mathematical model (1):

- Feedback state vectors include motor rotation speed ω; rotor flux ψ_{rd}; current i_{sd}, i_{sq} and magnetic flux angle θ_{rf}.
- The measured state vectors consist of the speed ω; the stator current $i_{s\alpha}$, $i_{s\beta}$. Then, these measured state vectors are used to calculate the rotor flux vetor ψ_{rd} and current vectors i_{sd}, i_{sq} in real-time.
- The remaining state variables will be measured using an estimator.
- Input vectors are the stator voltage vectors u_{sd}, u_{sq}.

This is a nonlinear model. When the flux has not reached a steady state, the currents i_{sd}, i_{sq} vary greatly. The nonlinearity in (1) is shown by the square of the current isq in the first equation, and the product of isd and isq in the second equation. At this time, the phenomenon of coupling occurs. When operating at speeds that are greater than the rated speed, it is required to reduce the magnetic flux ψ_{rd}. The flux controller has strong non-linear characteristics, affecting to the current loop i_{sd}. In addition, the bilinear feature is expressed in the product of state vectors i_{sd}, i_{sq} and the speed input of the motor ω.

According to [5], [6], the deadbeat stator current controller has ensured a fast, accurate current and de-coupling. Accordingly, the state model (1) is reduced to a third-order model as below:

$$\begin{cases} \dfrac{d\psi_{rd}}{dt} = a\psi_{rd} + aL_m i_{sd} \\ \dfrac{d\theta_{rf}}{dt} = z_p\omega + \dfrac{a}{\psi_{rd}} L_m i_{sq} \\ \dfrac{d\omega}{dt} = \dfrac{3}{2} z_p \dfrac{L_m}{J T_r} \psi_{rd} i_{sq} - \dfrac{z_p}{J} m_L - \dfrac{d}{J}\omega \end{cases} \tag{2}$$

with $a = \dfrac{R_r}{L_r}$.

It can be seen that the state model (2) is a 3rd order model, in which the first equation is for the magnetic flux, the second equation is for the rotor flux angle, the third equation is for the rotation. The state model is divided into two subsystems. The first subsystem includes the flux equation and the rotational flux of the rotor, and the second subsystem consists of the rotating motion equation characterized by uncertain parameters such as axis stiffness d, the moment of inertia J, and resistance torque m_W. The first subsystem will be applied to design speed and flux controllers due to this subsystem is less dependent on motor parameters, shaft connection structure, and resistance torque, etc..

III. DESIGN OF THE SPEED AND FLUX CONTROLLERS USING THE EXACT LINEARIZATION WITH STATE-DERIVATIVE FEEDBACK ALGORITHM

From the mathematical model of the IM, the first subsystem is as following:

$$\begin{cases} \dfrac{d\psi_{rd}}{dt} = a\psi_{rd} + aL_m i_{sd} \\ \dfrac{d\theta_{rf}}{dt} = z_p\omega + \dfrac{aL_m i_{sq}}{\psi_{rd}} \end{cases} \tag{3}$$

With state variables $x_1 = \psi_{rd}$; $x_2 = \theta_{rf}$; inputs $u_1 = i_{sd}^*; u_2 = i_{sq}^*$.

Thus, a mathematical model of the first subsystem (3) has two input variables and two output variables. It can be described by the following affine nonlinear equations:

$$\begin{cases} \dfrac{d\mathbf{x}}{dt} = \mathbf{f(x)} + \mathbf{H(x)u} \\ \mathbf{y = g(x)} \end{cases} \tag{4}$$

In which:

$$\mathbf{x} = \begin{bmatrix} \psi_{rd} \\ \theta_{rf} \end{bmatrix}; \mathbf{f(x)} = \begin{bmatrix} -a\psi_{rd} \\ z_p\omega \end{bmatrix}; \mathbf{H}(x) = \begin{bmatrix} aL_m & 0 \\ 0 & (aL_m)/\psi_{rd} \end{bmatrix}$$

Eq. (3) can be rewritten as follows:

$$\begin{cases} \dfrac{dx_1}{dt} = -a\psi_{rd} + aL_m i_{sd} \\ \dfrac{dx_2}{dt} = z_p\omega + aL_m i_{sq}/\psi_{rd} \end{cases} \tag{5}$$

Notice that in Eq. (5), the all output vectors contain the input vector *u*. So the relative order $r_1 = r_2 = 1$. Therefore the system of Eq.(5) can be linearized accurately. The input vector *u* is in the following form:

$$\mathbf{u} = \mathbf{H}^{-1}1\mathbf{(x)}[\dot{\mathbf{x}} - \mathbf{f(x)}] \tag{6}$$

To obtain an optimal control algorithm so that the input and output vectors are split as follows:

$$\mathbf{y}(t) = \mathbf{u}^*(t) \tag{7}$$

Eq. (6) is rewritten as below:

$$\mathbf{u} = \mathbf{H}^{-1}\mathbf{(x)}[\dot{\mathbf{x}} - \mathbf{f(x)} + (\mathbf{u}^* - \mathbf{g}(\mathbf{x}))] \tag{8}$$

From (8), explicitly implemented:

$$i_{sd} = \frac{1}{aL_m}\left[(\frac{d\psi_{rd}}{dt} + a\psi_{rd}) + (\psi_{rd}^* - \psi_{rd}) \right] \tag{9}$$

$$i_{sq} = \frac{\psi_{rd}}{aL_m}\left[(\frac{d\theta_{rf}}{dt} - z_p\omega) + (\omega^* - \omega) \right] \tag{10}$$

Substitute Eq. (9) and Eq. (10) into Eq. (2), yielding:

$$\begin{cases} \psi_{rd} = \psi_{rd}^* \\ \omega = \omega^* \\ \dfrac{d\theta_{rf}}{dt} = z_p\omega^* + \\ \dfrac{aL_m}{b\psi_{rd}^*}\left[\dfrac{d\omega^*}{dt} + (\dfrac{d}{J})\omega^* + (\dfrac{1}{J})m_L \right] \end{cases} \tag{11}$$

From Eq. (11), it can realize that the real signals of speed and flux equal to their reference signals, the error is zero.

To implement the speed controller (9) and the flux controller (10), an estimator is designed to estimate the IM parameters, the derivative of the flux, and the rotational flux of the rotor. In this project, the authors design a PD controller because is simple but effective. The estimated value is calculated by the following formula:

$$\frac{d}{dt}\begin{bmatrix} \hat{\psi}_{rd} \\ \hat{\theta}_{rf} \end{bmatrix} = \begin{bmatrix} -a\hat{\psi}_{rd} + aL_m i_{sd} \\ z_p\omega + (aL_m i_{sq})/\hat{\psi}_{rd} \end{bmatrix} \tag{12}$$

To implement a controller so that the stator current is fast and accurate, implemented by a deadbeat controller. This is one of the advantages of the controller, allowing the IM to operate at a full-speed range, with an arbitrary load torque, a moment of inertial (within the operating limits of the motor). The speed controller (9) and the flux controller (10) only need to know the rotor time constant, the value of the inductance, the number of pairs of poles, measured by the estimator (12). It is recognized that this control strategy uses a minimum number of IM parameters. Therefore, it is necessary to use an estimator. In addition, the estimator (12) improves significantly the stability of the system when there are deviations or inaccuracies of the calculated IM parameters.

IV. CONCLUSION

The advantage of the control strategy implement to design a speed and flux controller such as adjusting PI control parameters. The authors propose a solution to design a speed controller to minimize the dependence on these aforementioned factors, increase the durability of the drive system, resulting in a fast, accurate dynamic response and small torque range. Besides, this control method is least dependent on the motor parameters. It is also independent of the coefficients of the mechanical structure connecting the motor (damping coefficient, coefficient of shaft coupling), load torque, and moment of inertial. Therefore, the speed and flux responses are fast and accurate. As a result, the quality of the drive system is improved. The system's stability is good under the impact of changed or inaccurate motor parameters.

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