Simplified Predictive Torque Control for Surface-Mounted PMSM Based on Equivalent Transformation and Partition Method

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ABSTRACT Predictive torque control (PTC) is one of the most widely discussed strategies for high performance permanent magnet synchronous motor (PMSM) drive system because of its fast dynamic response and good stability. However, the complex prediction and optimization process requires a large amount of computation. Until now, few studies have carried out any systematic research on the simplification of the PTC. In this paper, a stepwise simplification method for the PTC strategy is proposed. Firstly, by using the equivalent transformation, the prediction process is simplified without affecting the control performance of the PTC. On this basis, the cost function is further simplified with additional presuppositions, and the optimization process is replaced by honeycomb-structured space partitioning for optimal voltage vector searching. Therefore, the computation burden caused by the multiple calculations in traditional PTC is greatly reduced. In addition, this paper provides the implementation details of the proposed control strategy. Moreover, the comparison between the traditional PTC and two simplified PTC is carried out in the evaluation of the transient performance, steady-state errors, control characteristics, computation burden, and compatibility with additional constraints. The experimental results prove the feasibility and effectiveness of the simplified algorithms, and this is beneficial to the future application of PTC strategy in the low-cost motor drive system. Moreover, the simplified PTC algorithm can achieve higher switching frequency, thereby improving the performance of torque and stator flux in the motor drive system with the application of new generation SiC power device.

INDEX TERMS PMSM, predictive torque control, computation burden, cost function.

I. INTRODUCTION

In the past few years, Finite control set model predictive control (FCS-MPC) has developed rapidly and attracted growing interest of the researchers across various fields such as grid-connected power converter, motor driver, and renewable energy. Compared with the traditional linear control strategy, FCS-MPC has many advantages like inherent decoupling, fast dynamic response, and easy inclusion of non-linear constraints [1]–[4]. To achieve upon advantages, the prediction and cost function in FCS-MPC strategy should be carefully designed.

In recent years, predictive torque control (PTC), which is kind of FCS-MPC strategy, has been widely applied in the researches of motor drive systems [5]–[7]. The traditional FCS-MPC algorithm treats the current as the controlled object, while the PTC takes the torque and flux linkage as the control target. Therefore, the PTC algorithm can achieve direct control of torque and flux linkage while preserving the advantages of FCS-MPC, providing higher torque transient response speed and lower torque ripple.

However, FCS-MPC requires more computation than conventional linear control algorithms, and there are several reasons for this: First, the prediction process of FCS-MPC depends on accurate description of the motor’s mathematical model. Secondly, FCS-MPC needs to predict the effects of all voltage vectors that the inverter can output, while the
required multiple calculation will greatly increase computation burden. Finally, the cost function calculation and the optimization process require a large amount of computation. Compared with FCS-MPC, PTC is more complex in prediction process, and it requires additional computation for flux and torque estimation.

Unlike the traditional linear control algorithm, the switching frequency of PTC is inconstant without using the pulse width modulation (PWM) technology, and the control period should be set with smaller value to achieve similar steady-state performance. Roughly, in the two-level inverter based motor drive system, the control period in PTC is about one-sixth of what it would if used in conventional linear control algorithm [8]. This causes the contradiction between the program execution time and acceptable control period. Furthermore, for the multi-level inverters, the complexity of the circuit topology and switching states are greatly increased, resulting in more time consumption in the multiple calculation of the prediction and optimization process. If the computation time required by PTC exceeds the control period, the performance of the control system will be affected. With the development of the microcontroller, the operating frequency and computing performance have been greatly improved, and the computation time required by multiple calculations can be reduced by introducing the multi-core microcontrollers or FPGAs. However, the adoption of high-performance microcontrollers will significantly increase the overall cost and limit the implementation of PTC in low-cost motor drive systems.

In recent years, several simplified methods for FCS-MPC have been proposed. For example, Literature [9] proposes a method to reduce the algorithm computation time by reducing the candidate vectors in the prediction procedure with adoption of the sector distribution on the space voltage vectors. However, since the proposed algorithm is not completely equivalent to the original one, the performance of the control strategy may be affected. In [10], [11], the simplified FCS-MPC for two-level converters, three-level converters, and matrix converters are proposed and discussed. The prediction and optimization process are simplified by using equivalent transformation. Moreover, the simplified algorithm can maintain the control performance of the traditional FCS-MPC with the additional constraints. In [12], the lyapunov principle is introduced into the design of partition method, thereby eliminating the unwanted voltage vector and simplifying the control strategy of the FCS-MPC for grid-connected converter.

Based on the approaches above, some simplified algorithms for PTC have been proposed. In [13], the number of candidate voltage vectors is reduced by introducing a direct torque control (DTC) switching table in the induction motor drive system; In [14], the complex weighting factor selection process in the cost function is simplified by using the TOPSIS algorithm. In [15], a cost function with the consideration of the voltage vector error is constructed based on duty cycle calculation for PTC, and the candidate voltage vectors and duty cycles are determined according to the geometric position of the reference voltage vector in the inverse triangular matrix. On this basis, the processes of the prediction and duty cycle calculation are combined into one, which effectively reduces the complexity of PTC algorithm. In [16], the control structure of the multi-step direct torque prediction algorithm is simplified by using the look-up table, avoiding the square root operation and trigonometric function operation in traditional PTC strategy. For the six-phase PMSM control system, several simplified PTC strategies are proposed. In [17], a voltage vector optimization algorithm is proposed to reduce the candidate vectors according to the stator flux vector and torque deviation. In other studies, some researchers reduce the candidate vectors by using the simplified prediction model and look-up table [18]. In [19], a deadbeat direct torque and flux control (DB-DTFC) method is proposed to obtain the optimal voltage vector, replacing the coordinated control of torque and flux by single-objective control, which simplifies the structure of the cost function and reduces parameter adjustment difficulty of PTC. The similar method is also applied in the control of the dual inverter-fed open-winding PMSM and induction machine (IM), converting the traditional cost function into a simple cost function and simplifying the structure of PTC [20], [21]. In the studies of the permanent magnet synchronous generator (PMSG) based wind power system, a cost function based on the torque and the reactive torque is proposed to simplify the structure of PTC algorithm [16]. In other circuit topologies such as the four-switch three-phase inverter-fed PMSM drive system, or the matrix converter-fed PMSM drive system, the simplified method for PTC algorithm is realized by converting the capacitor voltage offset (CVO) control and torque control into flux linkage control, or by reducing the candidate voltage vectors needed in the prediction process [22], [23].

All of the studies reviewed can reduces the complexity of the PTC, however, there are still several problems:

1. Most simplified PTC strategies are based on the duty cycle control strategy with constant switching frequency. However, The studies for the simplification of traditional PTC with variable switching frequency are still rare;

2. The simplification of PTC strategy requires several preconditions such as assuming the stator flux linkage remains constant during the control period, or assuming the d-axis stator current remains zero, etc. However, these preconditions do not exist in the traditional PTC strategy, and the related derivations and discussions can not prove whether the simplified strategy is equivalent to the traditional strategy. The corresponding assumptions may not be well-considered and experimentally validated. Therefore, it is difficult to prove whether the simplified strategy can maintain all the features of the traditional strategy.

3. Most researches focus on simplification in certain parts of PTC such as the prediction process or candidate vector selection. Other studies attempt to simplify the PTC strategy for specific topologies or electrical machines. However, the systematic optimization strategy for PTC is still lack
of theoretical basis and experimental study. Furthermore, the detailed implementation and comprehensive performance evaluation for the simplified PTC are rarely mentioned in relevant studies.

In this paper, the simplification of PTC with variable switching frequency is discussed with comprehensive and systematic studies. Firstly, the mathematical model of the surface-mount PMSM is established by using the vector-form state equations in α-β stationary reference frame. With the equivalent transformation, the multiple calculation in the prediction process is converted into a simple calculation for the variables $B_1$ and $B_2$, which realizes the simplification of the algorithm structure and the reduction of the computation burden. The simplification process does not contain preconditions and therefore will not affect the performance and features possessed by the traditional PTC. On this basis, additional preconditions are present to realize further simplification of the integrated algorithm. A specially designed honeycomb-like partition structure and the corresponding voltage vector optimization method are present, so the cost reduction burden. The simplification process does not contain additional preconditions and therefore will not affect the performance and features of the proposed control strategy.

Finally, with the discussion concerning about the impact of the simplification process on the PTC strategy, the simplified algorithms are evaluated in terms of steady-state performance, transient performance, inherent decoupling characteristics, control object adjustment characteristic, and additional constraints compatibility on surface-mount PMSM experimental platform. Moreover, the computation burden of the proposed algorithms is also measured and evaluated.

II. TRADITIONAL PTC FOR SURFACE-MOUNTED PMSM

In d-q synchronization reference frame, the state equations of PMSM can be expressed as

\[
\begin{align*}
\frac{d\psi_d}{dt} &= u_d - R_i d + \omega_c \psi_q \\
\frac{d\psi_q}{dt} &= u_q - R_i q - \omega_c \psi_d
\end{align*}
\]

where $\psi_d$, $\psi_q$ are the stator fluxes; $u_d$, $u_q$ are the stator voltages; $i_d$, $i_q$ are the stator currents; $R$ is the stator winding resistance. $\omega_c$ is the electrical angular velocity of the rotor.

The equations of $\psi_d$, $\psi_q$ are expressed as follows

\[
\begin{align*}
\psi_d &= L_d i_d + \psi_f \\
\psi_q &= L_q i_q
\end{align*}
\]

where $L_d$, $L_q$ are the stator winding inductance in d-q synchronization reference frame, $\psi_f$ is the permanent magnet flux of the rotor.

Substituting (2) into (1), one can obtain the state equations about $i_d$ and $i_q$,

\[
\begin{align*}
\frac{di_d}{dt} &= -\frac{R}{L} i_d + \frac{L_q}{L} \psi_f + \frac{u_d}{L} \\
\frac{di_q}{dt} &= -\frac{R}{L} i_q - \frac{L_d}{L} \psi_f + \frac{u_q}{L} - \frac{\psi_d}{L} \omega_e
\end{align*}
\]

For the surface-mounted PMSM, it has $L_d = L_q = L$, so (3) can be rewritten as

\[
\begin{align*}
\frac{di_d}{dt} &= -\frac{R}{L} i_d + \omega_c i_d + \frac{u_d}{L} \\
\frac{di_q}{dt} &= -\frac{R}{L} i_q - \omega_c i_d + \frac{u_q}{L} - \frac{\psi_d}{L} \omega_e
\end{align*}
\]

According to the Euler’s discretization, the derivative of the stator currents $i_d$, $i_q$ can be expressed as

\[
\begin{align*}
\frac{di_d}{dt} &= i_d(k+1) - i_d(k) \\
\frac{di_q}{dt} &= i_q(k+1) - i_q(k)
\end{align*}
\]

where $T$ is the control period of the motor drive.

Applying (5) to (4), one can obtain the prediction of the stator currents shown as follows.

\[
\begin{align*}
i_d(k+1) &= (1 - \frac{R}{L}) i_d(k) + T \omega_c i_d(k) + \frac{T}{L} u_d(k) \\
i_q(k+1) &= (1 - \frac{R}{L}) i_q(k) - T \omega_c i_d(k) + \frac{T}{L} u_q(k) - \frac{\psi_f}{L} T \omega_e
\end{align*}
\]

The stator flux in d-q synchronization reference frame can be written as

\[
\begin{align*}
\psi_d(k+1) &= L_i d(k+1) + \psi_f \\
\psi_q(k+1) &= L_i q(k+1)
\end{align*}
\]

The modulus of the stator flux is expressed as

\[
|\psi_s(k+1)| = \sqrt{|\psi_d^2(k+1) + \psi_q^2(k+1)}
\]

The electromagnetic torque $T_e$ can be calculated by

\[
T_e(k+1) = 1.5 p \left[\psi_d(k+1) i_d(k+1) - \psi_q(k+1) \psi_f(k+1)\right]
\]

where $p$ is the pole pairs;

The cost function $g$ can be defined as

\[
g = \lambda_T (T_e^* - T_e(k+1))^2 + \lambda_\psi (|\psi_s^*| - |\psi_s(k+1)|)^2
\]

where $\lambda_T$, $\lambda_\psi$ are the weighting factors of the cost function; $T_e^*$, $|\psi_s^*|$ are the reference values of the torque and stator flux modulus.

The 2-level inverter in motor drive system contains eight types of switching states. The corresponding voltage vectors are the non-zero vectors $V^1 - V^6$, and the zero vectors $V^0$, $V^7$, shown in Fig. 1.
During each control period, the voltage vectors $V^0$, $V^7$ are used for the flux and torque prediction with (6-9). The realization of the PTC strategy is to find the optimal voltage vector $V^0$ which minimizes the cost function defined in (10) then apply the corresponding switching state to the 2-level inverter in motor drive system, so it has

$$V^0 = \arg \min g_{V^0} \quad n = 0 - 7$$  

The equations of the surface-mounted PMSM in α-β stationary reference frame are expressed as follows

$$\begin{align*}
V &= RI + \frac{d\psi_s}{dt} \\
\psi_s &= LI + \psi_f
\end{align*}$$  

where $R$ is the stator winding resistance; $L$ is the stator inductance; $V$, $I$, $\psi_s$, $\psi_f$ are the stator voltage, stator current, stator flux and rotor flux in complex field, respectively, and it has: $V = u_\alpha + ju_\beta$; $I = i_\alpha + ji_\beta$; $\psi_s = \psi_{sa} + j\psi_{sb}$; $\psi_f = \psi_{fa} + j\psi_{fb}$; $|\psi_f| = \psi_f$

The electromagnetic torque $T_e$ can be expressed as

$$T_e = \frac{3p}{2} \text{Im}(\overline{\psi_s}I)$$  

where $\overline{\psi_s}$ is the complex conjugate of $\psi_s$; $p$ is the pole pairs.

The variables $\psi_s$, $\psi_f$ can be expressed as $|\psi_s|e^{j\theta_s}$, $|\psi_f|e^{j\theta_f}$, where $\theta_s$ and $\theta_f$ are the phase angles of $\psi_s$ and $\psi_f$, respectively. According to (12), the derivative of the stator flux modulus $|\psi_s|$ can be expressed as

$$\frac{d}{dt}|\psi_s| = \frac{1}{|\psi_s|} \text{Re} \left[ \overline{\psi_s}(V - RI) \right]$$  

Similarly, the derivative of the electromagnetic torque $T_e$ can be obtained by (12) and (13), and it has

$$\frac{dT_e}{dt} = -\frac{3p}{2L} \left[ R \text{Im}(\overline{\psi_s}\psi_f) + \omega_e \text{Re}(\overline{\psi_s}\psi_f) - \text{Im}(V\psi_f) \right]$$

By neglecting the impact of stator resistance $R$, (14) can be rewritten as

$$\tau_e = \frac{d}{dt}|\psi_s| = |V| \cos(\theta_v - \theta_s)$$  

where $\theta_v$ denotes the phase angle of the stator voltage $V$.

The stator voltage $V$ can be expressed as $|V|e^{j\theta_v}$. According to (15), the derivative of the electromagnetic torque $T_e$ can be replaced by

$$\frac{dT_e}{dt} = \tau_T + \tau_0$$  

where

$$\begin{align*}
\tau_T &= 3p |\psi_f| |V| \sin(\theta_v - \theta_f) \\
\tau_0 &= -\frac{RT_e}{L} - \frac{3pe_e |\psi_f| |\psi_s| \cos(\theta_v - \theta_s)}{2L}
\end{align*}$$

To simplify the equations in (18), defining that

$$k_e = \frac{3p |\psi_f|}{2L}$$

The derivative of the electromagnetic torque $T_e$ and stator flux modulus $|\psi_s|$ can be rewritten as

$$\begin{align*}
\frac{d}{dt}T_e &= \tau_0 + k_e |V| \sin(\theta_v - \theta_e) \\
\frac{d}{dt}|\psi_s| &= |V| \cos(\theta_v - \theta_s)
\end{align*}$$

To realize the simplification of PTC, the first step is to reduce complexity of prediction procedure. With the consideration of (21), by assuming an voltage vector $V^*$ which
can make \(| \Psi_s(k+1) | \) and \( T_e(k+1) \) reach to the reference values \(| \Psi_s(k+1) | \) and \( T_e(k+1) \) at the \( t(k+1) \) time instant, it has

\[
\begin{aligned}
T_e^*(k+1) &= T_e(k) + k_e \left[ V^* \right] \sin(\theta^* - \theta_e) + \tau_0 T \\
\left| \Psi_s^*(k+1) \right| &= \left| \Psi_s(k) \right| + \left| V^* \right| \cos(\theta^* - \theta_s) T
\end{aligned}
\]

(22)

where \( \theta^* \) is the angle of the voltage vector \( V^* \).

\( V^* \) is named as the prediction vector, and (22) can be further transformed to

\[
\begin{aligned}
T_e^*(k+1) - T_e(k) - \tau_0 T &= -u_d^* \sin \theta_e + u_d^* \cos \theta_e \\
\left| \Psi_s^*(k+1) \right| - \left| \Psi_s(k) \right| &= u_d^* \cos \theta_s + u_d^* \sin \theta_s
\end{aligned}
\]

(23)

where \( u_d^* = \left| V^* \right| \cos \theta^* ; u_d^* = \left| V^* \right| \sin \theta^* \).

The equations in (23) can be rewritten as

\[
AX = B
\]

(24)

where

\[
A = \begin{bmatrix}
-\sin \theta_e & \cos \theta_e \\
\cos \theta_s & \sin \theta_s
\end{bmatrix} ; \quad X = \begin{bmatrix}
u_d^* \\
u_d^*
\end{bmatrix} ; \quad B = \begin{bmatrix} B_1 & B_2 \end{bmatrix}
\]

The expression of \( u_d^* , u_d^* \) can be obtained by (24), that is

\[
\begin{aligned}
u_d^* &= \frac{B_2 \cos \theta_e - B_1 \sin \theta_s}{\cos(\theta_e - \theta_s)} \\
u_d^* &= \frac{B_1 \cos \theta_s + B_2 \sin \theta_e}{\cos(\theta_e - \theta_s)}
\end{aligned}
\]

(25)

On the other hand, the effects of the eight voltage vectors on the derivative of \( | \Psi_s | \) and \( T_e \) should be discussed. By substituting \( V^* \) into (20), there is

\[
\begin{aligned}
\frac{dT_e}{dt} &= \tau_0 + k_e \left| V^* \right| \sin(\theta_e - \theta_e) \\
\frac{d|\Psi_s|}{dt} &= \left| V^* \right| \cos(\theta_e - \theta_e)
\end{aligned}
\]

(26)

By using Euler’s discretization, (26) can be transformed into

\[
\begin{aligned}
T_e^{n+1} - T_e^n - \tau_0 T &= -u_d^n \sin \theta_e + u_d^n \cos \theta_e \\
\left| \Psi_s^{n+1} \right| - \left| \Psi_s^n \right| &= u_d^n \cos \theta_s + u_d^n \sin \theta_s
\end{aligned}
\]

(27)

By subtracting (27) from (23), one can get

\[
\begin{aligned}
T_e^{n+1} - T_e^n &= \frac{k_e T}{u_d^n} \sin \theta_e + (u_d^n - u_d^n) \cos \theta_e \\
\left| \Psi_s^{n+1} \right| - \left| \Psi_s^n \right| &= \frac{k_e T}{u_d^n} \cos \theta_s + (u_d^n - u_d^n) \sin \theta_s
\end{aligned}
\]

(28)

Based on (28) and (10), one can obtain

\[
\begin{aligned}
g &= \lambda \tau k_e^2 T^2 \left( (u_d^n - u_d^n) \sin \theta_e + (u_d^n - u_d^n) \cos \theta_e \right)^2 \\
+ \lambda \phi T^2 \left( (u_d^n - u_d^n) \cos \theta_s + (u_d^n - u_d^n) \sin \theta_s \right)^2
\end{aligned}
\]

(29)

Considering that the control period \( T \) does not affect the optimization results of the cost function, assuming that

\[
\lambda = \frac{\lambda \tau k_e^2 T^2}{\lambda \phi} = \frac{1.5 \phi}{L T^2}
\]

(30)

Equation (29) can be expressed as

\[
\begin{aligned}
g &= \lambda \left( (u_d^n - u_d^n) \sin \theta_e + (u_d^n - u_d^n) \cos \theta_e \right)^2 \\
+ \left( (u_d^n - u_d^n) \cos \theta_s + (u_d^n - u_d^n) \sin \theta_s \right)^2
\end{aligned}
\]

(31)

By substituting (25) into (31), one can obtain the cost function of the simplified PTC algorithm

\[
\begin{aligned}
g &= \lambda (B_1 + u_d^n \sin \theta_e - u_d^n \cos \theta_e)^2 \\
+ (B_2 - u_d^n \cos \theta_s - u_d^n \sin \theta_s)^2
\end{aligned}
\]

(32)

Equation (32) is the cost function obtained by equivalent transformation, and the corresponding simplified strategy is named as equivalent transformation based PTC (ET-PTC), with its structure diagram shown in Fig. 3.

![FIGURE 3. Simplified PTC strategy based on equivalent transformation.](image-url)
speed of the stator flux linkage in PLL can be adjusted. The calculation of the transcendental function calculation required by the simplified PTC is limited to the sine function and the cosine function, which can be easily implemented by using the look-up table. Moreover, the PLL can avoid the noise caused by the division calculation in the traditional algorithm.

In the practical implementation of the simplified PTC, the control algorithm must take one control period before outputting the optimal switching state, so it has a fixed time delay. In order to avoid the control performance degradation caused by the time delay, a compensation algorithm is established. According to (12), it has

\[
\begin{align*}
\frac{L}{d} \frac{di_{\alpha}}{dt} &= u_{\alpha} - R_i u_{\beta} + |\Psi_r| \sin \theta_e \\
\frac{L}{d} \frac{di_{\beta}}{dt} &= u_{\beta} - R_i u_{\alpha} - |\Psi_r| \cos \theta_e
\end{align*}
\]

(33)

By ignoring the stator resistance and applying the discretization method, it has

\[
\begin{align*}
i_{\alpha}(k+1) &= \frac{T}{L} (u_{\alpha} + |\Psi_r| \sin \theta_e) + i_{\alpha}(k) \\
i_{\beta}(k+1) &= \frac{T}{L} (u_{\beta} - |\Psi_r| \cos \theta_e) + i_{\beta}(k)
\end{align*}
\]

(34)

So the compensated currents can be obtained by (34), that is

\[
\begin{align*}
i_{\alpha} &= \frac{T}{L} (u_{\alpha} + |\Psi_r| \sin \theta_e) + i_{\alpha}(k) \\
i_{\beta} &= \frac{T}{L} (u_{\beta} - |\Psi_r| \cos \theta_e) + i_{\beta}(k)
\end{align*}
\]

(35)

where \(i_{\alpha}, i_{\beta}\) are the currents at \(k\) time instant; \(u_{\alpha}, u_{\beta}\) are the output voltages in \(\alpha-\beta\) stationary reference frame, corresponding to the optimal voltage vector \(V^o\) obtained by the PTC algorithm during the control period of \(k-1\) to \(k\) time instant.

On the basis of ET-PTC, considering the special case that \(\lambda = 1.0\), (32) can be changed to

\[
g = \left( (u_{\alpha}^n - u_{\alpha}^n) \sin \theta_e + (u_{\beta}^n - u_{\beta}^n) \cos \theta_e \right)^2 \\
+ \left( (u_{\alpha}^n - u_{\alpha}^n) \cos \theta_e + (u_{\beta}^n - u_{\beta}^n) \sin \theta_e \right)^2
\]

(36)

During the operation of the PMSM, the air-gap magnetic field is mainly generated by the rotor permanent magnet, while the magnetic field generated by the stator current is relatively small. Therefore, the angles of the stator flux and the rotor flux are very close to each other. It can be assumed that \(\theta_e \approx \theta_s\), then the above equation becomes

\[
g = (u_{\alpha}^n - u_{\alpha}^n)^2 + (u_{\beta}^n - u_{\beta}^n)^2
\]

(37)

That is

\[
g = |V^* - V^n|^2
\]

(38)

The equation above shows that, the cost function in PTC strategy can be approximately transformed into the voltage vector form, and the voltage vectors \((V^0 - V^7)\) which is nearest the prediction vector \(V^*\) will produce the minimum cost function result, so the optimal voltage vector can be selected by the following equation.

\[
V^o = \arg \min_g V^n \quad n = 0 - 7
\]

(39)

Moreover, the equation about the prediction vector calculation in (25) can be simplified as

\[
\begin{align*}
u_{\alpha} &= B_2 \cos \theta_e - B_1 \sin \theta_e \\
u_{\beta} &= B_1 \cos \theta_e + B_2 \sin \theta_e
\end{align*}
\]

(40)

The calculation of \(t_0\) in (18) can be further simplified as

\[
t_0 = -\frac{RT_e}{L} - \frac{1.5 p \omega_e |\Psi_s| \psi_r|}{L}
\]

(41)

The process of optimal voltage vector selection still needs 8 times of calculations for the cost function results by (37), and the complex comparison process between the results of \(g\) in (39).

Considering that the selection of the optimal voltage vector is to find the \(V^o\) which is nearest to \(V^n\), one can further simplify the PTC algorithm by partition the voltage vector space into 7 sections, which are sections I-VII, shown in Fig. 5.

In the honeycomb-structured section distribution, the voltage vector space is divided into 7 sections with different colors. The border (black dotted line) of the section VII consists of the perpendicular bisectors of the voltage vectors.
The borders (black dotted line) between the sections I~VI consist of the perpendicular bisectors of the hexagon’s six sides (red dotted line) in the voltage vector space. According to position where the end of the voltage vector \( V^m \) is located, the zero voltage vectors \( V^0 \) and \( V^7 \) belong to section VII, while the non-zero voltage vectors \( V^1 \sim V^6 \) belong to sections I~VI, respectively. It can be found that, if the end of voltage vector \( V^* \) falls in one section, the voltage vector \( V^n \) contained by this section will be nearest to \( V^* \). Therefore, according to the section which the prediction vector \( V^* \) falls in, the output voltage vector can be determined and then applied to the 2-level inverter of the motor drive.

A section identification method is designed with 2 steps summarized below:

**First**, the voltage vector space is divided into 6 sectors (shown in Fig. 6(a)) by using the perpendicular bisectors of the hexagon’s six sides. With the inverse Clarke transform on \( V^* \), it has

\[
\begin{align*}
\alpha = u^* - u^* \alpha + \frac{\sqrt{3}}{2} u^* \beta \\
\beta = \frac{1}{2} u^* - \frac{\sqrt{3}}{2} u^* \beta \\
\gamma = \frac{1}{2} u^* + \frac{\sqrt{3}}{2} u^* \beta
\end{align*}
\]

where

\[
\text{sgn}(x) = \begin{cases} 
1, & x \geq 0 \\
0, & x < 0
\end{cases}
\]

The range of the candidate voltage vectors are narrowed down to the zero vectors and one non-zero vector.

**Second**, the voltage vector space can be further divided into 7 sections by adding the perpendicular bisectors of the voltage vectors \( V_1 \sim V_6 \), shown in Fig. 6(b), and it can be found that the section where \( V^* \) falls can be identified by calculating the projection of \( V^* \) on the non-zero vector which belongs to the sector \( N \), so it has

\[
N^f = \begin{cases} 
N, & 0 \leq u^* \alpha + u^* \beta \leq \frac{2}{9} u^2_{dc} \\
0, & 0.7 \leq u^* \alpha + u^* \beta \leq \frac{2}{9} u^2_{dc}
\end{cases}
\]

With the identified section number \( N^f \), the optimal voltage vector is obtained, and the corresponding simplified strategy is named as space partition based PTC (SP-PTC), with its structure diagram shown in Fig. 7.

**IV. VERIFICATION VIA EXPERIMENT**

The experimental platform is shown in Fig. 8. The test motor is a surface-mounted PMSM with encoder, and the resolution is 2000 lines per rev. This motor is mainly used in the fields of CNC machines, door motor of elevator, robots, and industrial automation equipment.

The core board is based on Texas Instruments (TI) floating point digital signal processor (DSP) TMS320F28335 with computing power of 150 million instructions per second (MIPS), and the power device is based on Cree SiC MOSFET C2M0080120D.

The parameters of the surface-mounted PMSM control system is shown in Table 1. The control period is 50\( \mu \)s.

**TABLE 1. Parameters of the PMSM control system.**

| Symbol | Parameter | Value |
|--------|-----------|-------|
| \( n_r \) | rated speed | 1500r/min |
| \( T_N \) | rated torque | 1.27N.m |
| \( p \) | pole pairs | 4 |
| \( R \) | stator resistance | 0.56Ω |
| \( L_s \) | stator inductance | 1.55 mH |
| \( \Psi_0 \) | rotor flux | 0.042Wb |
| \( U_{dc} \) | DC Voltage | 50.0 V |

The steady-state and transient performance of the traditional PTC (T-PTC) and two simplified algorithms are analyzed, shown in Fig. 9.
In Fig. 9, the torque reference $T_e^*$ increases from 0.5$T_N$ to 1.0$T_N$ at 0.1s, while the flux linkage reference $|\psi^*_s|$ decreases from 1.0$|\psi_s|$ to 0.9$|\psi_s|$ at 0.2s. The detailed diagram of the step response waveforms shows that the three algorithms almost have the same transient response speed, and the details of the step response waveforms are similar. Furthermore, if the reference value of one controlled variable $T_e$ or $|\psi_s|$ changes, the transient performance of the other controlled variable will not be affected, which indicates that the performances of the simplified algorithms are consistent with the traditional PTC algorithm in the inherent decoupling characteristic. In addition, it can be determined that the PLL introduced in the simplified algorithms does not affect the transient performance of the control system.

The experimental results shown above also include the evaluation about the steady-state performance of the three control strategies. The steady-state errors of the torque and flux linkage are respectively defined as

$$
\begin{align*}
T_e^{\text{error}} &= \frac{1}{T_x} \int_0^{T_x} \left| T_e^* - T_e(t) \right| \, dt \\
\psi_s^{\text{error}} &= \frac{1}{T_x} \int_0^{T_x} \left| |\psi^*_s| - |\psi_s(t)| \right| \, dt
\end{align*}
$$

where $T_x = kT$; $k = 1000$.

It can be seen from the figures that, the steady-state errors of the torque and flux linkage in ET-PTC are almost the same as those in T-PTC. This indicates that the ET-PTC maintains the steady-state performance of traditional PTC.

In contrast, the SP-PTC has a relatively large steady-state error of flux linkage, which could possibly be due to the approximate setting $\theta_e \approx \theta_s$ in the simplification process. However, considering that the flux linkage reference $|\psi^*_s|$ is 0.042Wb in the above figure, the steady-state error of SP-PTC is about $0.82 \times 10^{-4}$Wb higher than that of T-PTC. This value is only about 0.2% of $|\psi^*_s|$, so the influence can be neglected.

Figs. 10-12 show the comparison of the stator flux trajectories, current total harmonic distortion (THD), and voltage vector selections between the traditional PTC and two simplified PTC, respectively.

In Fig. 10, the trajectories of the stator fluxes are plotted in the $\alpha\beta$ coordinate system. It can be seen that the shape of the...
In Fig. 10, the stator flux trajectories obtained by the three algorithms are similar, particularly in region A, region B, and region C.

In Fig. 11, the torque reference $T^* = 0.8T_N$, and the flux linkage reference $|\psi^*_s| = |\psi_s| = 0.042$ Wb. The THD analysis performed on the stator currents shows that the fundamental frequency amplitudes and THD results of the three algorithms are very close to each other. Moreover, the spectrums of the three algorithms in the frequency range of 0-20 kHz are very similar.

In Fig. 12, since the control effects of $V^0$ and $V^7$ on the motor are the same in the voltage vector selection, when the optimal voltage vector is zero vector, only $V^0$ is selected instead of $V^7$. It can be seen from the figures that, although the initial conditions of experiments cannot be exactly the same, the optimal voltage vector selection trends of the three algorithms are very close to each other.

In the PTC strategy, the setting value of the weighting factor will affect the proportion of torque and stator flux components in the cost function, thus affecting the control effect of stator flux and torque. Moreover, the waveforms of stator currents will also be affected. Fig. 13 shows the waveforms of the torque, stator flux and stator currents when the weighting factor is gradually changed.

In traditional PTC, the weighting factors of the cost function are $\lambda_\psi$ and $\lambda_T$, shown in equation (10); while the weighting factor of the ET-PTC is $\lambda_e$ which is determined by $\lambda_\psi$ and $\lambda_T$ in equation (30). Therefore, by using the same setting values of the $\lambda_\psi$ and $\lambda_T$, the steady-state performance associated with the weighting factor can be compared between the T-PTC and the ET-PTC.

In Fig. 13(a) and Fig. 13(b), $\lambda_\psi/\lambda_T$ is gradually decreased from $10.0k_e^2$ to $0.1k_e^2$, where $k_e$ is the constant value related to motor’s parameters, shown in equation (19). It can be seen from the figures that, with the decreasing of the $\lambda_\psi/\lambda_T$, the steady-state error of the torque $T_e$ gradually decreases, and at the same time, the steady-state error of the stator flux modulus $|\psi_s|$ gradually increases. Too high or too low setting values of $\lambda_\psi/\lambda_T$ will cause severe fluctuations in torque or stator flux modulus with higher distortion of the
stator currents. Considering about this feature, the users are needed to adjust the weighting factor according to the control targets, thereby changing the tendency of the cost function for different control targets ($T_e$ or $|\psi_s|$). For example, if the motor control system application requires lower control error of the torque, the user should specify the weighting factors to satisfy a relative small value of $\lambda_{\psi}/\lambda_T$, and this will sacrifice the control performance of stator flux in exchange for the improvement of the torque control performance. By comparing the experimental results, it can be seen that the control performance of the ET-PTC is the same as that of the T-PTC.

In the derivation process of the SP-PTC strategy, the weighting factor $\lambda$ applied in ET-PTC is removed, so the corresponding value cannot be directly set. In order to solve this problem, by removing the partitioning method, the modified cost function is used for optimization, shown below.

$$g = \lambda_v (u_{d}^n - u_{d}^s)^2 + (u_{p}^n - u_{p}^s)^2$$  \hspace{1cm} (46)

The corresponding experimental results of SP-PTC are shown in Fig. 13(c). It can be seen from the figures that, if the weighting factor $\lambda_v$ is gradually decreased, the steady performances of the torque and flux linkage are synchronously
deteriorated, and at the same time, the stator currents are significantly distorted. This shows that the proportion of the torque and stator flux components in the modified cost function of SP-PTC cannot be adjusted with the weighting factor $\lambda_v$, and the performance of one control target cannot be improved by deteriorating the performance of the other control target in the SP-PTC strategy.

In the middle section of the figures (0.15s−0.2s, where $\lambda_\psi/\lambda_T = 1.0k_e^2$ or $\lambda_v = 1.0$), the steady-state errors of the three control strategies on the torque, flux, and stator current are basically the same. This shows that the SP-PTC is only a special case where the weighting factor in T-PTC or ET-PTC is set to $\lambda_\psi/\lambda_T = 1.0k_e^2$. With this setting, the control performance of torque and stator flux can be well balanced.

In order to evaluate the compatibility of the simplified algorithm with additional constraints, Fig. 14 shows the experimental results after adding the switching frequency constraint.

In the motor drive system, higher switching frequency can improve the stator current waveform and reduce the torque ripple, but it will also bring additional switching losses, which may cause heat dissipation problems in the motor drive system. On the basis of SP-PTC strategy, the switching frequency constraint is added by modifying the cost function in (37), so there is

$$
g = (u_{a}^{\alpha} - u_{a}^{s})^2 + (u_{\beta}^{\alpha} - u_{\beta}^{s})^2 + \lambda_{sw}C_n^\alpha
$$

where $\lambda_{sw}C_n^\alpha$ is the additional switching frequency constraint term; $C_n^\alpha$ is the sum of the switching times of the power inverter at the $kT$ time instant.

In Fig. 14(a), with the increasing of the weighting factor $\lambda_{sw}$, the harmonic component of the stator currents gradually increases. When $\lambda_{sw}$ is set to 10 and 100 respectively, the current waveform does not change significantly, and the fluctuation of the torque $T_e$ and stator flux modulus $|\psi_s|$ basically remains unchanged. When the value of $\lambda_{sw}$ rises to 1000, obvious harmonic component appears in the current waveforms. Moreover, the fluctuation of $T_e$ and $|\psi_s|$ increases significantly.

In Fig. 14(b), as the weighting factor $\lambda_{sw}$ increases, the switching times $n_{sw}$ during each control period decreases gradually. It can be seen that, during each time interval of 0.5s, the cumulative switching times $N_{sw}$ are 1298, 1165, and 645, respectively. This means that the simplified PTC algorithm can adjust the switching frequency of the motor drive by setting appropriate value of the weighing factor, and it also proves that the SP-PTC strategy is compatible with the additional constraints of switching frequency like the traditional PTC.

Fig. 15 shows the stator currents, torque, stator flux modulus, and common mode voltage of the SP-PTC strategy after adding the common mode voltage constraint.

During the operation of the motor drive system, common mode voltages with high frequency and large fluctuation range may cause insulation and electromagnetic interference problems. On the basis of the SP-PTC strategy, the common mode voltage constraint is added by modifying the cost function shown in (37), so it has

$$
g = (u_{a}^{\alpha} - u_{a}^{s})^2 + (u_{\beta}^{\alpha} - u_{\beta}^{s})^2 + \lambda_{cm}u_{cm} - \frac{u_{dc}}{2}
$$

where

$$
u_{cm} = \frac{u_{dc}}{3}(s_{a} + s_{\beta} + s_{c})
$$

where the component $\lambda_{cm}|u_{cm} - u_{dc}/2|$ is the additional common mode voltage constraint term; $u_{cm}$ is the common mode voltage of the motor drive when $V_s^\alpha$ is selected as the output voltage vector during the control period $kT \sim (k + 1)T$; $\lambda_{cm}$ is the weighting factor of the common mode voltage constraint.

In the figure, when the weighting factor $\lambda_{cm}$ is set to a relatively small value ($\lambda_{cm} = 1.0$), the stator currents contain less harmonic component, and the fluctuation ranges of the torque and stator flux modulus are small; when the value of $\lambda_{cm}$ rises to 50, the stator currents contains obvious harmonic

![FIGURE 14. Control performance of SP-PTC under different setting values of weighting factor when adding the switching frequency constraints.](image)
component, and the steady-state errors of the torque and stator flux modulus rise significantly; when the value of \( \lambda_{cm} \) rises to 1000, the distortion of the stator current waveform and the steady-state errors of the torque and stator flux modulus don’t change substantially.

The fluctuation range of the common mode voltage decreases gradually with the increasing of the weighting factor \( \lambda_{cm} \). The fluctuation range of \( u_{cm} \) is 33V when \( \lambda_{cm} \) is set to 1.0, then the fluctuation range of \( u_{cm} \) decrease by 50% when \( \lambda_{cm} = 1000 \). The control performance of torque, stator flux and common mode voltage can be balanced by adjusting the value of weighting factor. So it can be proved that the SP-PTC is compatible with the additional constraints of common mode voltage as the traditional PTC.

The computation burdens of the traditional PTC and the two simplified PTC algorithms are shown in Table 2-Table 4. Considering the differences in algorithm structure between the traditional PTC and the two simplified PTC, this paper presents the computation amount of each stage in the whole PTC strategy. In the Table 2-Table 4, the symbol “+, −” indicates the addition or subtraction operation; “×” indicates the multiplication operation; “\( f(x) \)” indicates the trigonometric or square root operation; “>, <” indicates the numerical comparison process. Through the three PTC strategies, the division operation only involves the division between the variable and constant values, so it can be converted into multiplication operation.

The calculation amount statistics of \( \tau_{0} \) are merged into the step of \( B_1, B_2 \) calculation in Table 3 and 4. It should be noted that the calculation of \( \lambda \) and \( k_e \) does not require real-time measured values. Therefore, the variables \( \lambda \) and \( k_e \) do not need to be calculated online, and the corresponding calculation amount statistics are not included in the tables.

It can be seen from the statistical results that the calculation amount of “+, −”, “×”, and “\( f(x) \)” in the ET-PTC is significantly smaller than that of the T-PTC, and the calculation amount of “>, <” is slightly higher than the T-PTC. The calculation amount of “+, −”, “×”, and “\( f(x) \)” in the SP-PTC is further reduced compared to the ET-PTC, and the computation times of “>, <” in the SP-PTC is the same as that in the T-PTC. Therefore, the computation amounts of the three algorithms have: T-PTC > ET-PTC > SP-PTC. Furthermore, by analyzing the computation amounts of each

TABLE 2. The computations amount of each part in T-PTC algorithm.

| Stages               | Components            | + | − | × | f(x) | > | < |
|----------------------|-----------------------|---|---|---|-----|---|---|
| Delay compensation   | Clarke transform      | 3 | 3 |   |     |   |   |
|                      | Park transform        | 2 | 4 | 4 |     |   |   |
|                      | Delay compensation equation | 5 | 9 |   |     |   |   |
| Current prediction   | Park transform        | 14| 28| 28|     |   |   |
|                      | Current prediction equation | 35| 63|   |     |   |   |
| Torque & flux        | Flux prediction       | 14| 28| 7 |     |   |   |
| Prediction           | Torque prediction     | 7 | 21|   |     |   |   |
| Cost function        | -                     | 21| 28|   |     |   |   |
| Optimization         | -                     | 101| 184| 39| 7 |   |   |
| Total                | -                     | 58| 94| 37| 10 |   |   |

TABLE 3. The computations amount of each part in ET-PTC algorithm.

| Stages               | Components            | + | − | × | f(x) | > | < |
|----------------------|-----------------------|---|---|---|-----|---|---|
| Delay compensation   | Clarke transform      | 3 | 3 |   |     |   |   |
|                      | Park transform        | 2 | 4 | 4 |     |   |   |
|                      | Delay compensation equation | 4 | 4 |   |     |   |   |
| Torque & flux        | Flux estimation       | 2 | 2 | 2 |     |   |   |
| estimation           | Torque estimation     | 1 | 3 | 3 |     |   |   |
|                       | PLL calculation       | 8 | 4 | 3 |     |   |   |
| \( R_1, R_2 \)       | calculation           | - | 5 | 7 | 1  |   |   |
| Cost function        | -                     | 35| 63| 28|   |   |   |
| Optimization         | -                     | 7 |   |   |     |   |   |
| Total                | -                     | 58| 94| 37| 10 |   |   |

TABLE 4. The computations amount of each part in SP-PTC algorithm.

| Stages               | Components            | + | − | × | f(x) | > | < |
|----------------------|-----------------------|---|---|---|-----|---|---|
| Delay compensation   | Clarke transform      | 3 | 3 |   |     |   |   |
|                      | Park transform        | 2 | 4 | 4 |     |   |   |
|                      | Delay compensation equation | 4 | 4 |   |     |   |   |
| Torque & flux        | Flux estimation       | 2 | 2 | 2 |     |   |   |
| estimation           | Torque estimation     | 1 | 3 | 3 |     |   |   |
|                       | PLL calculation       | 8 | 4 | 3 |     |   |   |
| \( R_1, R_2 \)       | calculation           | - | 4 | 6 |     |   |   |
| \( \psi' \)          | calculation           | - | 2 | 4 | 4  |   |   |
| Optimization         | Partition method      | 6 | 7 | 4 |     |   |   |
| Total                | -                     | 30| 41| 12| 7  |   |   |

FIGURE 15. Stator currents, torque, stator flux modulus, and common mode voltage of the SP-PTC under different setting values of weighting factor when adding the common mode voltage constraints.
stage in the PTC algorithm, it can be seen from the tables that the computation amounts of the three algorithms in the delay compensation stage are basically the same. The traditional PTC needs a large amount of computations in the stages of currents prediction, torque & flux prediction and cost function; while the ET-PTC replaces the prediction processes with the torque & flux linkage estimation, $B_1$ & $B_2$ calculation, which greatly reduces the amount of computations. Based on the ET-PTC, the SP-PTC further reduces the computational complexity of the cost function by using the partition method.

Figs. 16-18 show the practical measurement results of the computation time at each step in the traditional PTC and the two simplified PTC. Fig. 19 shows the proportion of computation time spent on each step in the three algorithms. Fig. 20 shows the computation time of the SP-PTC before and after adding the additional constraints.

The control periods of the traditional PTC and the two simplified PTC are all set to 50µs. In Figs. 16-18, the measurement process for three-phase currents, DC voltage, and rotor position are required and then placed before the PTC algorithm. The corresponding execution time is approximately equal to 12.37µs because of the oversampling technology used to reduce the noise during the A/D conversion procedure. The PTC algorithm is composed of five stages, corresponding to Table 2-Table 4. The time consumption of T-PTC, ET-PTC and SP-PTC are 22.31µs, 18.86µs, 8.93µs, respectively, which proves that the simplified algorithms effectively reduce the execution time of the PTC algorithm. In addition, the time consumption of each stage is also measured. The computation time spent on the delay compensation stage in T-PTC, ET-PTC, and SP-PTC are 2.38µs, 2.32µs, 2.32µs, respectively. The computation time of prediction stage in T-PTC is 16.48µs (12.62µs for current prediction, 3.86µs for torque and stator flux prediction); while the
ET-PTC greatly reduces the corresponding computation time to 5.53\(\mu\)s (4.46\(\mu\)s+1.07\(\mu\)s), but time consumption in cost function increases from 2.32\(\mu\)s to 9.88\(\mu\)s. The SP-PTC simultaneously reduces computation time in the prediction and cost function calculation. In SP-PTC, the computation time of the stages (2)-(4) which are used to replace the prediction procedure in T-PTC is reduced to 5.54\(\mu\)s; while the computation time of the stage (5) which is used to replace the cost function and numerical comparison procedure in T-PTC is reduced to 1.07\(\mu\)s. Moreover, considering that the control frequency in the three algorithms is set to 20 kHz, one can find that the interrupt program execution time of the ET-PTC is reduced to 31.22\(\mu\)s, so the corresponding control frequency can be set to 30 kHz, which is higher than the original control frequency achieved by the T-PTC. The interrupt program execution time of the SP-PTC is further reduced to 21.30\(\mu\)s, which can increase the control frequency to 40kHz. Besides, in the applications with low environmental electromagnetic interference or low control accuracy requirement, oversampling techniques can be removed from the program to reduce the time consumption in the measurement stage and further increase the control frequency. With the rapid development of silicon carbide (SiC) power devices, the simplified PTC algorithm can achieve higher switching frequency, thereby improving the performance of torque and stator flux in the motor drive system. For traditional Si MOSFET, the proposed algorithm is also applicable. Benefiting from the reduction of the calculation amount, the control period has more idle time, which can realize other functions, such as sensorless control and motor fault detection. In addition, since the amount of calculation is reduced, PTC control algorithms can be applied to microcontrollers with lower processing speed and lower cost, and such low-cost motor driver designs usually use Si MOSFETs.

In Fig. 20, because the partition method is replaced by the modified cost function in the SP-PTC with additional constraints, the measured computation time will be higher than that expected. The experimental results show that the execution time of the SP-PTC with common mode voltage constraint is 15.41\(\mu\)s, while execution time of SP-PTC the with switching frequency constraint is 17.8\(\mu\)s, which is higher than the 8.93\(\mu\)s of the SP-PTC without additional constraints. Even so, these results are still lower than the computation time of 22.31\(\mu\)s in the T-PTC without additional constraints, which proves that the simplified algorithm can effectively reduce the execution time of the PTC with additional constraints.

V. CONCLUSION
In this paper, by analyzing the characteristics of PTC, the ET-PTC based on equivalent transformation is proposed. Different from other simplified algorithms, the ET-PTC completely maintains the characteristics of traditional PTC with simpler control structure. On this basis, the multi-cycle torque & flux prediction is transformed into single voltage vector prediction with approximate conditions, and the partition method is present to simplify the optimal voltage vector searching process. The simplification procedure uses a stepwise method, and it reveals the relationship between the traditional PTC and the simplified PTC. Furthermore, it discusses whether the simplification procedure will lose some features of the traditional PTC.

In the experiment, the transient performance, steady-state error, and weighting factor adjustment ability of the T-PTC, ET-PTC and SP-PTC are evaluated. The additional constraint compatibility of the simplified algorithms is also tested. Finally, the computation complexity and time consumption in each stage of the three algorithms are evaluated in detail. The experimental results show that the ET-PTC completely maintains the performance and characteristics of the traditional PTC while reducing the computation time. The SP-PTC further reduces the algorithm complexity and time consumption in the procedure of cost function calculation and optimization, and its control performance is almost same with the traditional PTC, although it loses the feature about adjusting the tendency of the control targets. Considering about the difference between the ET-PTC and SP-PTC, the users can choose the appropriate simplified PTC strategy according to the requirements of the control system application. Besides, the simplified PTC algorithms have good compatibility with additional constraints.

Compared with other simplified algorithms, this paper provides the detailed simplification procedure for the PTC, and it also describes concrete steps for the implementation of the simplified algorithm. The theoretical derivation and experimental results reveal the impact of the simplification procedure on the performance and characteristics of the traditional PTC. Moreover, the experimental results give comprehensive evaluation of the performance about the simplified PTC. The computation burden statistics and practical measured time consumption show that the simplified PTC effectively reduces the computation amount and the execution time of the control strategy, which will be convenient for its application in low-cost microcontrollers. Moreover, the simplified PTC algorithm can achieve higher switching frequency, thereby improving the performance of torque and stator flux in the motor drive system with the application of the new generation SiC power device.

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