Virtual-Vector-Based FCS Model Predictive Current Control with Duty Cycle Optimization for Dual Three-Phase Motors

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Abstract: This paper presents an improved finite control set model predictive current control (FCS-MPC) based on virtual vectors with duty cycle control for dual three-phase permanent magnet synchronous motor (DTP PMSM) drives. Virtual vectors synthesized by two basic voltage vectors of the inverter are introduced to suppress the harmonic current by the vector space decomposition (VSD) model. And the virtual vectors with the optimal duty cycle can be obtained by duty cycle control using zero vectors to generate a fixed switching frequency. Moreover, a simple scheme of centrosymmetric switching patterns is presented for the real-time implementation. In this way, the performance of current tracking is improved while the computational burden is reduced. Finally, results of simulation analysis and DSP test for the three different strategies are given to verify the validity of the proposed strategy.

1. Introduction

Electrified vehicles (EVs) are catching a worldwide attention during the past decades owing to its good propulsion performance, high energy efficiency, and low-carbon emissions [1]. Electric propulsion system with low vibration and high robustness has always been one of the focuses in EVs. In recent years, multiphase machine drive systems have been considered as promising candidates for EVs, benefiting from their low rating current per phase, suppressed torque ripple, and good fault-tolerance capability. Dual three-phase permanent magnet synchronous motors (DTP PMSMs), namely asymmetrical six-phase machines, have the most common and simplest topology of multi three-phase machines [2-4].

Model predictive control (MPC) is getting a sustained development in motor drive field due to the merits of optimal control with multivariable constrained, fast dynamic response, flexible handling ability of system nonlinearities. Normally, MPC strategies have been categorized into two types: continuous control set MPC (CCS-MPC) and finite control set MPC (FCS-MPC). Due to the simple control principle and modulation-free characteristic, the FCS-MPC has been successfully applied in multiphase drives, such as five-phase drives and dual three-phase drives [5-7].

Virtual vectors are first proposed for three-phase invertors to suppress the torque ripples and improve the flux predictive control performance. An easy implement was also achieved in a fully digital control system in [8]. Virtual vectors are synthetic vectors by combining two basic voltage vectors of the VSI. In recent years, virtual vectors are utilized in the control algorithm of dual three-
phase motors to suppress the harmonic currents and steady-state errors [9]. A predictive current control scheme based on virtual vectors was presented in [10], improving the current tracking. And switching patterns were modulated with fixed switching frequency in [11]. However, because of the high calculation burden, it is difficult to be applied in a real-time control system with high sampling frequency.

In this paper, a novel virtual-vector-based FCS-MPCC for a DTP PMSM is proposed, including improved duty cycle control to optimize control voltage vectors. Compared with the conventional FCS-MPCC, the duty cycle control strategy can cover the shortage of local optimal control vectors and improve the current tracking performance. In addition, the computational burden is reduced by off-line optimizing and looking up the switching sequence table on-line. And consequently, the running time of the algorithm on a digital signal processor (DSP) is reduced.

2. Two-Step Virtual-Vector-Based FCS Model Predictive Current Control Scheme of DTP PMSM

In this section, the conventional two-step FCS-MPCC is introduced, including the vector space decomposition (VSD) model of the asymmetrical DTP PMSM and the synthesis scheme of virtual vectors fed by 2L-VSI.

2.1. VSD Model of asymmetrical DTP PMSM

The asymmetrical DTP PMSM involves in two sets of Y-type three-phase stator windings shifted by 30 electrical degrees with separated neutral points, which is indicated by \((A, B, C)\) and \((X, Y, Z)\) as illustrated in figure 1. This type of motor is usually used to improve the machine performance because of its low torque ripple and excellent fault tolerance.

\[
\begin{bmatrix} d & q & z_1 & z_2 \end{bmatrix}^T = T_{6s/dqz1z2} \cdot [A \ B \ C \ X \ Y \ Z]^T
\]

where \(T_{6s/dqz1z2}\) is the VSD transformation matrix in [12].

According to the VSD method, the mathematical model of the DTP PMSM can be presented as

\[
\frac{d}{dt} \begin{bmatrix} \dot{i}_d \\ \dot{i}_q \end{bmatrix} = A_t \begin{bmatrix} \dot{i}_d \\ \dot{i}_q \end{bmatrix} + B_t \begin{bmatrix} \dot{u}_d \\ \dot{u}_q \end{bmatrix} + f_t
\]

\[
\frac{d}{dt} \begin{bmatrix} \dot{i}_{z1} \\ \dot{i}_{z2} \end{bmatrix} = C_t \begin{bmatrix} \dot{i}_{z1} \\ \dot{i}_{z2} \end{bmatrix} + D_t \begin{bmatrix} \dot{u}_{z1} \\ \dot{u}_{z2} \end{bmatrix}
\]

\[
T_e = 3p_0 (\dot{i}_q \psi_d - i_d \psi_q)
\]

where

\[
A_t = \begin{bmatrix} -R_s/L_D & \omega L_q/L_D \\ -\omega L_d/L_D & -R_s/L_q \end{bmatrix}, \quad B_t = \begin{bmatrix} 1/L_D \\ 1/L_Q \end{bmatrix}, \quad f_t = \begin{bmatrix} 0 \\ -\omega \psi_f/L_Q \end{bmatrix},
\]

\[
C_t = \begin{bmatrix} -R_s/L_Z & 0 \\ 0 & -R_s/L_Z \end{bmatrix}, \quad D_t = \begin{bmatrix} 1/L_D \\ 1/L_Q \end{bmatrix}.
\]
\(i_d, i_q, i_{x1}, i_{x2}\) represent the currents, \(u_d, u_q, u_{x1}, u_{x2}\) represent the voltage, \(R_s\) is the stator resistance, \(L_D, L_Q, L_Z\) are the inductances, \(\psi_d, \psi_q, \psi_f\) are flux linkages, and \(\omega\) represents the electric angular speed of motor.

2.2. Classical FCS-MPCC based on virtual vectors for DTP PMSM driven by 2L-VSI

The topology of the 2L-VSI driving the DTP PMSM consists of six phase legs, and each leg is characterized by two power switches. For simplicity, \(S_j (j = a, b, c, x, y, z)\) is defined as the switching status of each phase leg, where “1” indicates the upper switch is ON, while “0” indicates the lower switch is ON. According to the combinations of different switching states, 64 voltage vectors can be generated by the inverter, and the voltages mapped into two decomposed stationary reference frames can be calculated by

\[
v_{ab} = \frac{U_{dc}}{3} \left[ e^{j0} \ e^{j\frac{2\pi}{3}} \ e^{j\frac{4\pi}{3}} \ e^{j\frac{\pi}{6}} \ e^{j\frac{5\pi}{6}} \ e^{j\frac{3\pi}{2}} \right] S_j
\]

\[
v_{z1z2} = \frac{U_{dc}}{3} \left[ e^{j0} \ e^{j\frac{2\pi}{3}} \ e^{j\frac{4\pi}{3}} \ e^{j\frac{\pi}{6}} \ e^{j\frac{5\pi}{6}} \ e^{j\frac{3\pi}{2}} \right] S_j
\]

where \(U_{dc}\) is the dc line voltage, and \(S_j = [S_a \ S_b \ S_c \ S_x \ S_y \ S_z]^T\). \(v_{z1z2}\) is the voltage in \(z1-z2\) subspace, and \(v_{ab}\) is the voltage in \(\alpha-\beta\) subspace, which can be transformed into \(d-q\) subspace with Park transformation

\[
\begin{bmatrix}
  d \\
  q
\end{bmatrix} = \begin{bmatrix}
  \cos \theta & \sin \theta \\
  -\sin \theta & \cos \theta
\end{bmatrix}
\begin{bmatrix}
  a \\
  b
\end{bmatrix}
\]

(7)

The active voltage vectors in two subspaces obtained by (5) and (6) are shown in figure 2. The four groups of active voltage vectors have different phases in \(\alpha-\beta\) and \(z1-z2\) subspaces. Large vector \(v_L\) and medium-large vector \(v_{ML}\) have the same phase in \(\alpha-\beta\) subspace, for instance, but opposite phases in \(z1-z2\) subspace, so do \(v_{ML}\) and \(v_S\). Therefore, virtual voltage vectors can be combined by such two vectors so that their components in \(z1-z2\) subspace are zero. From the control point of view, the harmonic current in \(z1-z2\) subspace can be suppressed in this way. Assuming that the duty cycles of voltage vectors \(\{v_L, v_{ML}, v_S\}\) are \(\{d_L, d_{ML}, d_S\}\) during a sampling period. As shown in figure 3, two kinds of virtual vectors can be obtained [13].

Figure 2. Voltage vectors generated by six-phase 2L-VSI in (a) \(\alpha-\beta\) subspace and (b) \(z1-z2\) subspace.
In conventional virtual-vector-based FCS-MPCC scheme for DTP PMSM, only virtual vectors VVI are applied in order to reduce computation burden and achieve high voltage utilization. Since \( u_{x1} = 0, u_{x2} = 0 \) of virtual vectors, the currents in \( \alpha-\beta \) subspace are zero according to (3). Applying the forward Euler method to discretize (2), the predictive values of the currents can be obtained by

\[
\begin{bmatrix}
i_d^p(k + h) \\
i_q^p(k + h)
\end{bmatrix} = (I_2 + T_s A_k) \begin{bmatrix}
i_d(k + h - 1) \\
i_q(k + h - 1)
\end{bmatrix} + T_s B_k \begin{bmatrix}
u_d(k + h - 1) \\
u_q(k + h - 1)
\end{bmatrix} + T_s f_k
\]  

(8)

where

\[
A_k = \begin{bmatrix}
-R_s/L_D & \omega_k L_Q/L_D \\
-\omega_k L_D/L_Q & -R_s/L_Q
\end{bmatrix}, f_k = \begin{bmatrix}
0 \\
-\omega_k \psi_f/L_Q
\end{bmatrix}, I_2 = \begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix}, B_k = B_t, C_k = C_t, D_k = D_t,
\]

\( h \) is the prediction horizon, \( T_s \) is the sampling period, and \( \omega_k \) is the speed in instant \((k + h - 1)\).

Due to time delay between sampling and control output in the practical control system, delay compensation is needed to improve the prediction accuracy for MPCC. According to (8), let horizon \( h = 1 \) and we can obtain the currents \((i_d, i_q)\) in instant \((k + 1)\) using the known current and voltage vectors in instant \( k \). After that, the prediction currents in instant \( k + 2 \) can be calculated with horizon \( h = 2 \). In order to track the reference currents quickly and accurately, the cost function is defined as

\[
g = |i_d^* - i_d^p| + |i_q^* - i_q^p|
\]

(9)

where the currents with superscript * represent reference values, superscript \( p \) represents prediction values. For a DTP PMSM, in order to guarantee the maximum torque per ampere (MTPA), the reference value \( i_d^* \) is set to zero. In classical MPCC, the value of \( i_q^* \) is imposed by the speed PI regulator. Hence, MPCC is converted to the following optimization problem, i.e., the optimal switching state vector \( S(k + 2) \) need to be found to minimize the cost function

\[
g_1 = |i_d^p| + |i_q^* - i_q^p|
\]

\[
S(k + 2) = \arg\min_{\{VVI_1, VVI_2, \ldots, VVI_{12}, 0\}} (g_1)
\]

(10)  

(11)

3. The Proposed MPCC with Duty Cycle Control

In FCS-MPCC, thirteen groups of predicted currents corresponding to twelve virtual vectors and zero vector can be obtained by (8) during each sampling period, and the virtual vector \((VVI_{\min})\) that minimizes \( g_1 \) in (10) will be selected by exhaustive method (original VV-MPCC). However, the
control voltage vector $VV^{\text{min}}$ is just a local optimal solution, and the performance of original VV-MPC in steady state can be further improved.

Duty cycle control (DCC) is proposed here by combining zero vectors and virtual vectors, so that the optimal duty cycle of the virtual vectors can be obtained. After injecting the zero vectors $\{v_0, v_7\}$ with duty cycle $d_0$ into the virtual vector $VV^{\text{min}}$, the equivalent optimal voltage vector $V_s^{\text{opt}}$ becomes

$$V_s^{\text{opt}} = d_v VV^{\text{min}} + d_0 V_0 = d_v VV^{\text{min}}$$

where

$$V_s^{\text{opt}} = \begin{bmatrix} u_d^{\text{opt}} \\ u_q^{\text{opt}} \end{bmatrix}, VV^{\text{min}} = \begin{bmatrix} u_d^{\text{min}} \\ u_q^{\text{min}} \end{bmatrix}, V_0 = [0, 0]^T, d_v + d_0 = 1,$$ 

$d_v$ represents the duty cycle of the virtual vector $VV^{\text{min}}$ during a sampling period.

Apply (12) to (8), new predictive currents in instant $(k + 2)$ can be obtained as $i_d^{\text{pnew}}(k + 2)$ and $i_q^{\text{pnew}}(k + 2)$. In order to track the reference currents with minimum tracking errors, a new cost function is defined as

$$g_2(d_v) = (i_d^* - i_d^{\text{pnew}}(k + 2))^2 + (i_q^* - i_q^{\text{pnew}}(k + 2))^2$$

From a mathematical point of view, the optimal duty cycle $d_v$ can be calculated by solving the extreme point of (13), which is a classical quadratic optimization problem, namely,

$$\frac{\partial g_2}{\partial d_v} = 0$$

$$d_v = \left( e_d h_d + e_q h_q \right) + m_d i_d T_s + m_q i_q T_s + m_f h_q T_s \left( h_d^2 + h_q^2 \right) T_s \tag{15}$$

where

$$i_d = i_d(k + 1), i_q = i_q(k + 1),$$

$$e_d = i_d^*(k) - i_d, e_q = i_q^*(k) - i_q,$$

$$h_d = \frac{u_d^{\text{min}}}{L_D}, h_d = \frac{u_q^{\text{min}}}{L_Q},$$

$$m_d = \frac{\omega_k L_D}{L_Q} h_q + \frac{R_s}{L_D} h_d, m_q = \frac{R_s}{L_Q} h_q - \frac{\omega_k L_Q}{L_D} h_d, m_f = \frac{\omega_k \psi_f}{L_Q} h_q.$$ 

The duty cycle $d_v$ is only related to the virtual vector component in $d-q$ subspace in each control period. Consequently, once the virtual vector ($VV^{\text{min}}$) is figured out, zero vector duty cycle $d_0$ can be easily obtained by a further calculation using (15-17). For the real-time operation, $d_v$ and $d_0$ are constrained by the saturation algorithm as follows

$$d_v = \begin{cases} 0 & , & d_v \leq 0 \\ d_v & , & 0 < d_v < 1 \\ 1 & , & d_v \geq 1 \end{cases}$$

$$d_0 = 1 - d_v \tag{17}$$

Since the standard PWM waveforms symmetrical to the middle of the sampling period are easily produced by DSPs, the optimal control voltage vector obtained by (12) needs to be transformed into an adequate switching pattern before being generated by the inverter. The switching patterns are modified according to the simple method in [14]. Taking virtual vectors $\{VV_1, VV_2, VV_3, VV_4\}$ with the optimal duty cycle $d_v$ as examples, the switching patterns are demonstrated in figure 4. It is obvious that the modified switching signals meet the requirements of real-time systems such as DSP and dSPACE. Moreover, a fixed switching frequency can be maintained through combining virtual vectors and zero vectors, and the switching frequency is $1/T_s$. Each phase switching duty cycle of all twelve virtual vectors is presented in table 1.
Define the switching duty cycle of phase \( j (j = a, b, c, x, y, z) \) in table 2 as \( d_j \), and the actuation time of the upper switch during each switching period can be calculated as

\[
t_j^u = \left[ 1 - (d_j d_x + d_0/2) \right] T_s/2
\]  

(18)

The overall control scheme of the proposed MPCC based on virtual vectors with duty cycle control is given in figure 5, and the corresponding algorithm flowchart for the real-time implementation is shown in figure 6.
Figure 5. The proposed virtual-vector-based FCS-MPCC with duty cycle control.

Figure 6. The real-time flowchart of the proposed algorithm.

4. Results Analysis

4.1. Simulation verification

The simulation analysis is carried out based on the model of the DTP PMSM and algorithms using Matlab/Simulink. The parameters of the dual three-phase PMSM are listed in Table I, and the sampling frequency of the control algorithm is 10 kHz.

To evaluate the feasibility and effectiveness of the proposed method, MPCC based on virtual vectors with DCC (Method 3), several simulation results are obtained to compare Method 3 with MPCC based on virtual vectors without DCC (Method 1), and MPCC based on virtual vectors with variable amplitude (Method 2) proposed in [11]. The results are compared side by side and presented in figure 7.
All the three strategies are able to track current references. Comparing the results, we can see that Method 2 and Method 3 are superior to Method 1 in terms of current tracking and torque. In order to quantitatively evaluate the performance, the mean error of a feedback and reference value is adopted here

\[ e(x) = \frac{1}{N} \sum_{i=1}^{N} |x_i^* - x_i| \times 100\% \]  

(19)

where \( x \in \{i_q, T_e\}, x^* \in \{i_q^*, T_L\} \), \( N \) is the number of samples corresponding to a time window of 1 s, and \( \bar{x}^* \) represents the mean of \( x^* \) during the time window. The performance indicator according to (19) is summarized in Table 3.

| Parameter                        | Value |
|----------------------------------|-------|
| DC supply \( U_{dc} \) (V)       | 270   |
| Stator resistance \( R_s \) (Ω)  | 1.096 |
| d/q-axis inductance \( L_D/L_Q \) (mH) | 2.28/2.28 |
| Leakage self-inductance \( L_Z \) (mH) | 1.02  |
| Permanent magnet flux \( \psi_f \) (Wb) | 0.0752 |
| Pole pairs \( n_p \)             | 5     |

Figure 7. Simulation performance of a DTP PMSM operating at 1000 rpm and 10 Nm (\( T_L \)) with three control strategies. (a) The conventional MPCC based on virtual vectors (Method 1). (b) MPCC based on virtual vectors with variable amplitude [11] (Method 2). (c) The proposed MPCC based on virtual vectors with DCC (Method 3).

In Method 1, virtual vectors are directly used for predictive current control, but only the local optimal solution is obtained. Although the current tracking is realized, the ripple of current and torque is very large. Through variable amplitude and duty cycle control, the optimal solution of the cost function can be obtained theoretically and current tracking errors are reduced in Methods 2 and 3. In Method 2, the duty cycle of current vectors is equivalent to that of voltage vectors, as given in (20), and used for control, so the performance is not the best. Method 3 directly designs the cost function considering the voltage duty cycle, and the optimal duty cycle obtained is applied to virtual voltages duty cycle control. Hence, Method 3 provides the best current reference tracking and leads to the lowest torque ripple.
\[ f = \left( i_d^* - d_0 i_{d,0}(k+2) - d_v i_d(k+2) \right)^2 + \left( i_q^* - d_0 i_{q,0}(k+2) - d_v i_q(k+2) \right)^2 \]  

(20)

Moreover, the cost function of Method 2 is given in (20), where \( i_{d,0}, i_{q,0} \) represent the predicted currents due to the application of a zero voltage vector during a sampling period. Since computational burden by Method 2 will be increased by solving \( i_{d,0}, i_{q,0} \) before obtaining the optimal duty cycle \( d_v \), the proposed strategy in this paper also shows an advantage in saving computing resources in a practical application.

Table 3. Performance indicator to evaluate the three methods under 1000 rpm and 10 Nm.

| Mean Error | Method 1 | Method 2 | Method 3 |
|------------|----------|----------|----------|
| \( e(i_q) \) | 7.37%    | 2.11%    | 1.42%    |
| \( e(T_v) \) | 5.55%    | 1.36%    | 0.57%    |

4.2. Running time test on a DSP

Running time of the above three strategies is tested on a real time platform. The DSP utilized is 32-bit TMS320F28335 produced by Texas Instruments (TI) (Dallas, TX, USA), with a system clock up to 150 MHz. Table 4 lists the results of running time in one control cycle.

Table 4. Running time of the three methods on the DSP.

| Strategy     | Method 1 | Method 2 | Method 3 |
|--------------|----------|----------|----------|
| Running time (\( \mu s \)) | 88.17    | 96.13    | 95.22    |

Method 1 with the simplest strategy has the shortest running time. By contrary, Method 2 shows the worst performance in time consuming due to the extra calculation to solve (20). Therefore, Method 3 offering better current tracking and consuming less time is the most useable in real-time system. Last but not least, the running time could be decreased adapting a more powerful DSP, like dual-core TMS320F28379D with system clock up to 200 MHz of two CPUs, and then the switching frequency could be increased to improve the control performance.

5. Conclusions

This paper proposes an optimal duty cycle control strategy to improve the performance of the conventional virtual-vector-based FCS-MPCC. Two cost functions are designed, and the optimal control voltage vector is obtained by the two-step optimization: the first step optimization is to get the local optimal virtual vector, and then, the optimal duty cycle of the virtual vector is obtained by the second step optimization. After that, the switching patterns are modified in a simple way to generate the control voltage by a 2L-VSI. Moreover, the excellent current tracking performance is verified via simulation analysis, and the DSP test shows the short running time of the proposed strategy.

In fact, apart from the current tracking error, the phase angle error of the control voltage also has an effect on the steady-state performance of the motor. In the future work, we will take the phase angle of virtual vectors into account in cost function design, so as to improve the MPCC.

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