Dual-Vector Model Predictive Current Control of Permanent Magnet Synchronous Motor Drives With the Segment Golden Search Method

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ABSTRACT Dual vector model predictive current control (DVMPCC) has been receiving plenty of concern due to its excellent static and dynamic performance. However, the application time of the selected voltage vector may be a wrong value because the calculation of duty cycle for each voltage vector is independent of the cost function. Here, to overcome the drawbacks of the current calculation methods, the paper introduces a novel approach to calculate the duty cycles for the two voltage vectors in one sampling time by combining the duty cycle calculation into the cost function. In this algorithm, the segment golden search method is used to find the optimal duty cycles for any pair of two voltage vectors. The cost function compares the available pairs of voltage vectors based on the obtained optimal duty cycles of each pair of voltage vectors to select the optimal pair of voltage vectors and duty cycles. In addition, to reduce the complexity of the algorithm, the redundant pairs of voltage vectors are eliminated and the available pairs of the voltage vectors is limit to five. Furthermore, a novel switching pattern is introduced to reduce the average switching frequency. When the inverter adopts the proposed MPCC, the optimal switching time and optimal voltage vector pair are both obtained at the same time, and the result duty cycle is always a reasonable number within the sampling interval. Comparative studies between the proposed method and a recently introduced duty cycle-based MPCC are carried out. The performance of a permanent magnet synchronous motor drive under the proposed MPCC confirms superiority in terms of average switching frequency, better quality currents, and electromagnetic torque ripple.

INDEX TERMS Model predictive control, segment golden search method, permanent magnet synchronous motor, dual-vector MPCC.

I. INTRODUCTION

Model predictive current control (MPCC) is a widely employed control scheme for motor drive systems, photovoltaic power generation systems, and grid-connected renewable energy systems due to its simple concept and fast dynamic response [1]–[5]. MPCC takes advantage of the discrete nature as well as the limited number of available switching states of the power electronics converters [6]. In single-vector MPCC, all possible switching states are examined in each sampling period, and the state that minimizes a predefined cost function is selected and applied to the converter at the next sampling time [6]. The cost function in MPCC is commonly selected as an absolute or squared error of the predicted and the reference currents [7]–[10]. This method is very flexible, allowing the addition of additional control objectives and constraints by modifying the cost function.

In conventional single-vector MPCC, only one voltage vector is applied in the whole control period [11]. Because the switching state is determined directly based on the cost function value, the switching frequency can vary, and the produced currents contain a broad spectrum of harmonics [12]. A fast sampling rate is generally required in single-vector MPCC [13], [14] to achieve the acceptable steady-state performance. Current distortion can be reduced by using a short
Although the two functions have the same value at three x-value points, the minimum value of the two functions is located at different intervals.

Using the SGSM, the minimum value of the two functions is narrowed to the smaller interval.

Segments in the example.

Time for the voltage vector application. However, the short sampling time imposes a high computational burden on the digital signal processor (DSP).

To improve the steady performance of MPCC without increasing sampling frequency, dual-vector MPCC was proposed in [13], [15]–[18]. In each sampling period, two voltage vectors must be selected and applied. In [19], [20], improved MPCC strategies based on the concept of duty cycle control (D-MPCC) were proposed. The idea is based on the fact that it is unnecessary to apply an active voltage vector for the whole sampling period in many operating conditions. In these control strategies, an active voltage vector followed by a zero voltage vector is applied during one sampling period.
The performance of these control strategies is validated to fulfill the control objectives with lower average switching frequencies and an acceptable current quality. However, as discussed in [21], a zero voltage vector results in a reduction of the active power. Therefore, a torque ripple may be produced while a zero voltage vector is applied. To avoid this drawback, active voltage vectors followed by either a zero or an active voltage vector are employed in [6]. In this paper, a fuzzy logic modulator is employed to achieve a low current distortion, power ripple, and average switching frequency.

In general, the two vectors are selected first, and then the optimal duty ratio of each voltage vector is calculated in these methods. Because the two processes are independent of each other, one issue that may arise in dual-MPCC methods is the negative duty cycles of the active voltage vectors for some load conditions [23]. In most cases, these negative duty cycles are corrected to zero, which causes the selected voltage vector to be discarded in the next sampling time [23], [24].
vector and the selected voltage vector. Because the angle is always less than 60 degrees, negative duty cycles can be avoided. Then, the optimal duty cycle is recalculated by using the difference between the voltage error.

In [14], the duty cycle of an active voltage vector is calculated based on the projection of the current error vector onto the active voltage vector, which minimizes the cost function. In the paper, the computational burden is significantly reduced for two main reasons. First, only three candidate voltage vectors are tested in the cost function. Second, the calculation of the duty cycle avoids the use of trigonometric functions. However, the combination of two voltage vectors may not always be the optimal combination.

In [23], negative duty cycles are mitigated based on the fact that the volt-second effect of a positive voltage vector with a negative duty cycle is equivalent to that of a negative voltage vector with a positive duty cycle. In this case, negative voltage vectors with positive duty cycles are selected when there is a negative duty cycle. The average switching frequency is greater than that of the method in [14].

Although the methods in [13], [14], and [23] result in reduced current distortion, nonetheless they are either highly complex and/or less efficient. The reduced efficiencies in [13] and [23], [26], [27] are due to the increased average switching frequency. The performance of the low complexity method in [14], [28] could be improved further.

Here, the dual-vector MPCC with segment golden search method is presented to solve the problem of the calculating duty cycles of the two voltage vectors in the traditional dual-vector MPCC. The major contributions of this paper are as follows. First, a novel method for determining the optimal duty cycles which satisfy the local optimal for the cost function is proposed. Second, the method of eliminating redundant pairs of voltage vectors is proposed to reduce the complexity of the proposed MPCC. Third, a novel switching pattern is introduced to reduce the average switching frequency.
frequency of the inverter. Comparative experimental studies with a recently introduced duty cycle-based dual-vector model predictive controller [19] confirm that while the computational time and average switching frequency is reduced, the proposed MPCC results in improved current quality and torque ripple.

The paper is organized as follows: the segment golden search method is briefly explained in Section II. The cost function in the proposed method is designed by using the golden search method in Section III. The proposed MPCC is discussed in detail in Section IV. Section V gives the experimental results and discussion. The work is concluded in Section VI.

II. SEGMENT GOLDEN SEARCH METHOD

A. INTRODUCTION OF SGSM

Given a continuous real-valued function \( f(x) \) of a single variable, suppose that a minimum exists on the interval as in

\[
\min_{x \in D} f(x) \tag{1}
\]

where \( f: D \to R, D = [a, b] \).

The process of SGSM is as follow:

1) Select two points on the interval, \( a < c_1 < c_2 < b \);
2) Compare \( f(c_1) \) and \( f(c_2) \) to determine whether the root lies on \([a, c_1]\) or \([c_1, b]\);
3) Repeat until the interval is sufficiently small.

In the process used for this method, trigonometry operations, square root, and other complex calculations that complicate the digital implementation are not included. The associated computations are only simple multiplication and addition. The root can be found by evaluating function values at different points on the interval. Hence, SGSM is a simple, robust, and straightforward method that is suitable for digital implementation.

B. SGSM COMPARED TO THE BISECTION METHOD

The basic idea of SGSM is the same as the bisection method, which is carried out by narrowing the interval to approach the local minimizer. In the bisection method, the mid-point \( f((a + b)/2) \) is evaluated to determine which half contains the local minimizer in the bisection method. However, a single point on the interval \([a, b]\) is selected, which is not sufficient to find a minimum. For example, consider the two functions shown in Fig. 1a and Fig. 1b. Each evaluates to the same points at \((1, 4), (2, 1)\) and \((3, 3)\), but Fig. 1a has a minimum on the left-hand subinterval, while Fig. 1b has a minimum on the right-hand subinterval. According to the example, the local minimizer is difficult to find by using the bisection method in this situation.

In SGSM, \( c_1 \) and \( c_2 \) are two points selected on the interval \([1, 3]\). After evaluating the function values at the two points, the minimum value of the two functions is narrowed.
to \([c_1, 3]\) and \([1, c_2]\), respectively, as shown in Fig. 2a and Fig. 2b. Then, replacing the interval \([1, 3]\) with a new interval \([c_1, 3]\) or \([1, c_2]\), two points on the new interval can be selected. Another iteration can yield the narrower interval. The root can be found after several iterations. Hence, SGSM is always useful for finding extreme values in comparison to the bisection method.

C. CHOOSING INTERMEDIATE POINTS

According to the prior example, each corresponding segment is shown in Fig. 3.

In Fig. 3, the more extended segment is the length of the new interval \([c_1, 3]\) or \([1, c_2]\), the shorter segment is the length of either interval \([1, c_1]\) or \([c_2, 3]\) and the sum is the length of the original interval \([1, 3]\).

If the ratio of the more extended segment to the shorter segment is equal to the ratio of the shorter segment to the more extended segment, then this ratio is the golden ratio \(\rho\).

The value of \(\rho\) is 0.618.

With the help of the golden ratio, the two intermediate points can be chosen as follows

\[
\frac{3 - c_1}{3 - 1} = \frac{3 - c_2}{3 - c_1} = 0.618. \quad (2)
\]

D. ALGORITHM OF SGSM

The root is supposed to be bracketed in an interval \([a_0, b_0]\).

Next, the function at two intermediate points \(a_1, b_1\), can be evaluated and yielding a new, smaller bracketing interval, either \([a_0, x]\) or \([x, b_0]\). The variable \(x\) is either \(a_1\) or \(b_1\).

The process continues until the bracketing interval is acceptably small. The selection of \(a_1\) and \(b_1\) should depend on the golden ratio to accelerate the convergence of iteration. The iteration of the golden search method can terminate when the size of the search brackets is sufficiently small, in comparison with some tolerance \(\varepsilon\).

The algorithm of SGSM is provided in SGSM Algorithm.

III. COST FUNCTION OF THE PROPOSED METHOD

To find the pair of voltage vectors and optimal duty cycles, the cost function is designed as follows

\[
J = |i_{sd, \text{ref}} - i_{sd}(n + 1)| + |i_{sq, \text{ref}} - i_{sq}(n + 1)| \quad (3)
\]

where \(i_{sd}(n + 1)\), \(i_{sq}(n + 1)\), and \(i_{sd, \text{ref}}\), \(i_{sq, \text{ref}}\) are the \(d-q\) axis stator current and reference stator current components, respectively. \(i_{sd, \text{ref}}\) is always equal to zero because the magnetic field is positioned to the \(d\)-axis. Two voltage vectors are applied in one sampling period, as shown in Fig 4.

In Fig. 4, \(\Delta T_1\) and \(\Delta T_2\) are the time interval of the application of the first voltage vector, \(v_{\text{fir}}\), and the following voltage vector, \(v_{\text{sec}}\); \(\Delta T\) is the sampling period. The next \(i_{sd}(n+1)\) and \(i_{sq}(n+1)\) in the cost function can be expressed as...
follows

\[
\begin{align*}
    i_{sd} (n + 1) &= i_{sd} (n) + i_{sd1} \left( \frac{\Delta T_1}{\Delta T} \right) + i_{sd2} \left( \frac{\Delta T_2}{\Delta T} \right) \\
    i_{sq} (n + 1) &= i_{sq} (n) + i_{sq1} \left( \frac{\Delta T_1}{\Delta T} \right) + i_{sq2} \left( \frac{\Delta T_2}{\Delta T} \right)
\end{align*}
\]

where \( i_{sd1}(\Delta T_1/\Delta T) \), \( i_{sq1}(\Delta T_1/\Delta T) \), and \( i_{sd2}(\Delta T_2/\Delta T) \), \( i_{sq2}(\Delta T_2/\Delta T) \) is the \( d-q \) axis stator current components of applying \( v_{df} \) for \( \Delta T_1 \) and \( v_{sec} \) for \( \Delta T_2 \), respectively. After substituting (4), (5) into (3), the cost function can be rewritten as follows

\[
J = \left| i_{sd} (n) + i_{sd1} \left( \frac{\Delta T_1}{\Delta T} \right) + i_{sd2} \left( \frac{\Delta T_2}{\Delta T} \right) \right| \\
+ \left| i_{sq} (n) - i_{sq1} \left( \frac{\Delta T_1}{\Delta T} \right) - i_{sq2} \left( \frac{\Delta T_2}{\Delta T} \right) \right|.
\]

IV. PROPOSED METHOD USING THE SEGMENT GOLDEN SEARCH METHOD

A. DISCRETE STATE EQUATIONS OF A PMSM

PMSM state equations in the rotor reference frame are used because sinusoidal quantities appear as constants in steady-state conditions [25], as also used in the proposed method. Moreover, based on the Euler-forward discretization method, discretization of the \( d- \) and \( q \)-axis stator voltage components which will be applied in the next sampling period can be written as

\[
\begin{align*}
    u_{sd} &= \frac{L_d}{\Delta T} i_{sd} (n + 1) - i_{sd} (n) + R_s i_{sd} (n) - \omega_e L_d i_{sq} (n) \\
    u_{sq} &= \frac{L_q}{\Delta T} i_{sq} (n + 1) - i_{sq} (n) + R_s i_{sq} (n) - \omega_e L_q i_{sd} (n) + \omega_e \lambda_{pm}
\end{align*}
\]

where \( i_{sd} \) and \( i_{sq} \) are the \( d-q \) axis stator current components; \( L_d \) and \( L_q \) are the \( d-q \) axis stator inductance components; \( \lambda_{pm} \) is the rotor permanent magnet flux; \( R_s \) is the stator resistance; and \( \omega_e \) is the electromechanical angular speed.

From (7) and (8), the stator current of the \( d- \) and \( q \)-axes at the end of sampling period with the selected voltage vectors should be as follows

\[
\begin{align*}
    i_{sd} (n + 1) &= \frac{\Delta T}{L_d} u_{sd} + i_{sd} (n) - \frac{R_s}{L_d} i_{sd} (n) + \frac{\Delta T}{L_d} \omega_e i_{sq} (n) \\
    i_{sq} (n + 1) &= \frac{\Delta T}{L_q} u_{sq} + i_{sq} (n) - \frac{R_s}{L_q} i_{sq} (n) + \frac{\Delta T}{L_q} \omega_e i_{sd} (n) - \frac{\Delta T}{L_q} \omega_e \lambda_{pm}
\end{align*}
\]
The inductances in both coordinates are equal in surface-mounted PMSM, i.e., \( L_d = L_q \), which leads to the cancellation of either \( L_d \) or \( L_q \) in equations (9) and (10).

**B. STATOR CURRENT EQUATION AFTER APPLYING THE FIRST VOLTAGE VECTOR**

During \( \Delta T_1 \), the first voltage vector \( v_{fir} \) is applied. The stator current \( i_s \) at the moment of \( (n + \Delta T_1/\Delta T) \) is shown in Fig.5.

Supposing that the first voltage vector is the active voltage vector, based on (9) and (10) the \( d \)- and \( q \)-axis stator current components at the end of \( \Delta T_1 \) can be expressed as follows

\[
 i_{sd} \left( n + \frac{\Delta T_1}{\Delta T} \right) = i_{sd} (n) \left( 1 - \frac{R_s \Delta T_1}{L_d} \right) + \Delta T_1 \omega_e i_{sq} (n) + \frac{u_{sd1} \Delta T_1}{L_d} \tag{11}
\]

\[
 i_{sq} \left( n + \frac{\Delta T_1}{\Delta T} \right) = i_{sq} (n) \left( 1 - \frac{R_s \Delta T_1}{L_q} \right) + \Delta T_1 \omega_e i_{sd} (n) - \frac{\Delta T_1}{L_q} \omega_e \lambda_{pm} + \frac{u_{sq1} \Delta T_1}{L_q} \tag{12}
\]

where \( u_{sd1} \) and \( u_{sq1} \) represent the \( d \)-\( q \) axis stator voltage components of the first voltage vector, which is applied in the period of \( \Delta T_1 \).

**C. STATOR CURRENT EQUATION AFTER APPLYING THE SECOND VOLTAGE VECTOR**

During the remainder of the sampling time, the second voltage vector is applied. At the end of this sampling time, the stator current can be shown in Fig.6.

If the second voltage vector is also an active voltage vector, then the stator current at the end of the sampling time can be expressed as follows

\[
 i_{sd} (n + 1) = i_{sd} \left( n + \frac{\Delta T_1}{\Delta T} \right) \left( 1 - \frac{R_s}{L_d} \Delta T_2 \right) + \Delta T_2 \omega_e i_{sq} \left( n + \frac{\Delta T_1}{\Delta T} \right) + \frac{u_{sd2} \Delta T_2}{L_d} \tag{13}
\]

\[
 i_{sq} (n + 1) = i_{sq} \left( n + \frac{\Delta T_1}{\Delta T} \right) \left( 1 - \frac{R_s}{L_q} \Delta T_2 \right) + \Delta T_2 \omega_e i_{sd} \left( n + \frac{\Delta T_1}{\Delta T} \right) - \Delta T_2 \omega_e \lambda_{pm} + \frac{u_{sq2} \Delta T_2}{L_q} \tag{14}
\]

where \( u_{sd2} \) and \( u_{sq2} \) represent the \( d \)-\( q \) axis stator voltage components of the second voltage vector \( v_{sec} \). The working time of this voltage vector \( \Delta T_2 \) is equal to \( (\Delta T - \Delta T_1) \). Accordingly, by replacing \( \Delta T_2 \) with \( (\Delta T - \Delta T_1) \), (13) and (14) can be
rewritten as follows

\[
\begin{align*}

i_{sd}(n + 1) &= i_{sd}(n + \frac{\Delta T_1}{\Delta T}) \left(1 - \frac{R_s}{L_d} (\Delta T - \Delta T_1)\right) \\
&\quad + (\Delta T - \Delta T_1) \omega_e i_{sq} \left(n + \frac{\Delta T_1}{\Delta T}\right) \\
&\quad + u_{sd2} (\Delta T - \Delta T_1) \\

i_{sq}(n + 1) &= i_{sq}(n + \frac{\Delta T_1}{\Delta T}) \left(1 - \frac{R_s}{L_q} (\Delta T - \Delta T_1)\right) \\
&\quad + (\Delta T - \Delta T_1) \omega_e i_{sd} \left(n + \frac{\Delta T_1}{\Delta T}\right) \\

\end{align*}
\]

(15)

\[\frac{(\Delta T - \Delta T_1)}{L_q} \omega_e \lambda_{pm} + \frac{u_{sq2} (\Delta T - \Delta T_1)}{L_q}.
\]

(16)

**D. ANALYSIS OF THE STATOR CURRENT EQUATION AT THE END OF THE SAMPLING TIME**

From (15) and (16), the relationship between the current and the applied voltage vectors is not explicit because there are too many variables.

However, substituting (11), (12) into (15), (16), combining like terms and performing a simplification, (15) and (16) can be rewritten as
Because the motor’s parameters, the sampling values of stator currents and the measured value of the motor’s speed have been obtained at sampling time, two voltage vectors and time of application $\Delta T_1$ are only variables in (17) and (18), as shown at the bottom of this page.

In other words, if the voltage vector pair is known, then the stator current equation is a quadratic function of $\Delta T_1$. After substituting (17), (18) into (3), the cost function is a complicated expression. However, given a pair of voltage vectors, the cost function will also be a quadratic function of $\Delta T_1$.

$$i_{sd}(n+1) = \frac{- (L^2_s \omega_e^2 + R^2_s) i_{sd}(n) + (2R_s \lambda_{pm} - u_{sq1}) \omega_e L_s + R_s u_{sd1}}{L_s^2} \Delta T_1^2 \\
+ \frac{((L^2_s \omega_e^2 + R^2_s) i_{sd}(n) - 2R_s \lambda_{pm} \omega_e) \Delta T + (-\lambda_{pm} \omega_e^2 + u_{sq1} \omega_e) L_s \Delta T + (u_{sd1} - u_{sd2}) L_s - R_s \Delta T \Delta T_1}{L_s^2} \\
+ \frac{(-R_s \Delta T + L_s) i_{id}(n) + (\omega_e i_{sq}(n) L_s + u_{sd2}) \Delta T}{L_s}$$

$$i_{sq}(n+1) = \frac{- (L^2_s \omega_e^2 + R^2_s) i_{sq}(n) + (2R_s \lambda_{pm} - u_{sd1}) \omega_e L_s - R_s (\lambda_{pm} \omega_e - u_{sq1}) L_s}{L_s^2} \Delta T_1^2 \\
+ \frac{((L^2_s \omega_e^2 + R^2_s) i_{sq}(n) - 2R_s \lambda_{pm} \omega_e i_{ld}(n) + L_s u_{sd1} \omega_e + R_s (\lambda_{pm} \omega_e - u_{sq1}) \Delta T + L_s (u_{sq1} - u_{sq2}) \Delta T_1}{L_s^2} \\
+ \frac{(-R_s \Delta T + L_s) i_{sq}(n) + (\omega_e i_{ld}(n) L_s - \lambda_{pm} \omega_e + u_{sq2}) \Delta T}{L_s}.$$ 

**E. ELIMINATING REDUNDANT CANDIDATE VOLTAGE VECTOR PAIRS**

The different combinations of seven different voltage vectors are examined to determine the optimal pair. Forty-nine (49) voltage vector pairs are available to test. If all are examined, then the computational burden will be increased. To address this issue, five different voltage vector pairs are examined in the proposed method based on the voltage vector location in the space [14].

In Fig. 7, it can be seen that when $i_{sq}(n+1)$ is on either the left- or right-hand side of $i_{sq}(n+1)$, the active voltage
vectors that point to the reference of the current vector must be considered as candidate voltage vectors. This is due to the fact that larger current errors, $\varepsilon(n+1)$, would result if active voltage vectors pointing in the opposite direction were applied. Therefore, a criterion based on the relative position of the reference current vector, $i^*_{n+1}$, with respect to the naturally decreasing current vector, $i_{0n}(n+1)$, can be articulated: when $i^*_{n+1}$ is on the left-hand side of $i_{0n}(n+1)$, the active voltage vectors that satisfy the criterion below must be selected.

Depending on the location of the reference stator current and naturally decreasing current vector in Fig. 7, the available voltage vectors should be $v_3$, $v_4$, and $v_5$. Hence, the combinations of three active voltage vectors and one zero voltage vector are, e.g., $(v_3, v_0)$, $(v_4, v_0)$, $(v_5, v_0)$, $(v_3, v_4)$, $(v_4, v_5)$. In these five vector pairs, the first three pairs comprise one active voltage vector along with one zero voltage vector. The remaining vector pairs are comprised of two consecutive active voltage vectors to reduce the number of switching transitions per control period.
FIGURE 29. Performance of the PMSM drive under the proposed MPCC at 30 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

FIGURE 30. Performance of the PMSM drive under DC-DVMPCC at 30 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

FIGURE 31. Performance of the PMSM drive under the proposed MPCC at 0.5 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

FIGURE 32. Performance of the PMSM drive under DC-DVMPCC at 0.5 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

F. OPTIMAL SWITCHING TRANSITIONS BETWEEN TWO CONTROL PERIODS

Because there are two voltage vectors in one sampling time, the switching between the second voltage vector in this control period and the first voltage vector in the next control period should be noted. Based on the types of the two voltage vectors, i.e., an active voltage vector or zero voltage vector, there are two different combinations, e.g., \((v_0 \rightarrow v_4)\), \((v_3 \rightarrow v_4)\), as shown in Fig.8.
FIGURE 33. Performance of the PMSM drive under the proposed MPCC at 0 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

FIGURE 34. Performance of the PMSM drive under DC-DVMPCC at 0 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

SGSM Algorithm

Select two points \(a_1, b_1 \in [a_0, b_0]\)

\[ a_1 = b_0 - \rho (b_0 - a_0), \quad b_1 = (1 + \rho)(b_0 - a_0) - b_0 \]

if \(f(a_1) \neq f(b_1)\) then

\[ x = b_1, \quad x^* \in [a_0, x] \]

else

\[ x = a_1, \quad x^* \in [x, b_0] \]

if \(x - a_0 > \varepsilon\) or \(b_0 - x > \varepsilon\)

Repeat

\[ x^* = \frac{a_1 + x}{2} \] or \[ x^* = \frac{b_1 + x}{2} \]

end

Output: \(x^*\)

FIGURE 35. Dynamic response of the proposed MPCC from speed 60 rad/s to -60 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

In Fig. 8a, the voltage vector pair of an active voltage vector and zero voltage vector is applied in two control periods. The vector pair in the sequence of the active voltage vector and zero voltage vector can help to limit the switching transitions in two control periods to at most two. As mentioned in Section D, the consecutiveness of the active voltage vectors can be used to guarantee one switching transition per control period. However, if two adjacent control periods have the same two active voltage vectors, as shown in Fig. 8b, then the order of the two active voltage vectors in the second control periods should be reversed, which could reduce the number of switching transitions from three to two in the two sampling periods.

In comparison to other sequences of voltage vectors between two control periods, the proposed sequences of consecutive active voltage vectors and one active voltage vector and one zero voltage vector require fewer switching transitions.
Optimal Switching Sequences Algorithm

\[
\begin{align*}
&\text{if } (V_{\text{optimal},1}(n+1) == V_{\text{optimal},1}(n) \text{ and } V_{\text{optimal},2}(n+1) == V_{\text{optimal},2}(n)) \\
&\quad \text{if } V_{\text{optimal},2}(n+1) == 0 \text{ or } 7 \\
&\quad \quad V_{\text{buffer}} = V_{\text{optimal},1}(n+1) \\
&\quad \quad V_{\text{optimal},1}(n+1) = V_{\text{optimal},2}(n+1) \\
&\quad \quad V_{\text{optimal},2}(n+1) = V_{\text{buffer}}; \\
&\text{else} \\
&\quad \text{keep original sequence} \\
&\text{else} \\
&\quad \text{keep original sequence} \\
&\text{end} \\
\end{align*}
\]

Output: \((V_{\text{optimal},1}(n+1), V_{\text{optimal},2}(n+1))\)

---

**FIGURE 36.** Dynamic response of DC-DVMPCC from speed 60 rad/s to -60 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

**FIGURE 37.** Dynamic response of the proposed MPCC from speed 314 rad/s to -314 rad/s. (a) Rotor mechanical speed. (b) Stator phase current. (c) Electromagnetic torque.

### G. OPTIMAL VOLTAGE VECTOR PAIR AND SWITCHING TIME CALCULATION WITH SGSMS

In the proposed method, five pairs of voltage vectors form five cost functions. As previously mentioned, SGSMS can be used to find a minimum of a single-variable function \(f(x)\). Therefore, the optimal value of \(\Delta T_1\), which allows each function minimization to be found with SGSMS. By comparing the minimization of five functions, the optimal voltage vector pair, and corresponding time of application, \(\Delta T_1\), can be obtained.

Figure 9 illustrates the flowchart of the proposed MPCC. In the first step, one of the five candidate pairs of voltage vectors that are chosen by the method already discussed in Section IV(E) is selected. Next, the currents at the \((n+1)\)th sample are calculated using the measured currents and voltages. Then, the cost function of the pair of voltage vector is expressed. By using the golden search method, the optimal time that minimizes the cost function is found. After five iterations, both the optimal pair of voltage vectors and optimal switching time are obtained. Additional details of the golden search method are explained in the following sections.

As shown in Fig. 10, the calculation used to obtain the optimal duration of the dual-voltage vectors is based on the algorithm given in Section II(D). First, the golden segment is calculated, and then, the two endpoints of the interval are selected. Second, after checking the sign of the error between the two endpoints, the updated interval and two endpoints in the new interval are determined. Finally, if the convergence of the algorithm is checked, the mean of the two endpoints is the optimal duration of the first voltage vector.

### H. STABILITY OF THE PROPOSED MPCC

The time derivative of the stator current in the \(d-q\) frame can be written as

\[
\frac{di_{dq}}{dt} = \frac{1}{L} u_{dq} - (R_s + \omega_L L_s) i_{dq}
\]  

(19)

Based on Lyapunov’s stability theory, a control Lyapunov function is defined to analyze the closed-loop stability of the proposed control strategy. If the Lyapunov function candidate \((V(x))\) satisfies the following criteria, then the global asymptotic stability of the proposed control strategy system
The Lyapunov function is defined as a quadratic function of the current error, \( \tilde{i} \), as follows:

\[
V(\tilde{i}) = \frac{1}{2} K_i \tilde{i}^2
\]

where \( K_i \) is a constant positive gain. From the definition, the selected Lyapunov function in (21) satisfies the first three criteria given in (20). If the time derivative of the Lyapunov function is negative, then global asymptotic stability can be achieved. Using the chain rule and (19), the time derivative of the Lyapunov function is calculated as

\[
\frac{dV}{dt} = \frac{\partial V}{\partial \tilde{i}} \frac{d\tilde{i}}{dt} = K_i \left( \frac{1}{L} u^*_{dq} - \frac{1}{L} ((R_s + \omega_e L_s) i_{dq}) - \frac{di_{dq} \cdot ref}{dt} \right)
\]

where

\[
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\]

It is supposed that the control input reference \( u^*_{dq} \) is in the form of (25); as a result, the time derivative of the Lyapunov function in (24) can be simplified to (26)

\[
u^*_{dq} = \left( (R_s + \omega_e L_s) i_{dq} + L \frac{di_{dq \cdot ref}}{dt} \right) - L (i_{dq} - i_{dq \cdot ref})
\]

Based on (26), for \( \tilde{i} \neq 0 \), the time derivative of the Lyapunov function is negative. This means that for \( t > 0 \), \( V(t) \leq V(0) \); therefore, \( i, \tilde{i}, \) and \( u^*_{dq} \) are bounded. Consequently, \( \dot{V}(\tilde{i}) \to 0 \) as \( t \to \infty \); therefore, \( \tilde{i} \to 0 \).

As a result, by selecting (21) as the Lyapunov function, the closed-loop control system is asymptotically stable, provided that the control input reference given in (27) and shown at the bottom of this page is valid. In the discrete time domain, (27) can be approximated as (28).

The space vector modulation theory of a two-level converter, the reference of the voltage vector can be expressed as a convex combination of the realizable inputs [6]. It has been shown that at least one of input vectors is stabilizing [30]. Based on this fact, the stability of the proposed method can be guaranteed by selecting the voltage vector pair that can satisfy (28). If there is no voltage vector pair to satisfy
the criteria, then only one voltage vector is selected.

\[
\frac{dV}{dt} = \frac{\partial V}{\partial i} \frac{\partial i}{\partial t} = K_i \left( \frac{1}{L} u_{\text{ref}1, dq} + (1 - d) u_{\text{ref}2, dq} \right) - \frac{di_{\text{ref}, dq}}{dt} \leq 0 \quad (27)
\]

\[
\frac{dV}{dt} \approx K_i \left( i(k+1) - i_{\text{ref}, dq}(k) \right) \times \left( \frac{1}{L} \left( u_{\text{ref}1, dq}(k) + (1 - d) u_{\text{ref}2, dq}(k) \right) \right) - \frac{1}{L} \left( (R_s + \omega_c L_s) i_{\text{ref}, dq}(k) \right) - \frac{i_{\text{ref}, dq}(k+1) - i_{\text{ref}, dq}(k)}{T_s} \leq 0 \quad (28)
\]

**V. EXPERIMENTAL AND SIMULATION RESULTS**

**A. SIMULATION AND EXPERIMENTAL SETUP**

A block diagram of the proposed MPCC is shown in Fig.11. As the first step, the measured currents, voltages, and rotor position at the \(n\)th sample are fed into the current prediction to obtain the currents in the \((n+1)\)th sample. Next, the redundant active voltage vectors are eliminated by the proposed method. Then, the error of the velocity, \(d\)-axis current and candidate pairs of the voltage vectors are fed to the modulator of the cost function minimization and golden search method to determine the duty cycles of the two voltage vectors in the next sampling period. To ensure stability, if no voltage vector pair can satisfy the criteria in Section IV(H), then only one voltage vector is selected. Eventually, the voltage vector pair that minimizes the cost function is selected, and the corresponding switching state is applied.

Based on the control diagram in Fig. 11, the simulation model of the PMSM is established by using MATLAB/Simulink to demonstrate the effectiveness of the proposed method. The compared method is the two-voltage vector model predictive control with a duty-cycle method, referred to as DC-DVMPC. In the simulation, the sampling time is 10 \(\mu\)s. The parameters of the machine and controller are listed in Table 1. The models are simulated to test the two methods based on the same situations.

Various experimental tests are also carried out using a dSPACE setup, as shown in Fig.12. The setup consists of a dSPACE controller, a PMSM, and a feeding two-level inverter. During the tests, all variables are displayed on Controldesk via an onboard AD converter, while the sampling time is 100 \(\mu\)s. The voltage is measured by the voltage sensor.
Moreover, the performance of the proposed method is compared with DC-DVMPC. The control strategies are implemented in the dSPACE with a time step of 100 µs. The machine and control parameters are also listed in Table 1. Two methods will be based on these. The switching frequency of the PWM is set to 10 kHz.

### B. SIMULATION RESULTS

In the simulation, the speed references are set at 70 rad/s, 60 rad/s, 50 rad/s and 30 rad/s to compare the steady state of the proposed MPCC and DC-DVMPPCC at different speeds.

In Fig. 13(a) and (c), when the PMSM runs under the proposed MPCC at 70 rad/s, the output speed and torque contain little error and ripple, respectively. In Fig. 13(b), there is no obvious distortion in the stator current of the PMSM.

In Fig. 14(a) and (c), the output speed and torque with DC-DVMPPCC at 70 rad/s have significant error and ripple, respectively. The stator current in Fig. 14(b) depicts that the distortion is more than that of Fig. 13(b).

The waveforms of the speed, stator current and torque with these two algorithms at different speed reference are provided in Fig. 15 through Fig. 20. Both approaches can trace the reference value correctly. However, the speed error, torque ripple and distortion of the stator current in the proposed MPCC are less than those of DC-DVMPPCC.

The dynamic performances of the two methods are tested by varying the speed reference from 314 rad/s to -314 rad/s at the time of 1s. The dynamic performance of the proposed MPCC is shown in Fig. 21. The response time from 314 rad/s to -314 rad/s shown in Fig. 21(a) is approximately 0.32 s. In addition, the stator current of PMSM has a good sinusoidal output wave, which is shown in Fig. 21(b). In Fig. 21(c), the torque ripple is not big.

In Fig. 22(a), the output speed error with DC-DVMPPCC is close to that of the proposed MPCC. However, the response time of the speed change is approximately 0.35 s, which is greater than 0.32 s in Fig. 21(a). The stator current in Fig. 22(b) has the more obvious harmonics in comparison to Fig. 21(b). The waveforms of the torque in Fig. 22(c) has a bigger ripple in comparison to Fig. 21(c).

The dynamic performance of DC-DVMPPCC and the proposed MPCC method are investigated in the second
two scenarios. Depending on these simulation results, the dynamic performance with the proposed MPCC has advantages in the reduced response time, increased sinusoidal stator current, and lesser torque ripple over DC-DVMPCC due to the optimal pair of voltage vectors and accurate duration of the applied two voltage vectors in the proposed MPCC.

C. EXPERIMENTAL RESULTS

The performance of DC-DVMPCC and the proposed MPCC method is investigated in two scenarios.

In the first scenario, the motor is operated at 70 rad/s under a load torque of 0.25 Nm. Fig. 23 depicts the performance of the proposed MPCC. By comparing the stator currents illustrated in Fig. 23(d) and Fig. 24(d), the proposed MPCC method has a lower stator current THD. In addition, the torque ripple is lower, as seen in Fig. 23(c) and Fig. 24(c). The calculation of the torque ripple is as follows

$$T_{\text{rip}} = \frac{\sum_{i=1}^{n} |T_e(i) - T_L|}{n}.$$  \hfill (29)

The torque ripples of the proposed MPCC and DC-DVMPCC are 0.023 Nm and 0.025 Nm, respectively. The measured average switching frequency of the proposed MPCC and DC-DVMPCC are 3.73 kHz and 4.43 kHz, respectively. The turnaround times for the proposed method and DC-DVMPCC are 13.4 µs and 15.5 µs.

The performances with the proposed MPCC at the speed of 60 rad/s, 50 rad/s, 30 rad/s, 0.5 rad/s, and 0 rad/s are shown in Fig. 25, Fig. 27, Fig. 29, Fig. 31, and Fig. 33, respectively. The waveforms of DC-DVMPCC at the same speeds as the proposed MPCC are shown in Fig. 26, Fig. 28, Fig. 30, Fig. 32, and Fig. 34, respectively. The associated experimental results are summarized in Table 2.

When the motor run at 0.5 rad/s under the proposed method and DC-DVMPCC in Fig. 33 and Fig. 34, the speed error is approximately 0.015 rad/s and 0.034 rad/s, respectively. The current THD of the proposed MPCC is approximately 156% in comparison to the value of 180% determined for DC-DVMPCC. The torque ripples of the proposed MPCC and DC-DVMPCC are 0.16 Nm and 0.18 Nm, respectively.

For the PMSM used in this paper, the current THD is quite high. This result is mainly relevant to the machine itself. Both methods yielded similar results after being tested on the same motor, which is independent of the control methods. The harmonic spectrum and torque ripple reduction are evident at normal and high speeds in the proposed method.

These results confirm that the proposed MPCC is useful in obtaining a better performance than DC-DVMPCC over a wide speed range. The turnaround time of the proposed MPCC is approximately 13.5 µs at the speed of 70 rad/s, which is lower than the value of 15 µs in DC-DVMPCC. The effectiveness of the proposed MPCC in reducing complexity is also confirmed.
The dynamic response of the two methods is evaluated as the second scenario. As depicted in Figs. 35 and 36, step changes in reference speed, from 60 to -60 rad/s, and a load torque of 0.25 Nm are applied. In Fig. 35(a), the dynamic response time of the proposed MPCC is 0.20 s. This features a faster dynamic response in comparison to the response time of DC-DVMPC, which is 0.22 s, as shown in Fig. 36(a). The torque ripples of the proposed MPCC and DC-DVMPC at 30 rad/s are 0.046 Nm and 0.051 Nm, respectively. The rotor speed increased from 30 rad/s to 60 rad/s, and the torque ripple of the proposed MPCC and DC-DVMPC are 0.038 Nm and 0.042 Nm.

The performance of the proposed MPCC and DC-DVMPC under the reference speed change from 314 rad/s to -314 rad/s is shown in Figs. 37 and 38, respectively. The response time of the proposed MPCC is 0.85 s. In comparison to the 0.95 s for DC-DVMPC, these figures also show that the proposed MPCC has a faster dynamic response.

The average switching frequency decreases with the increase of the motor speed for both methods. However, the proposed MPCC maintains a significantly smaller switching frequency as the speed is varied. In all case studies, the proposed MPCC results in a significantly reduced THD and complexity of the control algorithm.

The summarized results in Table 2 confirm that the proposed MPCC is effective in improving the current quality at the reduced average switching frequency and computational time compared with DC-DVMPC.

The third case study compares the performance of the two control methods in the presence of motor parameter uncertainties. Accordingly, a 50% increase in the stator resistance and a 30% decrease in the stator inductance are introduced in the proposed MPCC and DC-DVMPC. The motor is operated at 70 rad/s, 60 rad/s, 50 rad/s, and 30 rad/s under a load torque of 0.25 Nm. Figs. 39, 41, 43, and 45 depict the
performance of the proposed MPCC at speeds of 70 rad/s, 60 rad/s, 50 rad/s, and 30 rad/s, respectively. The performance of DC-DVMPC at speeds of 70 rad/s, 60 rad/s, 50 rad/s, and 30 rad/s is illustrated in Figs. 40, 42, 44 and 46, respectively. With the erroneous assumption of motor parameters at both speeds of operation, the stator current THD increases under both control methods. However, the increase is smaller under the proposed MPCC. The same applies to changes in the torque ripple (see Table 3).

VI. CONCLUSION
In this paper, a dual-vector MPCC with SGS for PMSM is presented to overcome the shortcoming of the duty cycle calculation in conventional dual-vector MPCC. In comparison with the DC-DVMPC, it mainly has three different aspects. First, a novel method to calculate the optimal durations of the two voltage vectors was proposed to guarantee the obtained value is local optimal. Second, a method of elimination the reluctant voltage vectors was introduced to reduce the complexity of the proposed MPCC. Third, a new switching sequence was proposed in proposed MPCC to reduce the average switching frequency of the inverter. The simulation and experimental results demonstrate that the current THD of the proposed MPCC declines by 3%–5% in normal operation compared with the DC-DVMPC. Moreover, the execution time is also decreased remarkably. In addition, the average switching frequency of the inverter is less than the DC-DVMPC. Moreover, the dynamic response of the proposed MPCC is also faster than the DC-DVMPC. However, in comparison with single vector FCS-MPC, the proposed method is more time consuming than the single vector FCS-MPC.

In addition, we are developing the proposed algorithm for reducing the complexity further, and the research achievements will be presented in the following paper.

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