**TELKOMNIKA Indonesian Journal of Electrical Engineering**
Vol. 12, No. 9, September 2014, pp. 6587 ~ 6593
DOI: 10.11591/telkomnika.v12i9.5074


# Back-stepping Adaptive SVM Direct Torque Control of SPMSM Drive system

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***Abstract***

*A novel back-stepping adaptive control strategy is proposed to direct torque control (DTC) of surface permanent magnet synchronous motor (SPMSM) in order to solve some problems existed in conventional DTC, such as high flux , torque and current ripple, variable switching frequency, and acoustic noises. PI speed regulator and two hysteresis regulators in the conventional DTC system are substituted by the proposed back-stepping adaptive controller respectively. The output of the designed controller makes space vector modulation (SVM) possible, which further ensures the inverter switching frequency to be fixed. The stability of the controller is verified via Lyapunov stable theory.Simulation results show that the presented control strategy can solve these problems effectively and show strong robust to external disturbance and noise.*

***Keywords***: *permanent magnet synchronous motor, direct torque control, back-stepping adaptive control, space vector modulation*



## 1. Introduction

Permanent magnet synchronous motor (PMSM) has been widely used employed in servo applications such as robotics, chip-mount machines, NC machine, and hard disk drives because its features of low noise, low inertia, high efficience, and low maintenance cost [1, 2]. During the past years, direct torque control (DTC) scheme for PMSM drives has receive enormous attention in industrial motor drive application due to its potential advantages, for example, its structure is simple because all calculations are implemented in stationary; the field-weakening control becomes easier because the stator flux linkage can been controlled directly in the DTC system; sensorless control becomes possible because the method does not need accurate rotor position information [3]. In addition, Compared with vector control, DTC directly manipulates the final output voltage vector without the need for inner current loops, hence eliminating the inherent delay caused by current loops and featuring a high dynamic response.

Desipite the merits aforementioned, convertional DTC employs two hysteresis comparators and a heuristic switching table to obtain quick dynamic response, which will cause some drawbacks, including high flux , torque and current ripple, variable switching frequency, high sampling requirement for digital implementation, and difficulty to accurately control at low speed and high frequency noise caused by high torque ripple [4]. In the past decades, numerous methods have been presented to address these problems of conventional DTC. Many of them [5-7] employ space vector modulation and PI regulator to achieve fixed switching frequency and low flux, torque and current ripple. However, this scheme is relatively noisy [8]. Multilevel inverter is introduced to obtain more voltage vectors [9], but it will increase the hardware cost and system complexity. In the literature [10, 11], more accurate and complex switching table is constructed by dividing one sampling period into several intervals, which can achieve excellent performance, but they are usually complicated and rely much on the knowledge of motor parameters. Two PI regulators are used to improve the performance of a DTC system in [12]. This method requires the continuous information of the stator flux vector position; hence, the drive performance relies heavily on the accuracy of the stator flux estimation. Moreover, PI controller is sensitive to changes in motor parameters and external

*Received November 7, 2013; Revised May 12, 2014; Accepted June 15, 2014*

load. Sliding mode variable structure based DTC control scheme is investigated in [13-15], though this scheme can reduces flux, torque and current ripples significantly, it requires gain scheduling and is parameter dependant. Moveover, sliding mode control has the inherent chattering phenomenon. Recently, predictive control has also been introduced to motor dirves. A predictive control based three-level inverter-fed DTC method is proposed in [16, 17], which utilizes a predictive horizon greater than one to obtain reduced switching frequency while keeping flux, torque, current, and neutral point potential within their respective hysteresis bands.

The back-steeping adaptive control is a systematic and recursive design methodology for nonlinear system [18]. In this method, if the control inputs selected properly, the controlled system can be stabilized quickly. In this paper, back-steeping adaptive control is used to improve the performance of DTC SPMSM drives. The control voltages are synthesized by using SVM strategy, which ensure that switching frequency is constant. The stability of the system is verified via Lyapunov stable theory.

This paper is organized as follows. Section 2 introduces the dynamic mathematical model of SPMSM drive system. Then, controller is designed in detail and the stability of the controlled closed systems is verified according to Lyapunov stability theory. Section 3 presents the simulation results to illustrate the effectiveness of the method. Finally, Section 4 concludes.

## 2. Controller Design for SPMSM Drive System

In this section, it is given the dynamic model of the SPMSM drive system and at the same time the controller is designed in detail. Then, The stability of the proposed control scheme is verified via Lyapunov stable theory.

### 2.1. Dynamic Mathematical Model of the SPMSM Drive System

A three-phase SPMSM drive system can be modeled in the α-β frame as:

$$\begin{cases} \dot{i}_\alpha = (u_\alpha - R_s i_\alpha - E_\alpha)/L_d \\ \dot{i}_\beta = (u_\beta - R_s i_\beta - E_\beta)/L_q \\ \dot{\psi}_\alpha = u_\alpha - R_s i_\alpha \\ \dot{\psi}_\beta = u_\beta - R_s i_\beta \\ \dot{w} = [P(T_e - T_L) - B_n w]/J \end{cases}, \tag{1}$$

$$T_e = \tfrac{3}{2} P(\psi_\alpha i_\beta - \psi_\beta i_\alpha), \tag{2}$$

$$\Psi = \psi_\alpha^2 + \psi_\beta^2, \tag{3}$$

Where $i_\alpha$, $i_\beta$, $u_\alpha$, $u_\beta$, $\psi_\alpha$, $\psi_\beta$, $E_\alpha$, and $E_\beta$ are the stator currents, voltages, flux linkage, and electromotive-force (EMF) respectively, $L_d$, and $L_q$ are the inductances, $R_s$ is the stator resistance, *P* is the number of poles, $T_e$ and $T_L$ is electromagnetic torque and load torque, $w$ is the motor speed, $B_n$ is the viscous friction coefficient, $J$ is the rotor inertia, $\Psi$ is the stator fulx linkage norm, and $\begin{cases} E_\alpha = -Pw\psi_f \sin\theta \\ E_\beta = Pw\psi_f \cos\theta \end{cases}$, $\psi_f$ is the permanent magnet fulx linkage, $\theta$ is the rotor angle.

### 2.2. Controller Design

The configuration of the proposed backstepping adaptive control scheme for SPMSM drive system is depicted in Figure 1. It can be seen from Figure 1 that the proposed control scheme is comprised of speed control and flux and torque control. The output of backstepping speed controller is the reference electromagnetic torque $T_e^*$, and the output of flux and torque controller are the reference voltage $u_{s\alpha}^{*}$ and $u_{s\beta}^{*}$, then, the SPMSM-fed is drived via SVM technology.

The control objective is to design the backstepping adaptive DTC controller so that the motor speed can track any desired command $w^{*}$. The controller is designed step by step as follows.

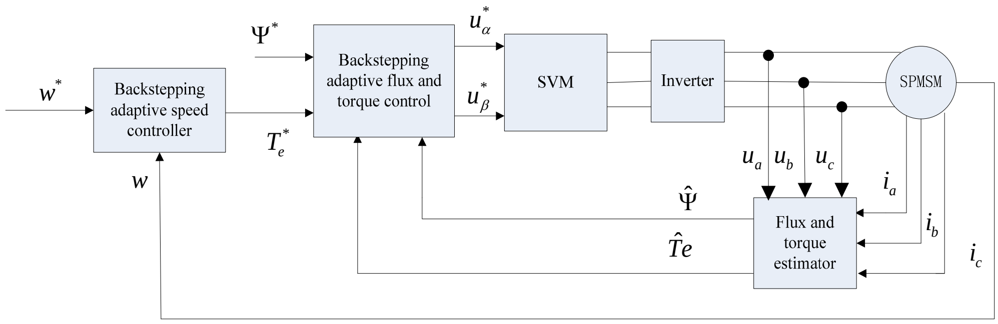



Figure 1. Block Diagram of the Proposed Backstepping Adaptive DTC Scheme for PMSM Drive System

Step 1: Speed controller design

Define the following speed track error: $e_w = w^{*} - w$

Then, the derivative of speed track error can be represented as:

$$\dot{e}_w = -\dot{w} = -[p(T_e - T_{L)} - B_n w]/J \ . \tag{4}$$

Define the Lyapunov function: $V_1 = \frac{1}{2}e_w^2$

Take the derivative of the Lyapunov function $V_1$, and note (4), we can get:

$$\dot{V}_1 = e_w \dot{e}_w = -e_w[p(T_e - T_{L)} - B_n w]/J \ . \tag{5}$$

In order to guarantee $\dot{V}_1 \le 0$, we select the following control input:

$$T_e^{*} = \tfrac{1}{P}(B_n w + k_w J e_w) + T_L \ , \qquad k_w > 0 \tag{6}$$

Then, $\dot{V}_1 = -k_w e_w^2 \le 0$.

However, the load torque is unknow and it need to be estimated adaptively. Then, the speed controller output becomes:

$$T_e^{*} = \tfrac{1}{P}(B_n w + k_w J e_w) + \hat{T}_L \ , \qquad k_w > 0 \tag{7}$$

Where $\hat{T}_L$ is the estimated load torque.

Substituting (7) into (4), the speed error dynamics becomes:

$$\dot{e}_w = [\tilde{T}_L - k_w J e_w]/J \ . \tag{8}$$

where $\tilde{T}_L = T_L - \hat{T}_L$.

Step 2: Flux and torque controller design
Define the flux and torque error as follows:

$$\begin{cases} e_\Psi = \Psi^* - \Psi \\ e_T = T_e^* - T_e \end{cases}. \tag{9}$$

Note (1)-(3), we can get that:

$$\dot{e}_\Psi = -\dot{\Psi} = -2(\psi_\alpha u_\alpha + \psi_\beta u_\beta - R_s \psi_\alpha i_\alpha - R_s \psi_\beta i_\beta) \tag{10}$$

$$\dot{e}_T = \dot{T}_e^* - \dot{T}_e = \frac{B_n - k_w J}{JP}[p(T_e - T_L) - B_n w]$$

$$-\tfrac{3}{2}P[u_\alpha(i_\beta - \frac{\psi_\beta}{L_d}) - u_\beta(i_\alpha - \frac{\psi_\alpha}{L_q}) - \frac{\psi_\alpha}{L_q}(R_s i_\beta + E_\beta) + \frac{\psi_\beta}{L_d}(R_s i_\alpha + E_\alpha)] \tag{11}$$

Define the Lyapunov function: $V_2 = V_1 + \frac{1}{2}(e_\Psi^2 + e_T^2 + \frac{1}{\gamma}\tilde{T}_L)$

Take the derivative of the Lyapunov function $V_2$, and note (5), (8), (10)-(11), we can get:

$$\dot{V}_2 = \dot{V}_1 + e_T \dot{e}_T + e_\Psi \dot{e}_\Psi - \tfrac{1}{\gamma}\dot{\hat{T}}_L \tilde{T}_L$$
$$= \frac{e_w}{J}(\tilde{T}_L - k_w J e_w) - 2e_\Psi(\psi_\alpha u_\alpha + \psi_\beta u_\beta - R_s \psi_\alpha i_\alpha - R_s \psi_\beta i_\beta) +$$
$$e_T\left\{\frac{B_n - k_w J}{JP}[p(T_e - T_L) - B_n w] - \tfrac{3}{2}P[u_\alpha(i_\beta - \frac{\psi_\beta}{L_d}) - u_\beta(i_\alpha - \frac{\psi_\alpha}{L_q}) - \frac{\psi_\alpha}{L_q}(R_s i_\beta + E_\beta) + \frac{\psi_\beta}{L_d}(R_s i_\alpha + E_\alpha)]\right\}$$
$$-\tfrac{1}{\gamma}\dot{\hat{T}}_L \tilde{T}_L$$

In order to guarantee $\dot{V}_2 \le 0$, we select the following control inputs:

$$\begin{cases} u_\alpha = E\left[\frac{2\psi_\beta}{3P}\dot{T}_e^* + \frac{\psi_\alpha \psi_\beta}{L_q}(R_s i_\beta + E_\beta) - \frac{\psi_\beta^2}{L_d}(R_s i_\alpha + E_\alpha) + f_\alpha\right] \\ u_\beta = -E\left[\frac{2\psi_\alpha}{3P}\dot{T}_e^* - \frac{\psi_\alpha \psi_\beta}{L_q}(R_s i_\alpha + E_\alpha) + \frac{\psi_\alpha^2}{L_d}(R_s i_\beta + E_\beta) + f_\beta\right] \end{cases}. \tag{12}$$

Where,

$$\begin{cases} f_\alpha = \frac{2\psi_\beta}{3P}k_T e_T + (i_\alpha - \frac{\psi_\alpha}{L_q})(R_s \psi_\alpha i_\alpha + R_s \psi_\beta i_\beta + \frac{1}{2}k_\Psi e_\Psi) \\ f_\beta = \frac{2\psi_\alpha}{3P}k_T e_T - (i_\beta - \frac{\psi_\beta}{L_q})(R_s \psi_\alpha i_\alpha + R_s \psi_\beta i_\beta + \frac{1}{2}k_\Psi e_\Psi) \\ E = (\psi_\alpha i_\beta + \psi_\beta i_\alpha - \frac{2\lambda}{L_q})^{-1} \end{cases}$$

And the adaptive update law is: $\dot{\hat{T}} = \frac{\gamma}{J}\left[e_w - (B_n - k_w J\right]$, $k_\Psi > 0$, $k_T > 0$.

Then, $\dot{V}_2 = -k_w e_w^2 - k_T e_T^2 - k_\Psi e_\Psi^2 \le 0$.

Hence, according to Lyapunov stability theory, the designed controller is asymptotically stable.

## 3. Simulation Results

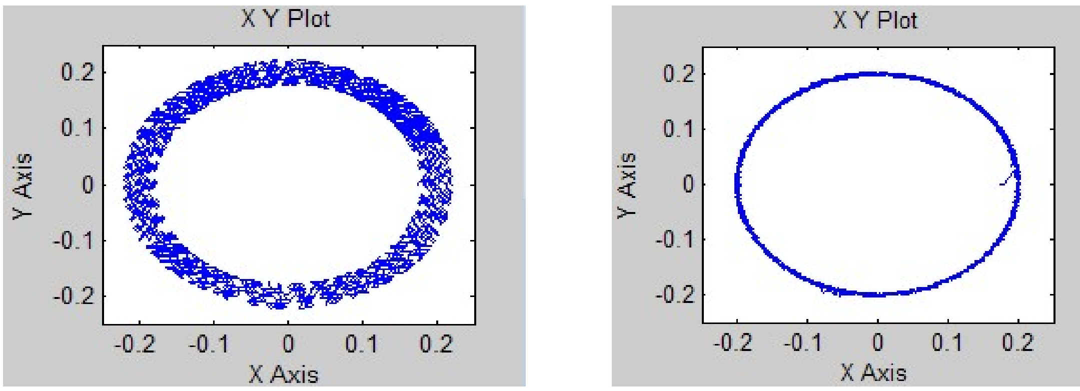
To study on the effectiveness of the proposed scheme, simulations of the traditional DTC-hysteresis-based (CDTC) and DTC-backstepping-based (BPDTC) PMSM drive is performed using a Matlab/Simulink package. The parameters for the PMSM drive system are as shown in Table 1. Chosen algorithm ode4 with fixed-step size 1e-5 and given flux linkage is $\psi_f = 0.2Wb$. In the CDTC, the hysteresis band of flux linkage controller is selected $\Delta\Psi = 0.02Wb$ and the hysteresis band of torque controller is selected as $\Delta T_e = 0.4Wb$. All results are obtained from one opertation point of 1000r/min without load, and the load is changed to 8N.m when t=0.03s.

The control parameters are selected as follows: $k_p = 0.2$ and $k_i = 0.02$ in CDTC; $k_w = 5000$, $k_\Psi = 600$ and $k_T = 2000$ in BDDTC. The simulation results are shown in Figure 2-3.

Table 1. Parameters of SPMSM used in this Paper

| | |
|---|---|
| Stator resistance $R_s$ | 2.875 $\Omega$ |
| d-axis inductance $L_d$ | 8.5e-3 H |
| q-axis inductance $L_q$ | 8.5e-3 H |
| Magnet flux linkage $\psi_f$ | 0.2 Wb |
| Number of pole pairs P | 2 |
| Inertia J | 8e-4 kg.m^2 |
| Friction factor | 1e-4 N.m.s |

It can be seen from Figure 2 that the flux linkage response are very fast in both CDTC and BPDTC, and it takes about 0.0005s to reach its amplitude 0.2Wb. Then, the flux linkage keeps constant throughout the running. But it is obvious that BPDTC has smaller amplitude fluctuation than CDTC. The amplitude of flux linkage changes from 0.196WB to 0.204Wb using BPDTC,but is changes from 0.178Wb to 0.224Wb for CDTC.




(a) (b)

Figure 2. The Stator Flux Linkage Waveforms of the CDTC (a) and BPDTC (b) for SPMSM

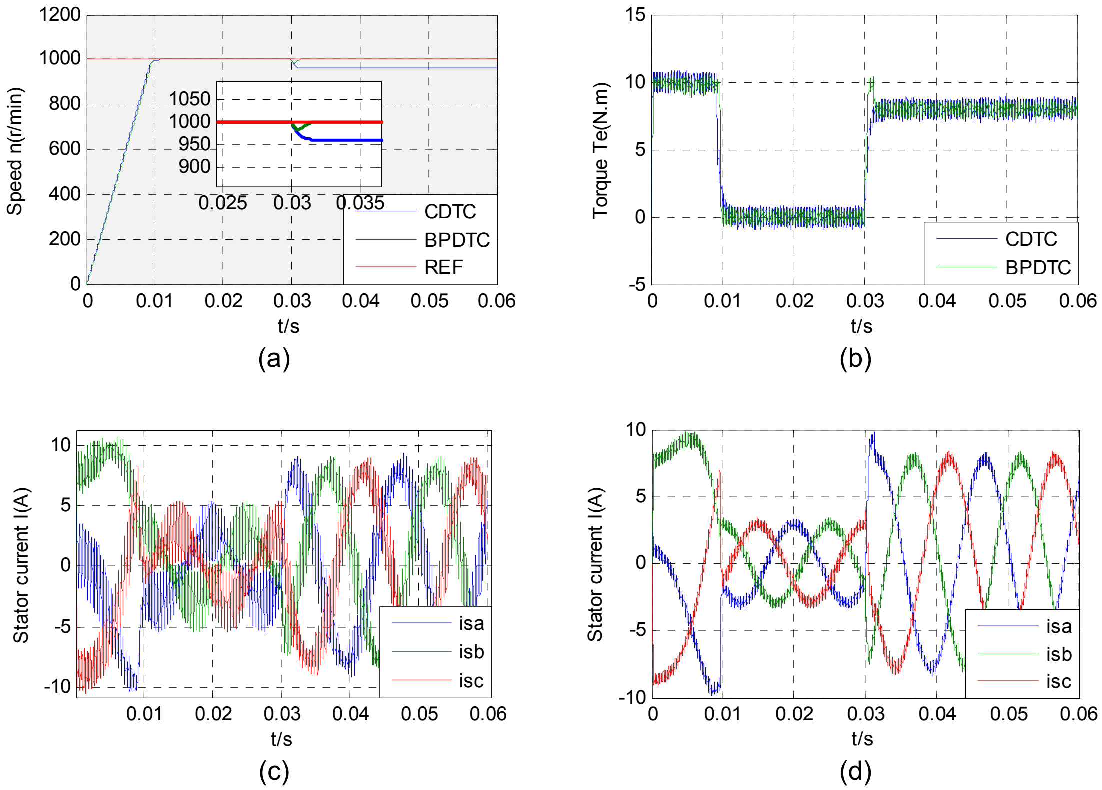



Figure 3. Response Curve of CDTC and BPDTC for SPMSM: a) speed, b) electromagnetic torque, c) currents of CDTC d) currents of BPDTC

Figure 3 shows the state trajectories of speed, electromagnetic torque, and stator current for the mothods of CDTC and BPDTC. We can see from Figure 3(a) that the presented method has better robustness than CDTC. It has only a slight perturbation and then quickly restored to its reference speed (taking about 0.002s. but CDTC has about 40r/min speed error and it need long time to restored to its reference speed. Figure 3(b)-(d) show that the proposed method has less torque and current perturbations than traditional ones (CDTC).

Form obove simulation results, it is clearly shown that the proposed control scheme is superior than traditional ones.

## 4. Conclusion

We develop a novel direct torque control scheme that accounts for loaduncertainty in a SPMSM drive system. This controller is designed based on backstepping adaptive theory. The advantages of the proposed controller are as follows:

1) It has strong robustness for uncertain load;
2) The flux linkage, torque, and current ripples are significantly reduced.
3) SVM scheme is used to get constant switching frequency.

Future research should investigate the implementation of the proposed control scheme by using an experimental setup.

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