JPE 18-5-8

https://doi.org/10.6113/JPE.2018.18.5.1369
ISSN(Print): 1598-2092 / ISSN(Online): 2093-4718

Model Predictive Torque Control of Surface Mounted Permanent Magnet Synchronous Motor Drives with Voltage Cost Functions

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Abstract

In this paper, a model predictive torque control (MPTC) without the use of a weighting factor for surface mounted permanent-magnet synchronous machine (SPMSM) drive systems is presented. Firstly, the desired voltage vector is predicted in real time according to the principles of deadbeat torque and flux control. Then the sector of this desired voltage vector is determined. The complete enumeration for testing all of the feasible voltage vectors is avoided by testing only the candidate vectors contained in the sector. This means that only two voltage vectors in the sector need to be tested for selecting the optimal voltage vector in each control period. Thus, the calculation time can be reduced when compared with the conventional enumeration method. On the other hand, a novel cost function that only includes the dq-axis voltage errors between the desired voltage and candidate voltage is designed to eliminate the weighting factor used in the conventional MPTC. Thus, the control complexity caused by the tuning of the weighting factor is effectively decreased when compared with the conventional MPTC. Simulation and experimental investigation have been carried out to verify the proposed method.

Key words: Computation burden, Predictive control, Torque control, Weighting factor

I. INTRODUCTION

The direct torque control (DTC) method is widely applied in PMSM systems. The applied voltage vector of the DTC method is obtained according to a switching table, which is established based on the error signs of both the torque and the flux. It has a very fast dynamic response by directly controlling the stator flux magnitude and torque. However, disadvantages, i.e., torque fluctuations and an unfixed switching frequency, obviously exist in the conventional DTC method [1], [2]. As a possible alternative control strategy, model predictive torque control (MPTC) has recently gained a lot of attention. In the MPTC method, the torque and stator flux at the next instant are predicted based on a discrete model of the motor. Then the optimal voltage vector is determined in each control period according to the optimization of an operating cost function. The selected optimal voltage vector, which is one of the seven basic voltage vectors and can minimize the torque and flux errors, is applied to the motor by an inverter [3]-[5]. The major advantages of MPTC include its intuitive concept, straightforward implementation, good capability in terms of handling the constraints of system states and the fact that it does not require a vector modulation strategy. However, this method has high computational burden on the system hardware, which is undesirable in the real-time implementations. This is especially true in the case of multilevel converters and long horizon prediction, where the calculation time rises exponentially [6], [7].

Manuscript received Jul. 25, 2017; accepted Mar. 17, 2018
Recommended for publication by Associate Editor M. P. Kazmierkowski.
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To solve this problem, a sphere decoding algorithm that can effectively reduce the computational burden is adopted in [8] and [9] to implement long horizon predictions. In order to implement a low calculation burden model predictive control, the binary search tr ee is introduced in [10]. To reduce the calculation effort and to extend the prediction horizon, a heuristic voltage vector pre-selection method, which uses the multiparametric programming from the Multiparametric Toolbox of Matlab software, is introduced in [11]. In addition, the results show the superior performance of the control system during transients and in the steady state.

In addition, in order to control both the torque and the stator flux simultaneously, an appropriate weighting factor of the conventional MPTC needs to be carefully designed, since the weighting factor selection is directly related to the control performance of the whole system. The existing design strategies for weighting factors usually require an exhaustive search, which is time consuming, and not intuitive [13], [14]. To improve the design process of the weighting factor, some methods have been proposed. In [15], a fuzzy decision-making strategy is introduced into the model predictive control of a matrix converter to adjust the value of the weighting factor. According to the basic principle of torque ripple minimization, an online weighting tuning method is presented in [16] to improve speed control performance. Although the aforementioned methods to find appropriate weighting factors have been reported, there is no universal theory to guide weighting factor design. Therefore, some model predictive control methods without a weighting factor have been presented. In [16], the control of the torque and stator flux are replaced by voltage control to avoid the weighting factor. However, the main optimization objective is not clear due to the lack of a theory analysis. In order to avoid the selection process of the weighting factors, a torque error based cost function and a stator flux error based cost function are established in [17]. Then, the voltage vector, which minimizes the sum of two cost functions, is selected as the optimal vector. Recently, a new duty cycle control method has been proposed in [18] to reduce the torque and flux ripples of the conventional direct torque control. In order to eliminate the weighting factor between the torque and stator flux, the torque is considered as the main control goal of the system while the stator-flux is included in the cost function as a constraint. This means that the stator-flux can be controlled within a limited hysteresis band.

In this paper, an improved MPTC algorithm for PMSM drives is presented, in which the computation burden reduction and weighting factor elimination are implemented simultaneously. Firstly, the desired voltage vector is predicted in real time according to deadbeat control technology. In addition, the sector position of the predicted desired voltage vector is used to optimize the selection range of where the candidate vector can be located. The optimal vector for minimizing the cost function can be selected from the vectors contained in the sector. This means that the selection range of the optimal vector can be narrowed down to two vectors, which avoids evaluating all of the feasible vectors (six nonzero vectors and two null vectors). Thus, the computation burden is reduced when compared with that of the conventional MPTC. On the other hand, a novel cost function is proposed to eliminate the complicated weighting factor design in the conventional MPTC. In this cost function, d-axis and q-axis voltage errors are adopted to replace the torque/flux error. In addition, the relationship between the proposed cost function and conventional cost function is derived, which shows that the control objective for both of the cost functions is the same. Therefore, the control complexity caused by the weighting factor is significantly reduced. Finally, simulations and experiments are carried out to verify the proposed method.

II. SPMSM Mathematical Model

The model of a SPMSM is given as [20],[21]:

$$\begin{cases} u_d = R_s i_d - \omega\psi_q + \dfrac{d\psi_d}{dt} \\ u_q = R_s i_q + \omega\psi_d + \dfrac{d\psi_q}{dt} \end{cases} \tag{1}$$

$$\begin{cases} \psi_d = L_d i_d + \psi_f \\ \psi_q = L_q i_q \end{cases} \tag{2}$$

$$T_e = 1.5 p \psi_f i_q \tag{3}$$

where:

u_d, u_q are the d- axis and q-axis stator voltages, respectively;
i_d, i_q are the d- axis and q-axis stator currents, respectively;
ψ_d, ψ_q are the d- axis and q-axis flux linkage, respectively;
L_d=L_q=L is the stator inductance;
R is the stator resistance;
ψ_f is the permanent magnet flux linkage;
ω is the angular velocity;
p is the number of pole pairs.

III. Conventional Model Predictive Torque Control

The basic control structure used in the conventional MPTC is displayed in Fig. 1(a). The torque and flux command is obtained based on the outer speed control loop, in which the flux command is obtained from the MTPA equation, namely, $\psi_s^* = \sqrt{\psi_f^2 + \left[L_q T_e^* / \left(1.5 p \psi_f\right)\right]^2}$. In digital implementation, a one-step control delay exists in the MPTC, which influences

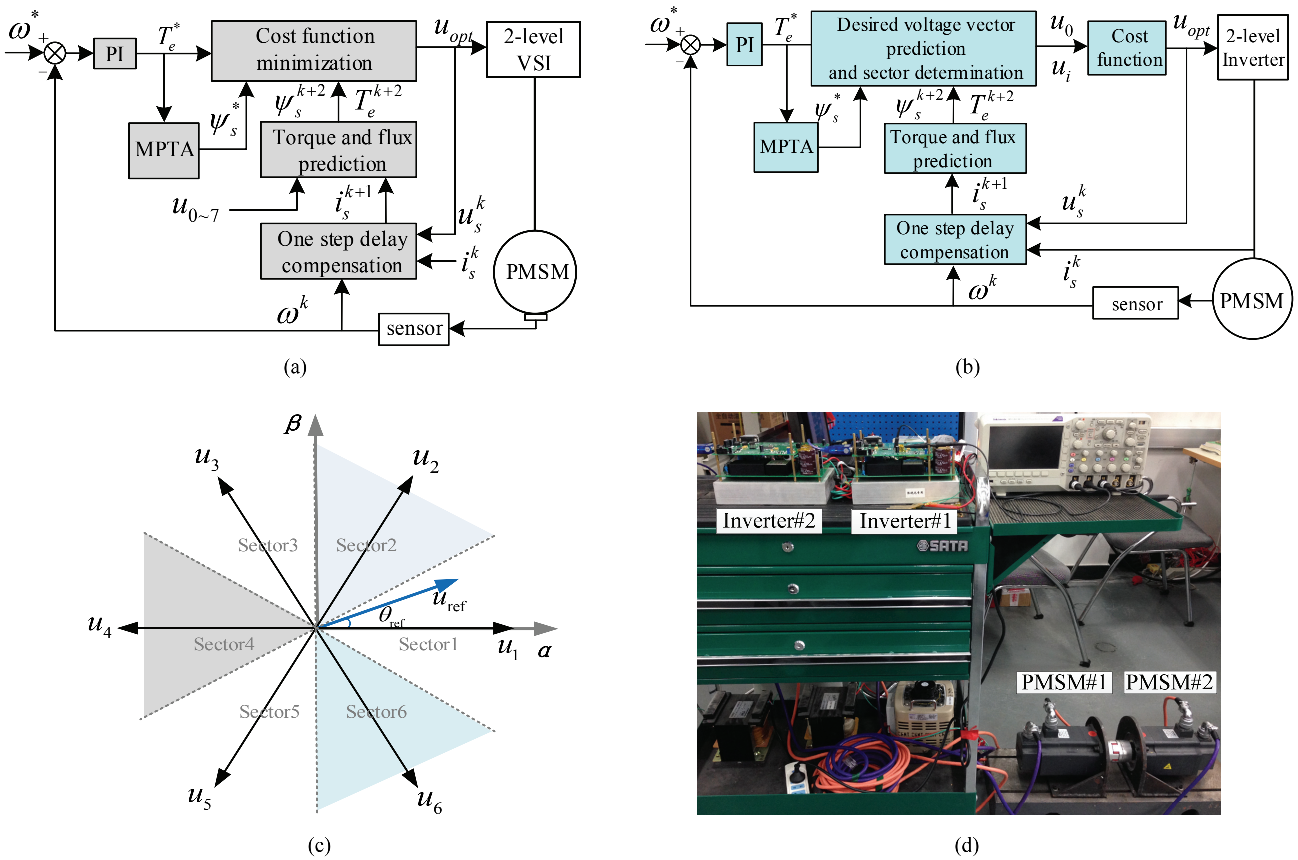


Fig. 1. Images of the: (a) Structure of the conventional MPTC, (b) Control structure of the proposed MPTC method, (c) Sector of the voltage vector, (d) Experiment platform of the control system.

the control performance of the digital control system. Thus, the compensation equation of the one-step control delay shown in [21] and [22] is adopted to predict the current values of the next control period and to replace the sampled currents i_d^k and i_q^k.

The conventional cost function, which contains the prediction errors of the torque and stator flux, is expressed as follows:

$$g = \left|T_e^* - T_e^{k+2}\right| + w\left|\psi_s^* - \left|\psi_s^{k+2}\right|\right| \tag{4}$$

Generally, in order to balance the control importance of the stator flux and the electromagnetic torque, the weighting factor w used in (4) should be designed according to the following expression [23]:

$$w = T_N / \left|\psi_{sN}\right| \tag{5}$$

where T_N is the rated torque, and ψ_{sN} is the rated stator flux amplitude. However, in most cases, the control responses of the torque and the stator flux are not satisfactory when the weighting factor is designed based on (5). Therefore, a lot of tuning work for the weighting factor is necessary by the way of simulations or experiments. If the weighting factor w in (4) is too high, which means that the control ability of the stator flux is more important than the control ability of the torque, the tracking performance of the torque performs poorly. Conversely, if the weighting factor w is selected too small, the control capability of the torque is more important when compared with the control ability of the stator flux. Under this control condition, fluctuations exist in the amplitude of the stator flux and the stator currents.

IV. Model Predictive Torque Control Without the Weighting Factor

According to above analysis, it can be found that the conventional method needs to test all of the feasible voltage vectors to select the optimal vector at every control period, which means lots of computations must be completed in one sampling interval. However, in order to improve the control performance of the MPTC, the sampling frequency generally needs to be selected so that it is as high as possible. Therefore, further practical application of the MPTC method is limited due to the dilemma between a large volume of computations and a short control period. In addition, since the stator flux and electromagnetic torque have different units and dimensions, the weighting factor of the cost function in the conventional method is required to balance the control between the torque and the stator flux. However, the design of the weighting factor is complicated, because tuning the weighting factor for optimum control performance is an iterative process.

To solve the above mentioned problems, an improved MPTC method, which eliminates the weighting factor and reduces the computation burden of the conventional MPTC, is proposed in this section. The control structure of the proposed method is displayed in Fig.1(b).

A. DB-DTC Based Desired Voltage Vector Prediction

In this paper, the DB-DTC algorithm is adopted to calculate the desired voltage vector, which ensures the convergence of the torque and stator flux error in one control period and is considered to be the desired voltage vector to reduce the range of the tested voltage vector. This means that the number of the candidate voltage vector is reduced. The concrete calculation procedure is presented as bellow.

The one-step delay compensation method reported in [21] and [22] is used to predict the current values of the next control period. Then the measured currents of the model (1), (2) and (3) are replaced. At this point, the following equation can be obtained according to (1) and (2):

$$\begin{cases} \psi_d^{k+2} = u_d^{k+1}T_s + \psi_d^{k+1} + \omega T_s \psi_q^{k+1} - \dfrac{R_s T_s}{L_d}(\psi_d^{k+1} - \psi_f) \\ \psi_q^{k+2} = u_q^{k+1}T_s + \psi_q^{k+1} - \omega T_s \psi_d^{k+1} - \dfrac{R_s}{L_q}\psi_q^{k+1}T_s \end{cases} \tag{6}$$

Then the torque equation (7) can be obtained by substituting the q-axis stator flux of (2) into (3).

$$T_e = \frac{3}{2} p\psi_f \frac{\psi_q}{L_s} \tag{7}$$

Taking the time derivative of the torque yields:

$$T_e^{k+2} - T_e^{k+1} = \frac{3}{2} p \frac{\psi_f}{L_s}\left(\psi_q^{k+2} - \psi_q^{k+1}\right) \tag{8}$$

Then substituting (6) into (8), the torque discrete equation can be obtained as [12], [24]:

$$T_e^{k+2} - T_e^{k+1} = \frac{3p\psi_f}{2L_s}\left(u_q^{k+1}T_s - \frac{R_s T_s \psi_q^{k+1}}{L_s} - \omega T_s \psi_d^{k+1} \right) \tag{9}$$

In addition, the relation equation between the stator flux and the voltage can be expressed as follows:

$$\begin{aligned} (\psi_s^{k+2})^2 &= (\psi_d^{k+2})^2 + (\psi_q^{k+2})^2 \\ &= (u_d^{k+1}T_s + \psi_d^{k+1} + \omega T_s \psi_q^{k+1})^2 + (u_q^{k+1}T_s + \psi_q^{k+1} - \omega T_s \psi_d^{k+1})^2 \end{aligned} \tag{10}$$

According to the basic principle of the deadbeat control, the torque and stator flux in the next period are used as reference commands, i.e., $T_e^{k+2} = T_e^*$ and $\psi_s^{k+2} = \psi_s^*$. Then based on equations (9) and (10), the desired voltage vector can be predicted as follows:

$$\begin{cases} u_d^{k+1} = \dfrac{-X_1 + \sqrt{X_1^2 - X_2}}{T_s} \\ u_q^{k+1} = \dfrac{B}{T_s} \end{cases} \tag{11}$$

where:

$$\begin{aligned} X_1 &= \psi_d^{k+1} + \omega \psi_q^{k+1} T_s \\ X_2 &= B^2 + 2B(\psi_q^{k+1} - \omega \psi_d^{k+1} T_s) + (\psi_d^{k+1})^2 + (\psi_q^{k+1})^2 \\ &\quad + \omega^2 T_s^2 [(\psi_d^{k+1})^2 + (\psi_q^{k+1})^2] - (\psi_s^*)^2 \\ B &= \frac{2L_q}{3p\psi_f}(T_e^* - T_e^{k+1}) + \frac{R_s T_s \psi_q^{k+1}}{L_q} + \omega T_s \psi_d^{k+1}. \end{aligned}$$

In order to obtain the phase angle of the desired voltage vector, the voltage vector obtained by (11) needs to be transformed to the α-β frame using the following equation [19]:

$$u_{ref} = \begin{pmatrix} u_\alpha^{\ k+1} \\ u_\beta^{\ k+1} \end{pmatrix} = \begin{pmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{pmatrix} \begin{pmatrix} u_d^{k+1} \\ u_q^{k+1} \end{pmatrix} \tag{12}$$

Based on (12), the phase angle of the desired voltage vector can be obtained by:

$$\theta_{\mathrm{ref}} = \arctan\left(\frac{u_\beta^{k+1}}{u_\alpha^{k+1}} \right). \tag{13}$$

B. Cost Function Design and Optimal Vector Selection

In order to quickly select the optimal voltage vector, the entire α-β plane is divided into six sectors, as shown as Fig. 1 (c). It can be seen that two basic vectors are contained within each sector. The two vectors are an active voltage vector and a null voltage vector. Then quickly judging the sector of the desired voltage vector can be implemented based on the angle calculation equation (13). Therefore, the candidate voltage vectors that need to be tested by the cost function are determined. Table I lists the relationship between the sector of the desired voltage vector and the candidate voltage vector. For example, if the desired voltage vector (u_{ref}) is located in sector 1, as shown in Fig. 1(c), the active vector u_1 and the null vectors are the candidate voltage vectors to be evaluated. The optimal vector is selected from u_1 and the null vector by minimizing the cost function. This means that only two voltage vectors need to be evaluated by the cost function in each control period, which avoids the evaluation of all of the voltage vectors. Therefore, the calculation burden is reduced when compared with the conventional MPTC method.

After determining the range of the candidate voltage vector, the evaluation of the candidate voltage vector by the cost function to select the optimal voltage vector is important. In this paper, a novel cost function, which contains the d-axis voltage error and the q-axis voltage error, is proposed as below:

TABLE I
RELATIONSHIP BETWEEN THE SECTOR OF THE DESIRED VOLTAGE VECTOR AND THE CANDIDATE VOLTAGE VECTOR

Sector location	one	two	three	four	five	six
Candidate Vectors	u_0,u_1	u_7,u_2	u_0,u_3	u_7,u_4	u_0,u_5	u_7,u_6
Vector selection	000, 100	111, 110	000, 010	111, 011	000, 001	111, 101

$$g = \left|u_{dref} - u_{di}\right| + \left|u_{qref} - u_{qi}\right| + I_{pro} \tag{14}$$

where $u_{\rm dref}$ and $u_{\rm qref}$ are d-axis and q-axis components of the desired voltage vector $u_{\rm ref}$; and $u_{\rm di}$ and $u_{\rm qi}$ represent d-axis and q-axis components of the candidate voltage vectors $u_{\rm i}$, which are selected according to the sector of the desired voltage vector and Table I. From (14), it can be seen that the weighting factor is eliminated due to the same dimension between the d-axis voltage and the q-axis voltage. In addition, the current limitation $I_{\rm pro}$ is included in (14) to limit the overcurrent problem. Its expression is shown as follows:

$$I_{\rm pro} = \begin{cases} 0 & (\left|i(k+2)\right| \le \left|i_{\max}\right|) \\ \infty & (\left|i(k+2)\right| > \left|i_{\max}\right|) \end{cases} \tag{15}$$

If the predicted current corresponding to the tested vector is less than the current limitation, $I_{\rm pro}$ is equal to zero, which means that overcurrent does not occur using this tested vector. On the other hand, if the currents of the tested voltage vector are over the current limit, $I_{\rm pro}$ is infinite.

According to the motor model, i.e., equation (1), (2) and (3), the following equations can be obtained:

$$\begin{aligned} \left|u_{\rm dref} - u_{di}\right| &= \left|\frac{d\psi_d^*}{dt} - \omega\psi_q^* - \frac{d\psi_d^i}{dt} + \omega\psi_q^i\right| \\ &= \left|\frac{d}{dt}\left(\psi_d^* - \psi_d^i\right) - \omega\left(\psi_q^* - \psi_q^i\right)\right| \end{aligned} \tag{16}$$

$$\begin{aligned} \left|u_{\rm qref} - u_{qi}\right| &= \left|Ri_q^* + \frac{d\psi_q^*}{dt} + \omega\psi_d^* - Ri_q^i - \frac{d\psi_q^i}{dt} - \omega\psi_d^i\right| \\ &= \left|\frac{2R}{3p\psi_f}T_e^* + \frac{d\psi_q^*}{dt} + \omega\psi_d^* - \frac{2R}{3p\psi_f}T_e^i - \frac{d\psi_q^i}{dt} - \omega\psi_d^i\right| \\ &= \left|\frac{2R}{3p\psi_f}\left(T_e^* - T_e^i\right) + \frac{d}{dt}\left(\psi_q^* - \psi_q^i\right) + \omega\left(\psi_d^* - \psi_d^i\right)\right| \end{aligned} \tag{17}$$

where $T_{\rm e}^{i}$ is predictive torque of the candidate voltage vector $u_{\rm i}$; and $\psi_{\rm d}^{i}$ and $\psi_{\rm d}^{i}$ are the predictive flux of candidate voltage vector $u_{\rm i}$. Therefore, the cost function can be rewritten as follows:

$$\begin{aligned} g =& \left|\frac{d}{dt}\left(\psi_d^* - \psi_d^i\right) - \omega\left(\psi_q^* - \psi_q^i\right)\right| \\ &+ \left|\frac{2R}{3p\psi_f}\left(T_e^* - T_e^i\right) + \frac{d}{dt}\left(\psi_q^* - \psi_q^i\right) + \omega\left(\psi_d^* - \psi_d^i\right)\right| + I_{pro} \end{aligned} \tag{18}$$

Then, the discrete form of (18) can be obtained:

$$\begin{aligned} g =& \left|\begin{aligned}&\frac{\psi_d^*(k+2) - \psi_d^i(k+2)}{T_s} - \frac{\psi_d^*(k+1) - \psi_d^i(k+1)}{T_s} \\ &-\omega(k)\left[\psi_q^*(k+1) - \psi_q^i(k+1)\right]\end{aligned}\right| + \\ &\left|\begin{aligned}&\frac{2R}{3p\psi_f}\left[T_e^*(k+1) - T_e^i(k+1)\right] + \frac{\psi_q^*(k+2) - \psi_q^i(k+2)}{T_s} \\ &-\frac{\psi_q^*(k+1) - \psi_q^i(k+1)}{T_s} + \omega(k)\left[\psi_d^*(k+1) - \psi_d^i(k+1)\right]\end{aligned}\right| + \end{aligned} \tag{19}$$

From equations (18) and (19), it is obvious that the minimizing equation (19) necessarily minimizes the stator-flux error and the torque error. In addition, it can be found that this cost function is similar to the conventional cost function (4) with a weighting factor, since they are both torque error and flux error based cost functions. However, the proposed cost function of this paper no longer needs a weighting factor. The control balance between the torque and the flux is automatically implemented according to the present operating condition, which means that the tedious work of tuning the weighting factor in the conventional MPTC can be avoided.

According to the above analysis, it can be found that the core ideas of the conventional method and the proposed method are the same, i.e., selecting an optimal voltage vector. However the implementation method is different for both of the methods. In the conventional method, the indirect voltage selection mode is adopted in the cost function, which means that the flux error and torque error are used as criterion to select the best vector. On the other hand, in the proposed method, the direct voltage selection mode is presented to construct the cost function. The best voltage vector is selected by directly comparing the desired voltage vector with the candidate voltage. The important advantage of this direct voltage selection mode is that the tedious work of tuning the weighting factor in the conventional MPTC can be avoided.

V. SIMULATION AND EXPERIMENTAL RESULTS

Simulations of the conventional MPTC and the proposed MPTC have been carried out in this section by MATLAB/ Simulink. In addition, the conventional MPTC method and the proposed MPTC method have been verified on a SPMSM drive system based on a floating point digital signal processor 28335. The experimental setup is shown in Fig. 1(d). The PMSM parameters and sampling frequency are listed in Table II.

A. Simulation Results

Fig. 2(a) shows the control performances under the control of the conventional MPTC method when different weighting factors are selected. If the weighting factor w is designed according to (5), i.e., w=20.25, the control performance of the conventional MPTC is not satisfactory, as shown in Fig. 2(a). However, the control performance becomes better and better

TABLE II
PARAMETERS OF A PMSM

Sampling frequency	10kHz
Number of pole pairs	p =3
Stator phase resistance	R =3Ω
Rated speed	n_N =2000r/min
Rotational inertia	J =0.00129kg·m2
Flux linkage	ψ_f =0.295Wb
d-axis and q-axis inductances	L_d =L_q =11mH

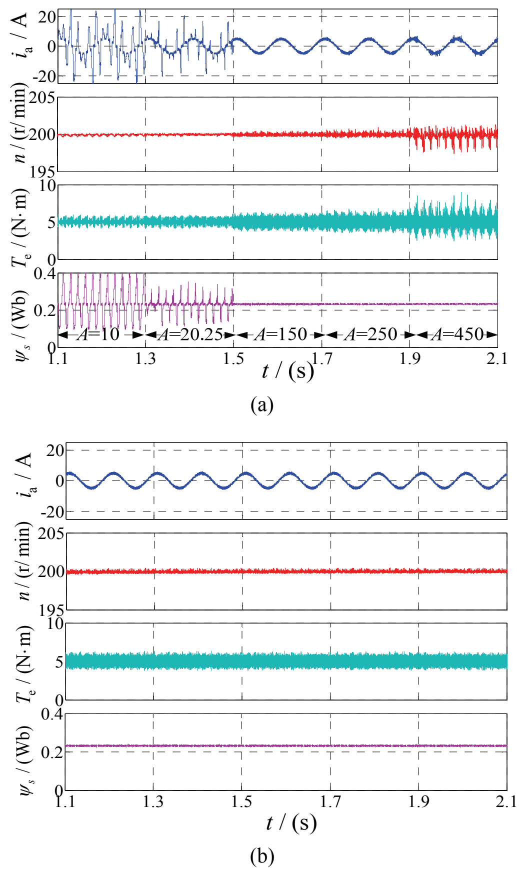


Fig. 2. Simulation results at a speed of 200r/min with a 50% rated load: (a) Conventional MPTC method under different weighting factors, (b) Proposed MPTC method.

as w increases. The best performance can be obtained when the weighting factor is 150. It should be noted that the control performances became bad when w further increases to 450. These results show that different values of the weighting factor have different control performances in the conventional MPTC method. Therefore, in order to gain satisfactory control performance, tuning work of the weighting factor is necessary. When compared with the conventional MPTC, the control performances of the proposed MPTC are shown in Fig. 2(b). It can be seen that the control performances are similar for the proposed MPTC and the conventional MPTC with the optimal weighting factor 150. However, the weighting factor in the conventional method is eliminated and the

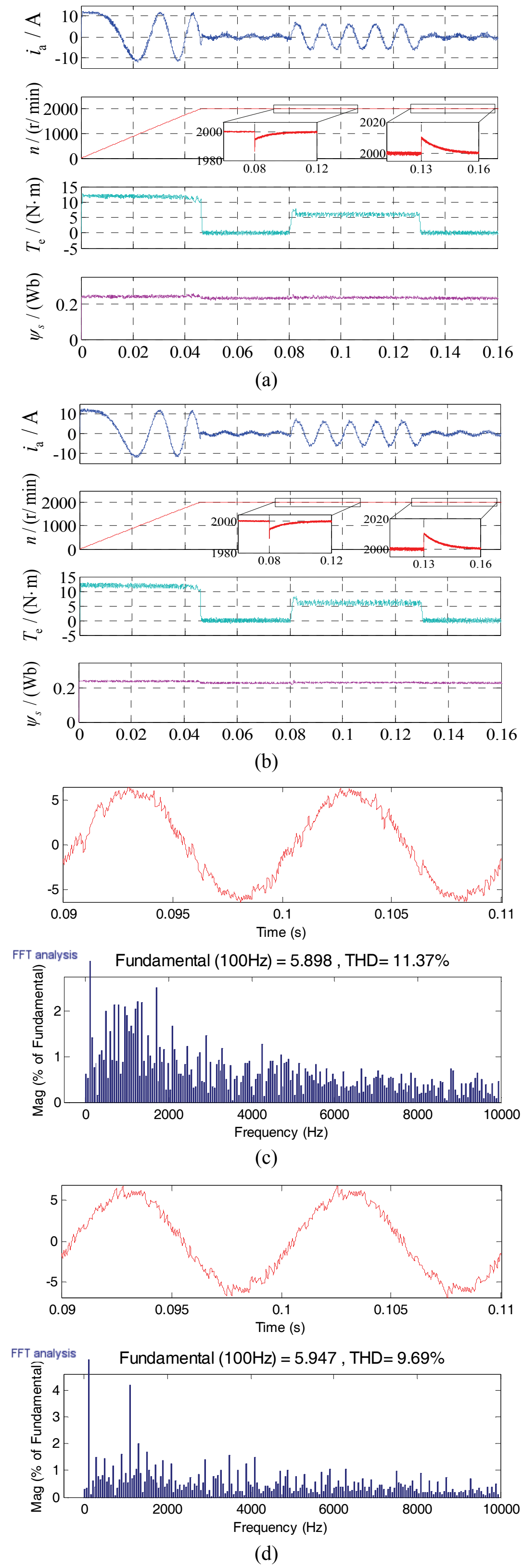


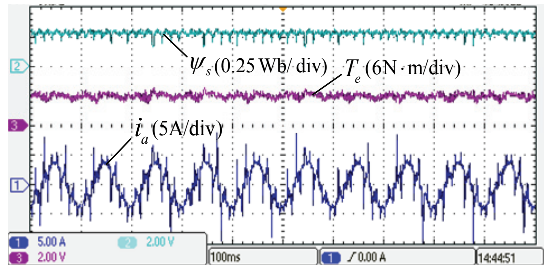
Fig. 3. Robustness test to external load disturbances and the harmonic spectrum of the stator current with the rated speed and rated torque: (a) Dynamic response of the conventional MPTC, (b) Dynamic response of the proposed MPTC, (c) THD of the conventional MPTC, (d) THD of the proposed MPTC.

tuning weighting factor is also avoided in the proposed MPTC, which is an advantage in the practical applications.

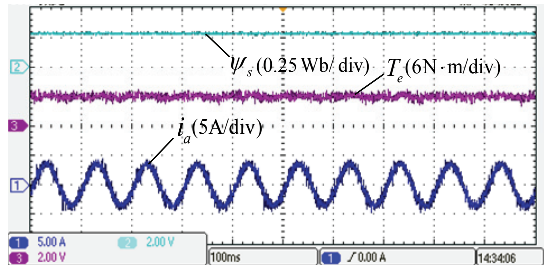
In order to further compare both methods, the dynamic responses and the current THD analysis under the rated load and rated speed are displayed in Fig. 3. In Fig. 3(a)-(b), the motor speed command is step changed from zero to the rated value. Then in order to test the dynamic performance, the load torque of the motor is suddenly changed from zero to the rated value at 0.08 seconds. Then this rated load torque is removed at 0.13 seconds. Additionally, the THD analyses of the steady state current under the control of both methods are shown in Fig. 3(c)-(d). From these simulation results, it is seen that the proposed MPTC has dynamic and steady-state control performances that are similar to those of the conventional method under an optimal weighting factor.

B. Experimental Results

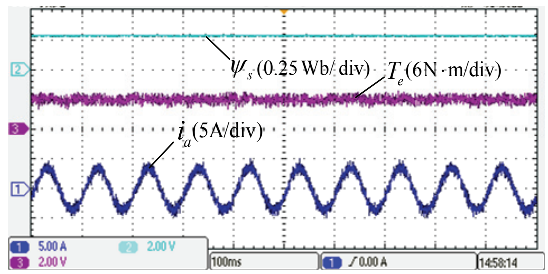
In the conventional MPTC, the tuning of the weighting factor, which is usually based on error-and-trial measures, is a time consuming process. Generally, the weighting factor w can be designed according to (5) to balance the control importance of the stator flux and the electromagnetic torque, which is 20.25 in this paper. However, when the weighting factor is designed based on (5), the control responses of the torque and the stator flux are not satisfactory according to the experimental results in Fig. 4. Fig. 4 shows the control responses of the conventional MPTC with different weighting factors (from 20.25 to 350) when the motor operates at a low speed of 200r/min. In addition, the phase currents shown in Fig. 4 are selected as the target of the THD analysis, and the analysis results are listed in Table III. It can be seen that when the weighting factor is 20.25, the fluctuation of the stator flux is 0.11Wb, the torque fluctuation is 0.85N.m and the current THD is 53.79%. If the weighting w is increased to 50, the stator flux ripple is reduced by 29%, from 0.11Wb to 0.023Wb. However, the ripple of the torque is increased by 44.7%, from 0.85N.m to 1.23N.m, and the current THD is reduced by 30.36%, from 53.79% to 23.43%. When the weighting factor varies from 50 to 150, the ripples of the stator flux are further reduced from 0.023 to 0.02, and the torque ripple is not increased. Instead, it is reduced to 1.2N.m. In addition, the current THD is further reduced from 23.43% to 19.91%. This means that a good control balance between the torque ripples and the flux ripples is achieved. Next, if a larger value is selected as the weighting factor, i.e., w=350, noticeable changes of the stator flux ripple cannot be seen. However, the ripples of the electromagnetic torque and the current THD clearly become higher. Experimental results indicate that good tuning of the weighting factor is necessary to balance the control performance of the electromagnetic torque and stator flux for the conventional method. In addition, the optimal weighting factor of this machine is 150 for the conventional MPTC.



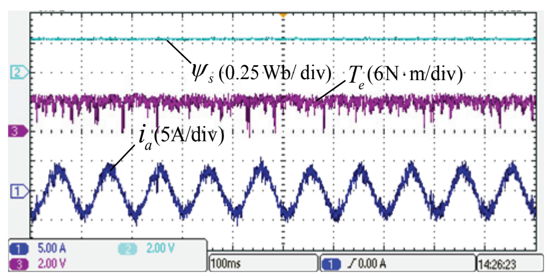
(a)



(b)



(c)



(d)

Fig. 4. Experimental results of the conventional MPTC with different weighting factors at 200r/min and the rated load: (a) w=20.25, (b) w =50, (c) w =150, (d) w =350.

TABLE III
QUANTITATIVE COMPARISON UNDER DIFFERENT WEIGHTING FACTORS

Weighting factor	ψ_s ripple(Wb)	T_e ripple(N.m)	THD(%)
20.25	0.11	0.85	53.79
50	0.023	1.23	23.43
150	0.02	1.2	19.91
350	0.021	6.1	25.44

In order to further compare the performances of both methods, the control responses of the proposed MPTC method when the motor operates at a low speed of 200r/min are shown in Fig. 5(a). Fig. 5(b)-(c) display THD comparisons of the

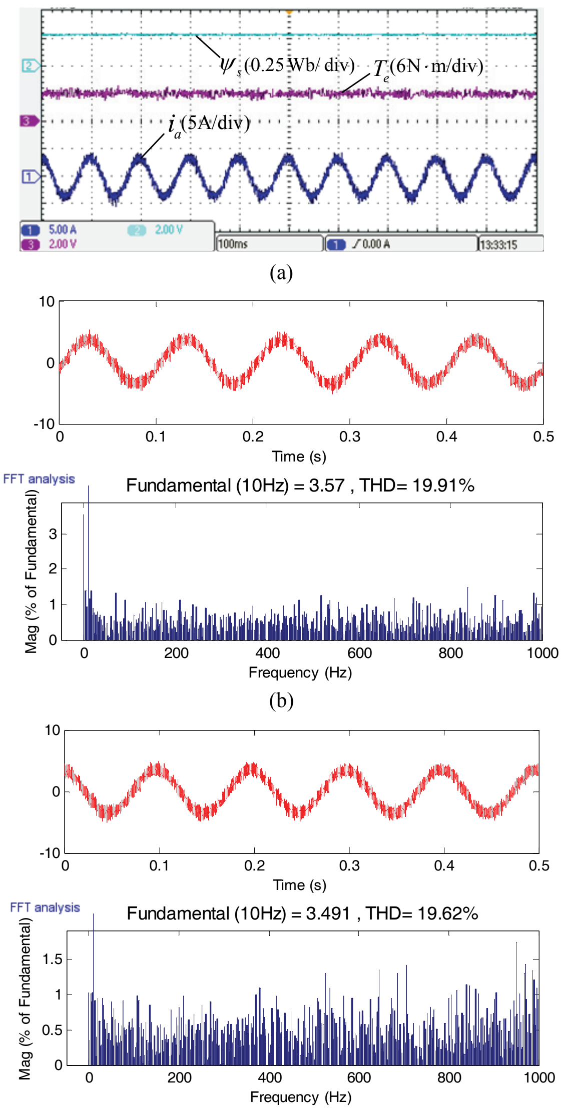


Fig. 5. Experimental results of the proposed MPTC, and a THD comparison between the conventional MPTC with w =150 and the proposed MPTC: (a) Steady response of the proposed MPTC at 200r/min and the rated load, (b) THD of the conventional MPTC with w =150, (c) THD of the proposed MPTC.

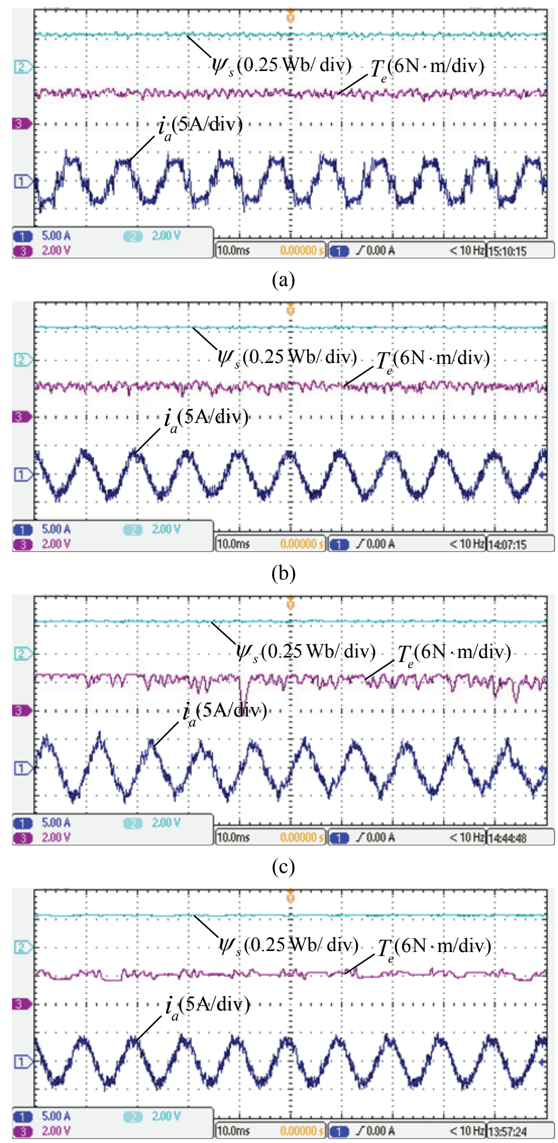


Fig. 6. Control performances of both methods with the rated speed and rated load: (a) Conventional MPTC with w=20.25, (b) Conventional MPTC with w=150, (c) Conventional MPTC with w =350, (d) Proposed MPTC.

conventional MPTC and the proposed MPTC. A comparison of Fig. 4 and Fig. 5 show that the control performances of the proposed MPTC method are similar to those of the conventional MPTC with the optimal weighting factor. However, unlike the conventional MPTC, the proposed MPTC method is able to balance the control performance of the stator flux and the electromagnetic torque without the weighting factor, which avoids the tuning that must be completed in the conventional MPTC. Similar conclusions can be obtained at a high speed of 2000 r/min, as shown in Fig. 6.

Comparison experiments of the dynamic responses also are completed, as shown in Fig. 7. It should be noted that the experimental results of the conventional MPTC are obtained under the condition of the optimal weighting factor. Fig. 7(a)-(b) show the performance when the motor accelerates from 500r/min to the rated speed 2000r/min. These results prove that rapid and non-overshoot speed tracking performance can be achieved under the control of both methods, when the speed reference suddenly varies. Experimental results of both methods at the rated speed are demonstrated in Fig. 7(c)-(d), when the load torque is suddenly changed from zero to a rated value. These results illustrate that both methods are able to achieve excellent anti-disturbance capability and quick dynamic performance.

In addition, in order to further verify the validity of the proposed MPTC, the line voltage and voltage vector trajectory are shown in Fig. 8(a)-(b), when motor operate at 500r/min with the rated load. The average switching frequencies of

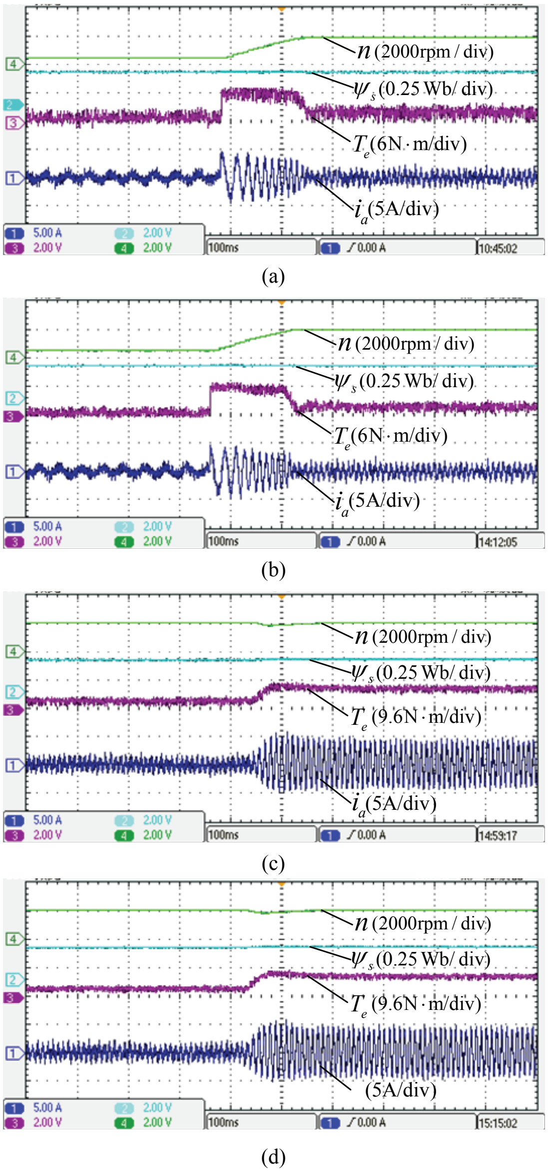


Fig. 7. Dynamic responses of both methods: (a) Response to speed reference variation with the rated load under the control of the conventional MPTC, (b) Response to speed reference variation with the rated load under the control of the proposed MPTC, (c) Response to a rated load disturbance at the rated speed under the control of the conventional MPTC, (d) Response to a rated load disturbance at the rated speed under the control of the proposed MPTC.

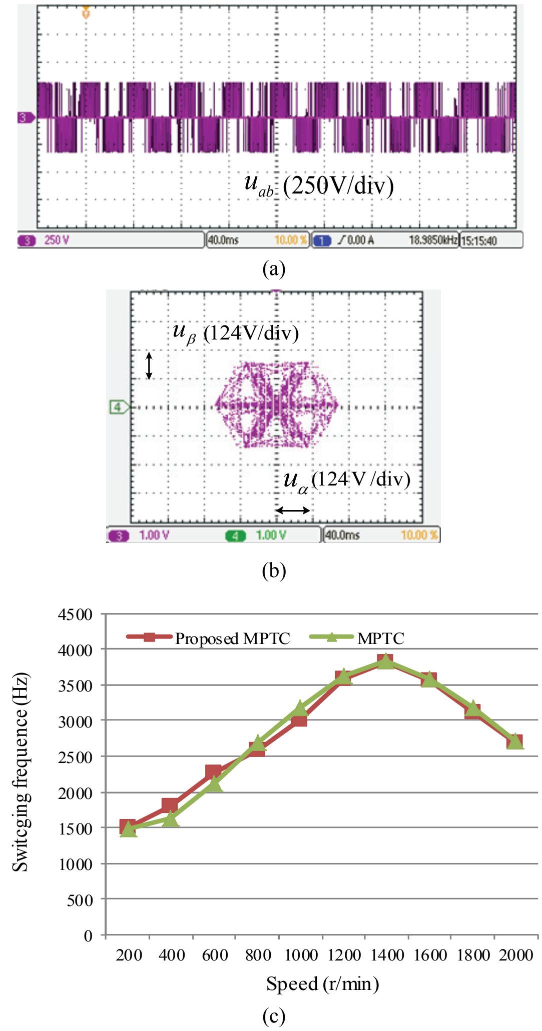


Fig. 8. Voltage waveform of the proposed MPTC and the average switching frequencies of both methods: (a) Line voltage waveform of the proposed method, (b) Voltage vector trajectory of the proposed method, (c) Average switching frequencies for the conventional MPTC and the proposed MPTC.

both methods at different speeds are shown in Fig. 8(c). This switching frequency is computed by counting the total switching jumps of the six switches of an inverter during a fixed period of 0.05s. According to Fig. 8(c), it can be seen that the switching frequencies of both methods are similar.

Based on the above experimental results, it can be clearly seen that the proposed MPTC inherits the advantage of the conventional MPTC, i.e., a quick torque response, while avoiding the disadvantage of the existence of the weighting factor. Therefore, the proposed method is more attractive and practical than the conventional MPTC.

Finally, a computation burden comparison between the proposed MPTC and the conventional MPTC is shown in Table IV, when the prediction step is one. The time in table refer to the time of code implementation. The conventional MPTC algorithm requires 65.11us to complete code operation. By comparison, 57.56us is required in the proposed method. This shows that the proposed MPTC method is able to reduce the calculation time by up to 11.6%. On the other hand, if the prediction step increases, such as two step prediction, the computation burden using the proposed method are significantly reduced, since the number of the candidate voltage vectors can be reduced from 49 to 4.

In summary, based on simulation and experimental results, it can be clearly seen that when compared with the conventional MPTC, the proposed MPTC can reduce the calculation time and eliminate the weighting factor.

TABLE IV
CALCULATION TIMES OF BOTH METHODS

Method	Conventional MPTC	Proposed MPTC
Times(us)	65.11	57.56
Call times of vectors	7	2
Time reduction(us)	7.55	

VI. CONCLUSIONS

An improved MPTC method is proposed in this paper, which has two main advantages when compared with the conventional MPTC method. Firstly, the candidate voltage vectors in the proposed method are reduced from eight to two based on the voltage sector of the desired voltage vector. Therefore, the calculation time can be significantly reduced when compared with the conventional enumeration method. Secondly, the weighting factor is eliminated in this paper by designing a cost function that only includes the voltage errors between the desired voltage and the candidate voltage. Thus, the control complexity caused by the tuning of the weighting factor is effectively decreased.

ACKNOWLEDGMENT

This work was sponsored in part by the research funds for the State Key Laboratory of Automotive Safety and Energy under Project No. KF1824, by the Beijing Natural Science Foundation under Grants #3172011, by National Natural Science Foundation of China under Grants #51507004, by the Young Top-Notch Talents Program of Beijing Excellent Talents Funding (2017000026833ZK12), and the Outstanding Young Scholars Fund of North China University of Technology.

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