Model predictive current control for dual three-phase PMSM with hybrid voltage vector
Fenghuang Cai\textsuperscript{1,a}, Fuyang Yang\textsuperscript{1}, Qin Qin Chai\textsuperscript{1}, and Jiahui Jiang\textsuperscript{1,b)

Abstract To improve the steady-state performance of the dual three-phase permanent magnet synchronous motor with high torque ripple and high harmonic current, this paper proposes a hybrid voltage vector model predictive current control algorithm (MPCC). Firstly, based on the virtual voltage vectors synthesized using the vector characteristics of the fundamental and harmonic subspaces, hybrid voltage vectors are synthesized from the virtual voltage vectors and the zero vector to increase the voltage vector amplitude range to reduce torque ripple and to suppress harmonic currents. Then a vector selection method is proposed to reduce the number of alternative vectors and the calculation burden of the MPCC. Finally, the realization of corresponding PWM modulation is given. The simulation results show that the method effectively suppresses harmonic currents and torque ripple and increases the steady-state performance of the motor.

Keywords: dual three-phase permanent magnet synchronous motor; model predictive control; hybrid voltage vector; predictive current control.

Classification: Power devices and circuits

1. Introduction

Three-phase permanent magnet synchronous motor (PMSM) is widely used in industry due to its simplicity of control [1]. Compared to three-phase motors, the dual three-phase PMSM has the advantages of high power density, high efficiency, low torque ripple and high reliability [2,3,4,5]. As in the case of other multi-phase motors, dual three-phase PMSM has a low-impedance harmonic plane where a smaller harmonic voltage can produce a larger harmonic current. Therefore, in addition to controlling the electromagnetic torque of the motor, the harmonic plane also needs to be controlled to obtain better operating performance [6,7].

There are quite a few papers in the literature devoted to studying the control problems arising from the dual three-phase PMSM that have been shown to achieve excellent steady-state performance [8,9,10,11,12]. In [13], a four-vector SVPWM control based on decoupled field-oriented control combined with space vector modulation techniques is used to reduce harmonic currents and achieve excellent steady-state performance. However, the generated PWM waveforms are not aligned bringing challenges to implement in DSP, such as increasing the switching frequency and exacerbate device losses. In [14,15], direct torque control is used to obtain a fast torque response during acceleration, deceleration, and load change dynamics. Although direct torque control has the advantages of simple structure, fast torque response, and high robustness, torque ripple, and harmonic currents are large [16,17].

Model Predictive Control (MPC) for motor has become popular in the last decade due to its fast response capability [18,19,20,21,22]. Although conventional MPCC has shown significant improvements in the system's dynamic performance, the problems of high torque ripple and high harmonic components still exist. In addition, the high computational burden brings a challenge to realize the controller [23]. For the harmonic problem in the motor, reference [24] proposes virtual vectors to synthesize the fundamental voltage vector so that its amplitude in the harmonic subspace is zero to suppress harmonic currents. However, the amplitude magnitude of the virtual vector is still limited and the torque ripple of the motor is still large. In [25], a model predictive torque control (MPTC) based on virtual vectors is proposed to control the torque directly. It can effectively suppress torque ripple and magnetic chain ripple problems, but the weighting factors for magnetic chain and torque in the cost function are difficult to determine. In [26], an MPTC, with virtual vectors and different cost functions attained according to the torque state of the system at two adjacent moments, is built to reduce harmonic currents and solve the difficulty of determining the weighting coefficients in the cost function. But this MPTC slightly increases the torque ripple.

In this paper, an MPCC with hybrid voltage vectors is proposed to address the above problems. The virtual...
vectors are first used to eliminate harmonic components, and then the vectors are further synthesized to widen the vector amplitude range. By judging the position of the reference voltage, the number of candidate voltage vectors is reduced, so as to reduce the computational burden. Moreover, the PWM module is designed to generate the optimal output voltage. The effectiveness of the change of control strategy is verified by simulation.

2. Dual three-phase PMSM mathematical model

In this paper, a dual three-phase PMSM shown in Fig. 1 consists of two sets of star-shaped three-phase windings 30° apart in space, the first set being ABC and the second set DEF.

![Fig. 1 Dual three-phase PMSM topology](image)

The mathematical model in the synchronous frame is [27]:

\[
\begin{bmatrix}
\begin{array}{c}
u_d \\ \nu_q
\end{array}
\end{bmatrix} = \begin{bmatrix}
\nu_d \\ \nu_q
\end{bmatrix} = T_{dq}\begin{bmatrix}
\nu_d \\ \nu_q
\end{bmatrix}
\]

(1)

While \(u_d, u_q\) are the stator voltage components in the \(d\)- and \(q\)-axis; \(T_{dq}\) is the Park matrix, and \(T_{dq} = \begin{bmatrix}
\cos \theta & \sin \theta \\
-\sin \theta & \cos \theta
\end{bmatrix}\)

\(u_d, u_q, u_r, u_{r'}, u_d, u_{q'}, u_r, u_{r'}\)

\(= T_{dq} \begin{bmatrix}
u_d \\ \nu_d
\end{bmatrix}
\)

(2)

\(= T_{dq} \begin{bmatrix}
u_d \\ \nu_d
\end{bmatrix}\)

(3)

where \(\alpha, \beta\) is the rotor electrical angle, \(u_i, \alpha, \beta\), \(i = A, ..., F\) are the stator voltage components in the nature coordinate system, and \(T_{dq}\) is the VSD transformation matrix and

\[
T_{dq} = \frac{1}{3}
\begin{bmatrix}
1 & -1 & -1 & \sqrt{3} & \sqrt{3} & 0 \\
0 & \sqrt{3} & -\sqrt{3} & 1 & 1 & -1 \\
1 & -1 & -1 & -\sqrt{3} & \sqrt{3} & 0 \\
0 & \sqrt{3} & -\sqrt{3} & 1 & 1 & -1 \\
1 & 1 & 1 & 0 & 0 & 0 \\
0 & 0 & 0 & 1 & 1 & 1
\end{bmatrix}
\]

(4)

The voltage equations in \(d-q\) plane and in the \(x-y\) plane are as follows:

\[
\begin{bmatrix}
u_d \\ \nu_q
\end{bmatrix} = \begin{bmatrix} R + L_d P & -\omega L_q \\ \omega L_d & R + L_q P \end{bmatrix}\begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega \psi_f \end{bmatrix}
\]

(5)

\[
\begin{bmatrix}
u_e \\ \nu_f
\end{bmatrix} = \begin{bmatrix} R \ 0 \\ 0 \ R \end{bmatrix}\begin{bmatrix} i_e \\ i_f \end{bmatrix} + \omega P\begin{bmatrix} i_e \\ i_f \end{bmatrix}
\]

(6)

where \(i_d, i_q\) are the components of the stator current in the \(d\)- and \(q\)-axis; \(L_d, L_q\) are the inductances in the \(d\)- and \(q\)-axis; \(\omega\) is the rotor electric angular velocity; \(\psi_f\) is the permanent magnet magnetic chain; \(R\) is the stator resistance; \(\nu_d, \nu_f\) are the voltage components in the \(x-y\) plane; \(i_e, i_f\) are the stator current components in \(x-y\) plane; \(L_e\) is the leakage self-inductance; \(P\) is the time derivative operator.

The equation for the magnetic chain is:

\[
\begin{bmatrix}
\psi_d \\ \psi_q
\end{bmatrix} = \begin{bmatrix}
L_d & 0 & 0 \\ 0 & L_d & 0
\end{bmatrix}\begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \psi_f \\ \psi_f \end{bmatrix}
\]

(7)

where \(\psi_d, \psi_q\) are the flux in the \(d\)- and \(q\)-axis; \(\psi_s, \psi_r\) are the flux in the \(x-y\) plane; the torque equation for a dual three-phase PMSM in the \(d-q\) plane is:

\[
T_e = 3n_p [\psi_d i_d - \psi_q i_q]
\]

(8)

where \(T_e\) is the electromagnetic torque; \(n_p\) is the number of pole pairs.

![Fig. 2 Basic voltage vectors](image)

The dual three-phase inverter has six bridges with a total of \(2^3 = 64\) switching states, each corresponding to a voltage vector [28]. However, remove the vectors in repeated positions, there are only 48 active voltage vectors. These voltage vectors can be divided into four groups: \(L1(0.644U_{dc}), L2(0.47U_{dc}), L3(0.33U_{dc}),\) and \(L4(0.172U_{dc})\). With VSD theory, these voltage vectors are divided into three mutually perpendicular spaces \(\alpha-\beta, x-y, \alpha-\alpha'\), as the star connection of the windings \(\alpha-\alpha'\) plane is zero. The basic voltage vector distribution is shown in Fig. 2. In order to improve the utilization of the voltage source on the DC side, the L1 group vectors with the highest amplitude in the \(\alpha-\beta\) plane are selected as the alternative voltage vectors [29].

The voltage Eq. (5) for the motor can be discretized
using the forward Euler equation as follows.

\[\begin{align*}
i_d(k+1) &= i_d(k) + \frac{T_s}{L_d} \left[ \omega_u(k) - R_i i_d(k) + L_{dL} i_q(k) \right] \\
i_q(k+1) &= i_q(k) + \frac{T_s}{L_q} \left[ \omega_u(k) - R_i i_q(k) + L_{qL} i_d(k) \psi_{fL}(k) \right]
\end{align*}\]

(9)

where \( k \) is the \( k \)th sampling moment; \( i_d(k), i_q(k) \) are the sampling currents of the \( d \)-and \( q \)-axis at \( k \)th sampling moment; \( \omega_u(k), \omega_u(k) \) are the sampled voltages of the \( d \)- and \( q \)-axis at \( k \)th sampling moment; \( T_s \) is the sampling period.

The dual three-phase PMSM energy conversion occurs in the \( \alpha-\beta \) subspace, where the vector voltage should be set high and the \( x-y \) subspace is not involved in the energy conversion, the smaller the voltage vector the lower the harmonic content. The cost function can be expressed as:

\[ g = \left| i_d^* - i_d(k+1) \right| + \left| i_q^* - i_q(k+1) \right| \]

(11)

where \( i_d^*, i_q^* \) are the reference current of \( d \)- and \( q \)-axis. To reduce control complexity, \( i_d^* \) is generally set to 0, \( i_q^* \) is determined by the set torque magnitude and obtained from the motor speed through the PI link.

3. MPCC with hybrid voltage vectors

3.1 Vector synthesis

According to Eq. (6), the \( x-y \) plane voltage vector determines the magnitude of the \( x \)- and \( y \)-axis currents. Therefore, if the virtual voltage vectors are constructed to have zero amplitude in the \( x-y \) plane, then the harmonic currents problem can be solved. In addition, from Fig. 2 it can be seen that in the \( \alpha-\beta \) plane V64 and V46 have different amplitudes but the same phase, indicating that V64 and V46 have the same effect on the \( d \)- and \( q \)-axis currents. In the \( x-y \) plane V64 and V46 have opposite phases, which means that they have opposite effects on harmonic currents. All the voltage vectors in groups L1 and L2 have this characteristic, so the virtual voltage vectors will be synthesized using the vectors in groups L1 and L2[30]. Taking V64 and V46 as examples, the action times of the vectors can be expressed as:

\[\begin{align*}
\eta_1 V_{64} - \eta_2 V_{46} &= 0 \\
\eta_1 + \eta_2 &= T_s
\end{align*}\]

(12)

Solving for the above equation calculates that \( \eta_1 = 0.269T_s \), \( \eta_2 = 0.731T_s \).

By synthesizing voltage vectors in groups L1 and L2, a group of virtual voltage vector with a magnitude of 0.597Ud can be constructed. The spatial distribution of the virtual voltage vector with a magnitude of 0.597Ud is shown in Fig. 3. In Fig. 3 plus the zero vector, a total of 13 voltage vectors are divided into 12 sectors. From Fig. 3, we can see that all virtual vectors have the same amplitudes. However, for motor control, mismatched voltage amplitudes will larger torque ripples, or influence the harmonic currents. Thus, how to regular the amplitude of the virtual vector is still a problem to be solved. For this, a hybrid voltage vector method is proposed to supply more amplitudes.

Fig. 3 Virtual voltage vectors

Taking sector I of the virtual voltage vector in Fig. 3 as an example. Sector 1 is subdivided into a 3-layer structure and divided equally into 9 subsectors as shown in Fig. 4. While each vertex of the subsector represents a voltage vector, and each new voltage vector is synthesized by VV12, VV1 and the zero vector.

Fig. 4 hybrid voltage vectors

However, the MPCC brings all alternative voltage vectors into the cost function, resulting in a heavy calculation burden. Thus, a vector selection method will be proposed in the next subsection to reduce the alternative voltage vectors.

3.2 Vector selection

According to Eq. (5), the reference voltage value can be calculated from the reference current, combined with the idea of field-oriented control to initially determine the optimal vector position, reducing the number of times to find the optimal calculation. \( d \) - and \( q \) - axis reference voltage equation is:

\[\begin{align*}
u_d^* &= R_i i_d + L_{dL} \frac{i_d^* - i_d}{T_s} - L_{dL} \omega_s \psi_{fL} \\
u_q^* &= R_i i_q + L_{qL} \frac{i_q^* - i_q}{T_s} + L_{qL} \omega_s \psi_{fL} + \psi_s \omega_s
\end{align*}\]

(13)

To facilitate the search for the optimal vector, the reference voltage needs to be converted to the \( \alpha-\beta \) plane,
which can be expressed as:

\[
\begin{align*}
\bar{u}_a &= \cos \theta u_b - \sin \theta u_c \\
\bar{u}_b &= \sin \theta u_a + \cos \theta u_c
\end{align*}
\]

(14)

According to the reference voltage vector, the magnitude and phase angle of the reference voltage vector in the α-β plane can be expressed as:

\[
V^* = \sqrt{u_a^2 + u_b^2}
\]

(15)

Here \(V^*\) is the amplitude of the reference voltage vector in the α-β plane and \(\theta^*\) is the angle between the reference voltage and the α-axis.

The large sector in which the reference voltage is located can be determined from \(\theta^*\). To simplify the analysis process, only the case when the reference voltage \(U^*\) is located in the sector I is discussed. The case when the reference voltage \(U^*\) is located in the other sectors can be analyzed using the same method by rotating vectors \(N\) times of 30°. Since the VV12 vector has an angle of 15° to the α-axis, the sector \(S\) in which the reference voltage is located can be expressed as:

\[
S = \left[ \frac{\theta^* + 15}{30} \right] + 1
\]

(17)

After the sector \(S\) is determined, only three alternative vectors chosen for the subsector need to be optimized.

To make it easier to find the subsectors, let index \((a,b,c)\) represent the vectors that fall in each subsector. The indexes are numbered using the distance of the reference voltage and the three boundaries. Define the three boundaries of the sector I as \(l_1, l_2, l_3\) respectively, and the distances of the reference voltage corresponding to the three boundaries as \(d_1, d_2, d_3\) respectively, the indexes of these triangles are denoted by \(a, b, c\) as follows:

\[
\begin{align*}
\begin{cases}
1 & 0 \leq d_1, d_2 < x_1 \\
2 & x_1 \leq d_1, d_2 < 2x_1 \\
3 & 2x_1 \leq d_3 < 3x_1
\end{cases}
\end{align*}
\]

(18)

\[
\begin{align*}
\begin{cases}
3 & 0 \leq d_3 < x_2 \\
2 & x_2 \leq d_3 < 2x_2 \\
1 & 2x_2 \leq d_3 < 3x_2
\end{cases}
\end{align*}
\]

(19)

\[
x_1 = \frac{1}{3} \times 0.597U_a \sin 30
\]

(20)

\[
x_2 = \frac{1}{3} \times 0.597U_a \sin 75
\]

(21)

According to Eq. (18-21), each subsector is labeled as shown in Fig. 4. The relationship between the three voltage vectors corresponding to each mark is shown in Table 1. The coefficients preceding the vectors are the time of action of each vector; if the combination of coefficients is less than one, the remaining time consists of the zero vector.

| Table 1 Alternative vector table |
|----------------------------------|
| Index | Alternative vectors |
|-------|---------------------|
| (1,1) | 1/3VV12,1/3VV10 |
| (2,1) | 1/3VV1,2/3VV1,1/3VV12+1/3VV1 |
| (1,2) | 1/3VV12,1/3VV1,1/3VV12+1/3VV1 |
| (1,2) | 1/3VV12,1/3VV1,1/3VV12+1/3VV1 |
| (3,1) | 2/3VV1,VV1,1/3VV12+2/3VV1 |
| (2,1) | 2/3VV1,1/3VV12+1/3VV1,1/3VV12+2/3VV1 |
| (2,2) | 1/3VV12+1/3VV1,2/3VV12+1/3VV1,1/3VV12+2/3VV1 |
| (1,2) | 2/3VV12,1/3VV12+1/3VV1 |
| (1,3) | 2/3VV12,2/3VV12+1/3VV1 |

By substituting three alternative voltage vectors into Eq. (9) the predicted currents can be obtained. The three predicted currents are then brought into the cost function of Eq. (11), and the voltage vector that minimizes the cost function is selected as the optimal vector. The corresponding PWM signal is then modulated to control the inverter bank.

3.3 PWM modulation

Consider that the average value of the inverter output voltage is only related to the PWM pulse width and is not influenced by the pulse position. According to the action time of each vector, the pulses are moved to the center position, which facilitates the signal generation and reduces the switching frequency. Taking the 1/3VV12 voltage vector as an example, VV12 consists of V45 (1,0,0,1,0,1) acting at 0.7317 and V54 (1,0,1,1,0,0) acting at 0.2697. The on/off times of the switching tubes at the same position of the two vectors are added up and synthesized to 1/3VV12 as (0.333,0.090,0.0,0.333, 0.0,0.090), the PWM waveform is shown in Fig. 5. The overall control block diagram of the system is shown in Fig. 6.
4. Simulation analysis

In this paper, a hybrid voltage vector MPCC is simulated in the Matlab/Simulink platform and compared with the virtual voltage vector MPCC [30]. The parameters are: permanent magnet flux $\psi_f = 0.095$ Wb, rotational inertia $J = 0.0016$ kg·m², d–q axis inductance $L_d = L_q = 20$ mH, stator resistance $R = 2$ Ω, pole pairs $n_p = 5$. The sampling frequency is set to 10 kHz.

Fig. 7 shows the steady-state response of the two control algorithms, which remain stable when the speed reaches a set value. In the phase current waveform, the virtual voltage vector based MPCC contains more harmonics than the hybrid voltage vector based MPCC. Taking the phase A current as an example, the harmonic content of the two is 14.24% and 5.13% respectively. It can be seen that the hybrid voltage vector based MPCC has a better optimization effect on the current harmonics.

Fig. 8 and Fig. 9 give the output torque for two control algorithms with motor speeds of 240r/min and 480r/min to further analyze the steady-state performance. As can be seen from the figures, the torque ripple is smaller in the latter two cases. To quantify the magnitude of torque ripple, the torque values for $N$ sampling cycles were collected and the average value of torque ripple was calculated by the following equation:

$$ T_{\text{rip}} = \sqrt{\frac{1}{N} \sum_{i=1}^{N} (T_i - T_{i,*})^2} \quad (22) $$

where $T_{\text{rip}}$ is the average value of torque ripple; $T_i$ is the current output torque; $T_{i,*}$ is the reference torque; and $N$ is the number of samples. Table 2 calculates the average value of torque ripple based on Eq. (22) for the period from 0.2s to 0.3s. It can be seen from the table that the proposed strategy is effective in reducing torque ripple.

![Fig. 7 Steady state performance. (a) Virtual vector based MPCC; (b) hybrid vector based MPCC.](image)

![Fig. 8 Torque output at 240r/min: (a)Virtual vector based MPCC; (b)Hybrid vector based MPCC.](image)

![Fig. 9 Torque output at 480r/min: (a)Virtual vector based MPCC; (b)Hybrid vector based MPCC.](image)

| Control strategy | Speed (r/min) | Torque ripple (N·m) |
|------------------|---------------|---------------------|
| Virtual MPC      | 240           | 0.3613              |
| Hybrid MPC       | 240           | 0.0864              |
| Virtual MPC      | 480           | 0.3724              |
| Hybrid MPC       | 480           | 0.0996              |

5. Conclusion

This paper proposes an MPCC method with hybrid voltage vectors. The 24 base voltage vectors are synthesized to build virtual voltage vectors to suppress harmonics, and the virtual voltage vectors are further synthesized to generate hybrid voltage vectors to reduce torque ripple and reduce the harmonics caused by the high-amplitude voltage vector at the same time. The reference voltage is calculated from the reference current, and the position of the reference voltage is predetermined in two steps to initially select an alternative set of vectors, reducing the number of model predictions to find the optimum. Simulation results show that MPC with hybrid voltage vectors can effectively reduce motor torque ripple and harmonic currents.
Acknowledgments

This work was funded by the Shandong Province Natural Science Foundation under Grant ZR2019QEE004 and the Fuzhou Science and Technology Plan Funding Project under Grant 2019-G-44.

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