

Torque ripples improvement of direct torque controlled five-phase induction motor drive using backstepping control

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ABSTRACT

The paper proposes Direct Torque Control (DTC) of a five-phase induction motor drive with reduced torque ripple. The method presented here is the DTC-Backstepping based on the classic DTC working with a constant switching frequency of the inverter. Another remarkable aspect is the complexity of the method proposed, both in the control unit of the inverter and in the number of correctors necessary for the control of the torque. The selection table and hysteresis have been eliminated. This method significantly improves the torque and flux oscillations and improves the dynamics of the drive by making it less sensitive to load torque disturbances. The proposed method is developed and designed using Matlab/SIMULINK to show the effectiveness and performances of the DTC-Backstepping.

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1. INTRODUCTION

Multiphase machines offer an interesting alternative to the reduction of the stresses applied to the switches as to the windings. Indeed, the multiplication of the number of phases allows a fractionation of the power and hence a reduction of the voltages switched to a given current. Moreover, these machines reduce the amplitude and increase the frequency of the torque ripples, thus enabling the mechanical load to filter them more easily. Finally, the multiplication of the number of phases offers increased reliability by allowing operation when one or more phases are open. Multiphase machines are present in the fields of marine, railway traction, petrochemical industry, avionics, automobile, etc. The rotor flux oriented control mostly employed in multiphase motor drives is based on effective control of the magnetic state. However, this structure generally requires the installation of a speed/position sensor on the shaft. Moreover, it remains very sensitive to the variations of the parameters of the machine, many research have developed this strategy of control applied for multiphase machines [1-6]. Direct torque control (DTC) has been initially developed for induction machines in the years 1986 and 1988 by TAKAHASHI and DEPENDBROCK. In the literature there is an extensive research on DTC applied for multiphase machines [7-14]. In [7,12] a DTC strategy for dual three-phase induction motor drives is discussed. In [8] describes an investigation of direct torque control for multi-phase permanent magnet synchronous motor drives. A DTC for five-phase synchronous motor has

been developed in [9]. Also in [11], authors established a DTC algorithms for a split-phase induction motor. Compared to vector control, the DTC control is much less sensitive to parametric variations and allows to obtain dynamics faster control. A good accuracy in measuring the position of the rotor is not necessary since only the sector in which the flux is located matters to determine the configuration to be used. On the other hand, the flux and torque of the machine must be estimated or observed using different ways. The synthesis of such estimators is not trivial and constitutes a difficulty for the implementation of this control [15]. By nature, torque oscillations exist in DTC drive. The reduction of the hysteresis bands with a given sampling period does not always have an effect on the amplitude of the torque oscillations. In this case, in order to reduce torque oscillations, it is necessary to decrease the sampling period. The switching frequency is not controlled, it varies according to the operating point. DTC with constant switching frequency using SVPWM is reported in [16,17], however, the algorithm is more complex, nevertheless, torque and flux oscillations are reduced. One find in [18], the extended DTC, this control uses selection table of optimal vectors and considers an additional input: the sign of the change of electromagnetic torque. In the extended DTC, the torque regulator has a three-level output, unlike the conventional DTC where only two levels were considered. One can observe a reduction in torque oscillations. In [19,20], the authors present an algorithm allowing to have a constant switching frequency, its main feature is the removal of the hysteresis regulators and the vector selection table, which eliminates the problems associated with them. With this control method, the inverter operates at a constant frequency, since a PWM technique is applied. The objective of this method is to realize a direct control of the stator flux vector in a reference frame linked to the stator, the polar components of these two vectors are obtained by their projections on the reference frame. From these components, the desired stator flux vector at a given instant is calculated. The PWM is applied to this vector to obtain the switching states of the inverter. Applications using fuzzy systems have been developed in [21,22], where the hysteresis blocks have been replaced by fuzzy controllers. Fuzzy logic is used to achieve a compromise between torque control and flux control, and they do not require an exact mathematical model of the machine. However, it draws strong current pulses which normally translates into higher torque ripple. The sliding mode control (SMC), due to its robustness to uncertainties and external perturbations, can be applied to uncertain and disturbed nonlinear systems. It is a question of defining a so-called sliding surface according to the states of the system. The synthesized global command consists of two terms: the first allows to approach this surface, the second allows the retention and the sliding along it [22]. The combination of SMC and DTC control reduces torque and flux ripples. Its main feature is the removal of the hysteresis regulators and the switching table, which eliminates the problems associated with them. The disadvantage of this association (DTC-SMG) is the use of the saturation function that introduces a static error that persists as well as the need to have a knowledge of the dynamics of the system. In [23], authors address a new approach to adapt the concept of the Predictive Control Model (MPC) to Direct Torque Control (DTC) in the five-phase induction motor control. The proposed algorithm improves the performance of a DTC controller by retaining the electromagnetic torque and the stator flux modulus within predefined hysteresis bands while minimizing switching losses. The MPC controller extracts the strings of switching sequences on the forecast horizon. Dynamic programming is implemented to select the switching sequences that minimize the cost function on power losses. Both simulations and experimental are carried out on specific models and the results verify the advantages of the proposed DTC method in comparison with conventional DTC. Model-based predictive direct control methods are advanced control strategies in the field of power electronics to control induction machines [24]. The torque control prediction (PTC) method evaluates the electromagnetic torque and the stator flux in the cost function. The switching vector selected for use in insulated gate bipolar transistors (IGBT) which minimizes error between references and predicted values. System constraints can be easily understood. Behavior and robustness and transient performance are evaluated. Due to the inconvenient of the DTC control such as high torque and currents ripples push us to think and to design new control to improve the direct torque control technique. The idea in this paper is to use the combination between the DTC control and a non-linear control in this paper is the Backstepping control. As the authors know there is no study in the literature treated this approach applied to the multiphase motor and specially five-phase induction motor.

2. DTC-BACKSTEPPING CONTROL OF FIVE-PHASE INDUCTION MOTOR

The induction motor drive system used to develop the proposed controller strategy is composed mainly of five-phase voltage source inverter and a symmetrical five-phase induction motor with phase shift of the windings equal to $2\pi/5$. All the components of the drive system are schematically illustrated in the Figure 1.

Using various assumptions, such as negligible core losses, uniform air gap, sinusoidal MMF distribution and symmetrical distributed windings, then the five-phase induction motor model using the vector space decomposition method presented in [25]. The model represented in two orthogonal planes, where the first plane is involved in the fundamental torque production ($\alpha - \beta$ representing the fundamental component and contains the harmonics of order $10n \pm 1, n = 0, 1, 2, 3, \dots$). The second plane is connected to the losses in the motor ($x - y$ represented the supply harmonics of the order $10n \pm 3, n = 0, 1, 2, 3, \dots$). Finally, the zero-sequence component is not considered due to the neutral point isolation (z -axis representing the supply harmonics of order $5n, n = 0, 1, 2, 3, \dots$). Using the selection of the stator currents in two planes $\alpha - \beta$ and $x - y$ and the rotor fluxes in the plane $\alpha - \beta$ as state variables, $x_1 = i_{s\alpha}, x_2 = i_{s\beta}, x_3 = \psi_{r\alpha}, x_4 = \psi_{r\beta}, x_5 = i_{sx}$ and $x_6 = i_{sy}$. The motor drive equations can be expressed in the following form

$$\begin{cases} \dot{x}_1 = c_1 x_3 + c_2 \omega x_4 - c_3 x_1 + c_4 v_1 \\ \dot{x}_2 = c_1 x_4 - c_2 \omega x_3 - c_3 x_2 + c_4 v_2 \\ \dot{x}_3 = -c_5 x_3 - \omega x_4 + c_6 x_1 \\ \dot{x}_4 = -c_5 x_4 + \omega x_3 + c_6 x_2 \\ \dot{x}_5 = -c_7 x_5 + c_8 v_3 \\ \dot{x}_6 = -c_7 x_6 + c_8 v_4 \\ \dot{\omega} = m_1(x_3 x_2 - x_4 x_1) - m_2 T_l - m_3 \omega \end{cases} \quad (1)$$

With the coefficients given by

$$c_1 = \frac{(1-\sigma)}{\sigma M \tau_r}, c_2 = \frac{(1-\sigma)}{\sigma M}, c_3 = \left(\frac{1}{\sigma \tau_s} + \frac{(1-\sigma)}{\sigma \tau_r} \right), c_4 = \frac{1}{\sigma L_s}, c_5 = \frac{1}{\tau_r}, c_6 = \frac{M}{\tau_r}, c_7 = \frac{R_s}{L_{ls}}, c_8 = \frac{1}{L_{ls}}$$

$$\sigma = 1 - \frac{M^2}{L_s L_r}, \tau_s = \frac{L_s}{R_s}, \tau_r = \frac{L_r}{R_r}, m_1 = \frac{P^2 M}{J_m L_r}, m_2 = \frac{P}{J_m}, m_3 = \frac{B_m}{J_m}$$

The inputs signals of system above are the applied stator voltages $v_1 = v_{s\alpha}, v_2 = v_{s\beta}, v_3 = v_{sx}$ and $v_4 = v_{sy}$. The equations represent the electrical speed ω and all the motor parameters, M is the mutual inductance, R_s is the stator resistance, R_r the rotor resistance, L_r is the rotor inductance, L_s is the stator inductance and L_{ls} is the stator leakage inductance. One presented a novel state variables to describe the novel model to developed the DTC-Backstepping controller, however, the separation of the model of five-phase induction motor is released into two different parts, mechanical part and electrical part, the new state variables are given as follows

$$\begin{cases} T_v = \psi_{r\alpha} i_{s\beta} - \psi_{r\beta} i_{s\alpha} = x_3 x_2 - x_4 x_1 \\ \psi_v = \frac{1}{2}(x_3^2 + x_4^2) \\ X_v = x_3 x_1 + x_4 x_2 \end{cases} \quad (2)$$

Where T_v is a virtual torque, the relation between the electromagnetic torque and the virtual torque is $T_{em} = m_1 T_v$. ψ_v is virtual flux, the relation between the rotor flux and the virtual flux is $\psi_v = \psi_r^2$, where $\psi_r^2 = \psi_{r\alpha}^2 + \psi_{r\beta}^2$. In practice the components $\psi_{r\alpha}$ and $\psi_{r\beta}$ are derived from the five-phase induction motor voltage model or current model given by

$$\begin{cases} \psi_{s\alpha} = \int (v_{s\alpha} - R_s i_{s\alpha}) dt \\ \psi_{s\beta} = \int (v_{s\beta} - R_s i_{s\beta}) dt \end{cases} \quad (3)$$

The rotor flux can be derived as following

$$\begin{cases} \psi_{r\alpha} = L_r(\psi_{s\alpha} - \sigma L_s i_{s\alpha}) \\ \psi_{r\beta} = L_r(\psi_{s\beta} - \sigma L_s i_{s\beta}) \end{cases} \quad (4)$$

The five-phase induction motor model can be separated into electrical and mechanical parts using the new state variables given above, the mechanical part is

$$\dot{\omega} = m_1 T_v - m_2 T_l - m_3 \omega \quad (5)$$

The derivative of the virtual torque is given by

$$\dot{T}_v = \dot{x}_3 x_2 + x_3 \dot{x}_2 - \dot{x}_4 x_1 - x_4 \dot{x}_1 \quad (6)$$

Replacing the derivative $\dot{x}_1, \dot{x}_2, \dot{x}_3$ and \dot{x}_4 from equation (1) one obtain:

$$\dot{T}_v = -(c_3 + c_5)(x_3 x_2 - x_4 x_1) - \omega(x_3 x_1 + x_4 x_2) - c_2 \omega(x_3^2 + x_4^2) + c_4(x_3 v_2 - x_4 v_1) \quad (7)$$

Using equation (2), the equation (7) can be rewritten as

$$\dot{T}_v = -(c_3 + c_5)T_v - \omega X_v - 2c_2 \omega \psi_v + c_4 u_T \quad (8)$$

Where $u_T = x_3 v_2 - x_4 v_1 = \psi_{r\alpha} v_{s\beta} - \psi_{r\beta} v_{s\alpha}$ The derivative of the virtual rotor flux $\dot{\psi}_v$ gives:

$$\dot{\psi}_v = \dot{x}_3 x_3 + \dot{x}_4 x_4 \quad (9)$$

Using equation (1), equation (9) can be rewritten as

$$\dot{\psi}_v = -c_5(x_3^2 + x_4^2) + c_6(x_3 x_1 + x_4 x_2) \quad (10)$$

Using equation (2), equation (10) becomes

$$\dot{\psi}_v = -c_5 X_v + c_6 \psi_v \quad (11)$$

The derivative of X_v gives as follows

$$\dot{X}_v = \dot{x}_3 x_1 + x_1 \dot{x}_3 + \dot{x}_4 x_2 + x_4 \dot{x}_2 \quad (12)$$

Using the equation (1), equation (12) becomes:

$$\dot{X}_v = -(c_3 + c_5)(x_3 x_1 + x_4 x_2) + \omega(x_3 x_2 - x_4 x_1) + c_6(x_1^2 + x_2^2) + c_1(x_3^2 + x_4^2) + c_4(x_3 v_1 + x_4 v_2) \quad (13)$$

Replacing the expressions of equation (2) in equation (13), one obtains:

$$\dot{X}_v = -(c_3 + c_5)X_v + \omega T_v + c_6(x_1^2 + x_2^2) + 2c_1 \psi_v + c_4 u_\psi \quad (14)$$

Where $u_\psi = x_3 v_1 + x_4 v_2 = \psi_{r\alpha} v_1 + \psi_{r\beta} v_2$ Indeed, for the mechanical part, the virtual torque T_v is controlled with a virtual torque voltage u_T and for the electrical part, the virtual rotor flux ψ_v is controlled by the virtual rotor flux voltage u_ψ . This will simplify the design of the DTC-backstepping controller in the following section based on the mechanical and electrical parts. One design by ω^* the electrical motor speed reference, the speed error is given by:

$$z_1 = \omega^* - \omega \quad (15)$$

The derivative of speed error can be rewritten as

$$\dot{z}_1 = \dot{\omega}^* - \dot{\omega} = \dot{\omega}^* - m_1 T_v + m_2 T_l + m_3 \omega \quad (16)$$

Choosing the derivative of the error $\dot{z}_1 = -k_1 z_1$ where k_1 is a positive constant. One can get the reference of the virtual torque T_v as

$$T_v^* = (\dot{\omega}^* + m_2 T_l + m_3 \omega + k_1 z_1)/m_1 \quad (17)$$

The error of the virtual torque gives by $z_2 = T_v^* - T_v$, the derivative of z_2 gives in equation (18).

$$\dot{z}_2 = \dot{T}_v^* - \dot{T}_v = \dot{T}_v^* + (c_3 + c_5)T_v + \omega X_v + 2c_2 \omega \psi_v - c_4 u_T \quad (18)$$

Choosing the derivative of the error $\dot{z}_2 = -k_2 z_2$, one can get the virtual torque voltage u_T as

$$u_T = (\dot{T}_v^* + (c_3 + c_5)T_v + \omega X_v + 2c_2 \omega \psi_v + k_2 z_2)/c_4 \quad (19)$$

Using equation (17) the equation (19) can be rewritten as

$$u_T = ((\ddot{\omega}^* + m_2 \dot{T}_l + m_3 \dot{\omega} - k_1^2 z_1)/m_1 + (c_3 + c_5)T_v + \omega X_v + 2c_2 \omega \psi_v + k_2 z_2)/c_4 \quad (20)$$

Finally, replacing the expression of $\dot{\omega}$ from equation (1), the expression of the virtual voltage u_T given as follows

$$u_T = ((\ddot{\omega}^* + m_2 \dot{T}_l - m_3 m_2 T_l - m_3^2 \omega - k_1^2 z_1 + m_1(m_3 + c_3 + c_5)T_v)/m_1 + \omega X_v + 2c_2 \omega \psi_v + k_2 z_2)/c_4 \quad (21)$$

One assume that the virtual rotor flux of the five-phase induction motor given by ψ_v^* , then its error is given by:

$$z_3 = \psi_v^* - \psi_v \quad (22)$$

The dynamic of the error of the virtual rotor flux is given by the following equation:

$$\dot{z}_3 = \dot{\psi}_v^* - \dot{\psi}_v = \dot{\psi}_v^* + 2c_5 \psi_v - c_6 X_v \quad (23)$$

Choosing the dynamic of the error $\dot{z}_3 = -k_3 z_3$, one obtains the virtual control X_v^* as:

$$X_v^* = (\dot{\psi}_v^* + 2c_5 \psi_v + k_3 z_3)/c_6 \quad (24)$$

Let is now define the error of the virtual control X_v as:

$$z_4 = X_v^* - X_v \quad (25)$$

Its dynamic error is given by:

$$\dot{z}_4 = \dot{X}_v^* - \dot{X}_v \quad (26)$$

Using equation (14), equation (26) becomes

$$\dot{z}_4 = \dot{X}_v^* + (c_3 + c_5)X_v - \omega T_v - c_6(x_1^2 + x_2^2) - 2c_1 \psi_v - c_4 u_\psi \quad (27)$$

Choosing the dynamic of the error $\dot{z}_4 = -k_4 z_4$, where k_4 is a positive constant, then the virtual rotor flux u_ψ obtains as

$$u_\psi = (\dot{X}_v^* + (c_3 + c_5)X_v - \omega T_v - c_6(x_1^2 + x_2^2) - 2c_1 \psi_v + k_4 z_4)/c_4 \quad (28)$$

Using equation (24), one can rewrite the equation (28) as

$$u_\psi = ((\dot{\psi}_v^* + 2c_5 \dot{\psi}_v - k_3^2 z_3)/c_6 + (c_3 + c_5)X_v - \omega T_v - c_6(x_1^2 + x_2^2) - 2c_1 \psi_v + k_4 z_4)/c_4 \quad (29)$$

Replacing the expression of $\dot{\psi}_v$ gives by equation (11) in equation (29), one obtains

$$u_\psi = ((\ddot{\psi}_v^* + 2c_5(-2c_5 \psi_v + c_6 X_v) - k_3^2 z_3)/c_6 + (c_3 + c_5)X_v - \omega T_v - c_6(x_1^2 + x_2^2) - 2c_1 \psi_v + k_4 z_4)/c_4 \quad (30)$$

From the two virtual voltages, both for torque and flux, one can acquire the stator voltages in the $(\alpha - \beta)$ frame.

$$\begin{cases} v_1 = v_{s\alpha} = (x_3 u_\psi - x_3 u_T)/2\psi_v \\ v_2 = v_{s\beta} = (x_4 u_\psi + x_3 u_T)/2\psi_v \end{cases} \quad (31)$$

Now, One need to determinate the control voltages v_3 and v_4 , then one define the following current errors z_5 and z_6

$$\begin{cases} z_5 = i_{sx}^* - i_{sx} = i_{sx}^* - x_5 \\ z_6 = i_{sy}^* - i_{sy} = i_{sy}^* - x_6 \end{cases} \quad (32)$$

The dynamic of the currents errors z_5 and z_6 given as follow

$$\begin{cases} \dot{z}_5 = \dot{i}_{sx}^* - \dot{i}_{sx} = \dot{i}_{sx}^* - \dot{x}_5 \\ \dot{z}_6 = \dot{i}_{sy}^* - \dot{i}_{sy} = \dot{i}_{sy}^* - \dot{x}_6 \end{cases} \quad (33)$$

Choosing the derivative of the errors $\dot{z}_5 = -k_5 z_5$ and $\dot{z}_6 = -k_6 z_6$, the stator voltages in the reference $(x - y)$ obtained in the following equation

$$\begin{cases} v_3 = v_{sx} = (\dot{i}_{sx}^* - c_7 x_5 + k_5 z_5)/c_8 \\ v_4 = v_{sy} = (\dot{i}_{sy}^* - c_7 x_6 + k_6 z_6)/c_8 \end{cases} \quad (34)$$

3. STABILITY ANALYSIS

Usually it is difficult to find the asymptotic stability of time-varying systems because it is very difficult to find Lyapunov functions with a negative definite derivative. We know in case of autonomous (time invariant) systems, if V is negative semi-definite, then it is possible to know the asymptotic behaviors by invoking invariant-set theorems. However, the flexibility is not available for time-varying systems. This is where Babarlat's lemma comes into picture.

3.1. Lemma: Barbal't's Lemma

Suppose $f(t) \in C^1(a, \infty)$ and $\lim_{t \rightarrow \infty} f(t) = \gamma$ where $\gamma < \infty$. If f' is uniformly continuous, then $\lim_{t \rightarrow \infty} f'(t) = 0$.

3.2. Proof

One will prove the result by contradiction. $\lim_{t \rightarrow \infty} f(t) \neq 0$. Then $\exists \epsilon > 0$ and a monotone increasing sequence $\{t_n\}$ such that $t_n \rightarrow \infty$ as $n \rightarrow \infty$ and $|f'(t)| \geq \epsilon$ for all $n \in N$. Since $f'(t)$ is uniformly continuous, for such ϵ , $\exists \chi > 0$ such that $\forall n \in N$.

$$|t - t_n| < \chi \Rightarrow |f'(t) - f'(t_n)| \leq \epsilon/2$$

Hence if $t \in [t_n, t_n + \chi]$ then

$$|f'(t)| = |f'(t_n) - (f'(t_n) - f'(t))| \geq |f'(t_n)| - |f'(t_n) - f'(t)| \geq \epsilon - \epsilon/2 = \epsilon/2$$

Then since $f(t) \in C^1$ one have

$$\left| \int_a^{t_n+\chi} f'(t) dt - \int_a^{t_n} f'(t) dt \right| = \left| \int_{t_n}^{t_n+\chi} f'(t) dt \right| \geq \int_{t_n}^{t_n+\chi} |f'(t)| dt \geq \int_{t_n}^{t_n+\chi} \epsilon/2 dt = \chi\epsilon/2$$

However

$$\begin{aligned} \lim_{t \rightarrow \infty} \left| \int_a^{t_n+\chi} f'(t) dt - \int_a^{t_n} f'(t) dt \right| &= \lim_{t \rightarrow \infty} |f(t_n + \chi) - f(t_n)| \\ &= \left| \lim_{t \rightarrow \infty} f(t_n + \chi) - \lim_{t \rightarrow \infty} f(t_n) \right| = |\gamma - \gamma| = 0 \end{aligned}$$

This is a contradiction. Therefore $\lim_{t \rightarrow \infty} f'(t) = 0$

3.3. Application of the Lemma

One assume the Lyapunov function is

$$V = \frac{z_1^2 + z_2^2 + z_3^2 + z_4^2 + z_5^2 + z_6^2}{2} \quad (35)$$

The derivative of the Lyapunov function given by

$$\dot{V} = z_1 \dot{z}_1 + z_2 \dot{z}_2 + z_3 \dot{z}_3 + z_4 \dot{z}_4 + z_5 \dot{z}_5 + z_6 \dot{z}_6 = -k_1 z_1^2 - k_2 z_2^2 - k_3 z_3^2 - k_4 z_4^2 - k_5 z_5^2 - k_6 z_6^2 \quad (36)$$

From equation (36), one have $\dot{V} \leq -k_1 z_1^2$, then

$$V(0) - V(\infty) \geq \int_0^\infty -k_1 z_1^2 dt \quad (37)$$

V is bounded. According to Barbalat's lemma one can have

$$\lim_{t \rightarrow \infty} z_1 = 0 \quad (38)$$

Also, one can have

$$\lim_{t \rightarrow \infty} z_2 = \lim_{t \rightarrow \infty} z_3 = \lim_{t \rightarrow \infty} z_4 = \lim_{t \rightarrow \infty} z_5 = \lim_{t \rightarrow \infty} z_6 = 0 \quad (39)$$

At last, according equations (38) and (39), one can notice that motor speed, torque and rotor flux can track reference values asymptotically.

4. SIMULATIONS RESULTS AND DISCUSSION

DTC-Backstepping control system block diagram is shown in Figure 1. Whenever the reference speed command ω^* is given, system compares it with the actual speed ω . The error is used to determine the virtual reference torque T^* . The reference virtual torque is compared with actual torque T . The torque error obtained is used to obtain the appropriate virtual voltage vector u_T . The second reference input that is flux compare with actual flux. The error added to determinate the virtual control X . Like torque control loop, flux control loop is also operated to select the appropriate virtual voltage vector u_ψ . However, After the determination of the two virtual voltage vectors which used to obtain the actual voltage vectors $(\alpha - \beta)$. The two other voltage vectors $(x - y)$ are obtains from a small loop current controller. A program in Matlab/Simulink environment has been designed for both the voltage source inverter (VSI)-fed five-phase induction machine, and performed to analyze the viability of the developed control technique a number of simulations have been performed. A small prototype around 1.5 kW symmetrical five-phase induction machine with four poles has been used. Table 1 shows the parameters of the machine, which correspond to those of the real five-phase induction machine.

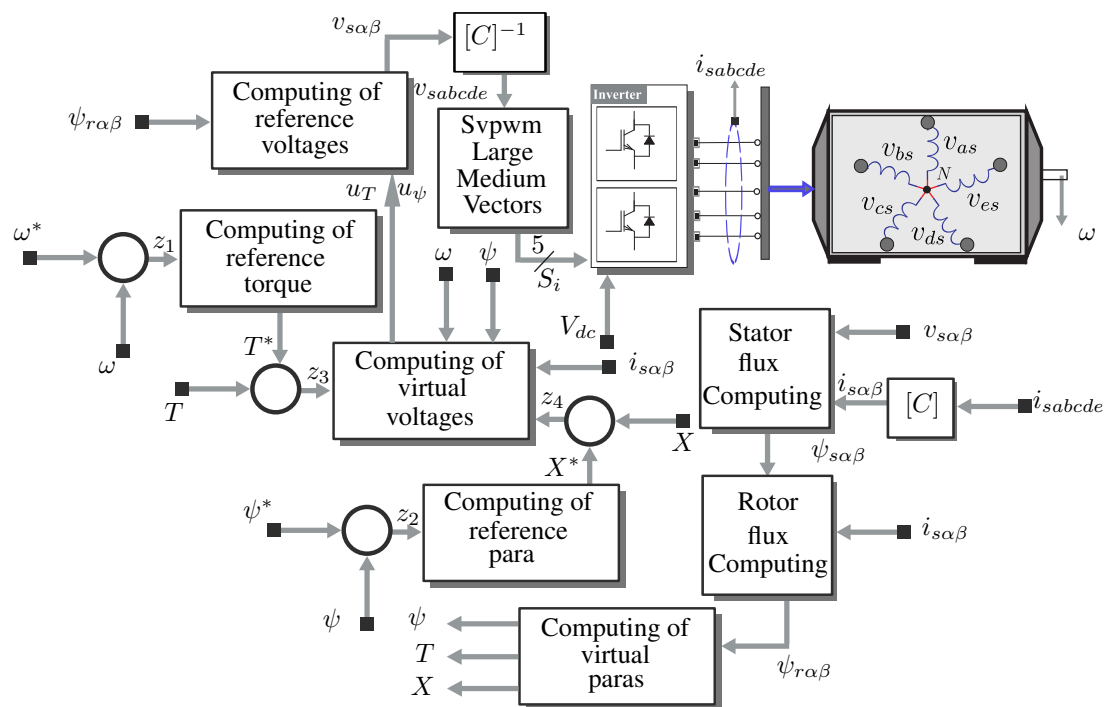


Figure 1. Scheme of Direct torque Backstepping Control of five-phase Induction Motor

Table 1. Parameters of the five-phase Induction Machine

Motor Data				
Parameters	Symbol	Value	Units	
Stator resistance	R_s	10	Ω	
Rotor resistance	R_r	6.3	Ω	
Mutual Inductance	M	0.42	H	
Stator leakage inductance	L_{ls}	0.04	H	
Rotor leakage inductance	L_{lr}	0.04	H	
Moment of inertia	J_m	0.03	$kg.m^2$	
Stator inductance	L_s	0.46	H	
Rotor inductance	L_r	0.46	H	
Number of pole pairs	P	2	-	
Rated stator flux	ψ_n	1.27	Wb	

Figures 2 and 3 summarize the simulation results. A start-up transient under no-load conditions, with speed references reversion from 400 rpm/min to -400 rpm, is considered, and the motor-speed, electromagnetic-torque, stator-flux, and stator-current responses and the torque ripple are shown for both methods (DTC and DTC-Backstepping). The obtained speed response using the DTC-Backstepping method is quicker than that using the DTC method, while the settling time is practically the same. By using DTC-Backstepping technique, torque ripples is reduced from 2.5 N.m to 1.2 N.m as shown in figure.3. Hence torque ripple is reduced by 52%. Flux ripple is reduced from 0.1 Wb to 0.01 Wb as shown in figure.3. Therefore flux ripple is reduced by 90%. It is observed that the DTC-Backstepping which based on non-linear approach allows effectively to reduce torque ripple as well as flux ripple. From the simulation figures, it can be concluded that torque and flux direct backstepping control has small torque, speed, flux and current ripple. Torque and flux direct backstepping control have better performance than classical direct torque control. The scheme can not only avoid the effect of rotor resistance to rotor flux orientation, but also reduce effect of hysteresis to torque and flux. From simulation results, one view that system design can achieve direct torque and flux control and motor speed tracking performance is better than classical direct torque control. The control scheme can realize system state variable decoupling efficiently.

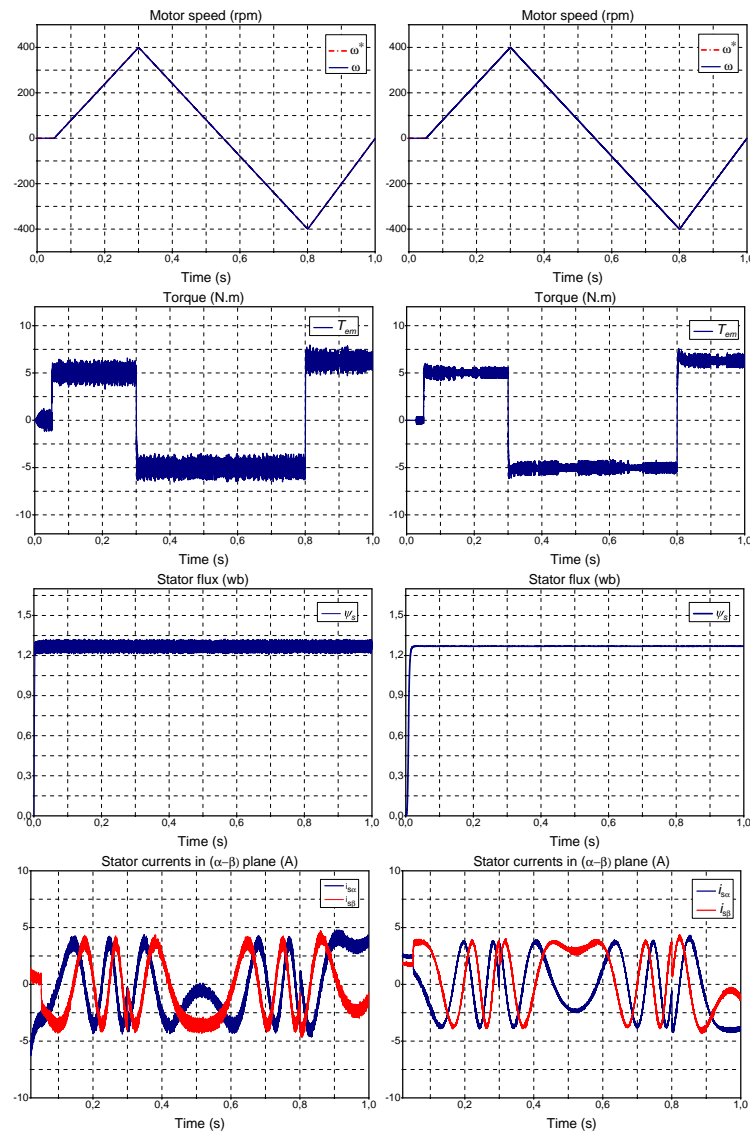


Figure 2. Motor speed, torque, stator flux and currents left side DTC and right side DTC-backstepping

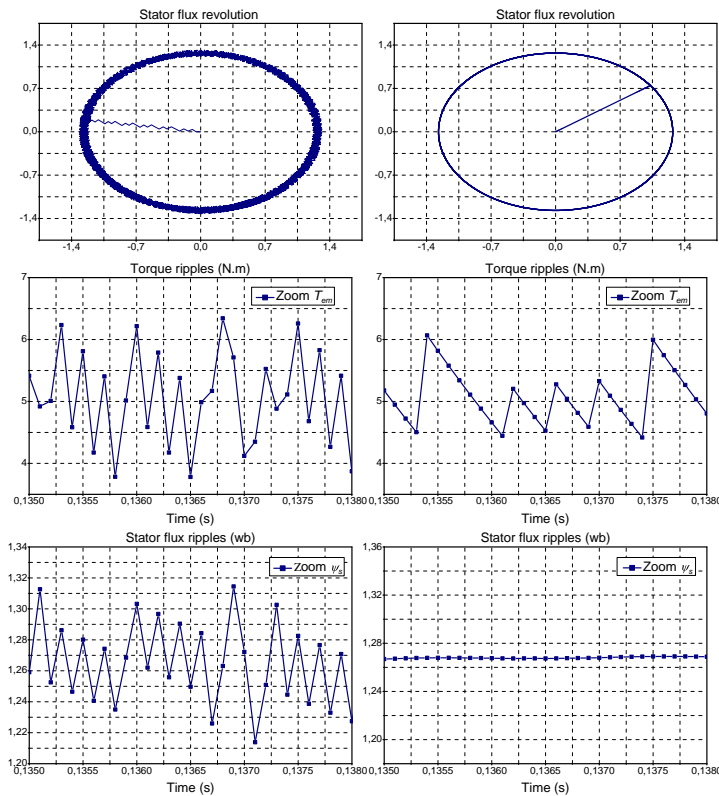


Figure 3. Stator flux revolution, torque and flux ripples, left side DTC and right side DTC-backstepping

5. CONCLUSION

This paper separates induction motor into mechanical part and electrical part. Backstepping coupling is applied to realize torque and flux decoupling. This can reduce the couple between state variable, and can make speed, torque and flux fast track reference values. System design is based on static coordinate of stator. The control scheme synthesizes the direct torque and vector control, which not only realizes torque and rotor flux direct control, but also have the little torque and rotor flux ripple. Backstepping control design can not only provide better speed, torque and flux tracking performance, but also assert system robust performance under speed reversal. Moreover, the torque and flux are respectively reduced by 52% and 90%.

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