A modified duty-modulated predictive current control for permanent magnet synchronous motor drive

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Abstract
Predictive current control (PCC) is a promising technique for adjustable speed drives due to its intuitive features such as quick response, easy inclusion of non-linearities and constraints as well as multi-variable operation. The conventional predictive current control (C-PCC) employs a cost function to choose the optimum vector, which reduces the error between the reference and the predicted current. Even though the C-PCC can achieve excellent dynamic performance with a single active voltage, the steady-state performance is affected with larger torque and flux ripples. When two voltage vectors are applied in a control interval, the steady-state performance is improved. This paper proposes a modified duty-based PCC for permanent magnet synchronous motor (PMSM) drives, employing an active and null vector in control interval, where the duration for which the optimal active vector is applied is based upon a simple error based calculation. The proposed duty calculation requires a simple constant and eliminates the need for current slope calculation, which results in parameter insensitive operation. The proposed method is compared with C-PCC and recently published works based on duty calculation. The simulation and hardware experiments conducted on a PMSM drive confirm the effectiveness of the proposed method with reduced torque and flux ripples.

1 | INTRODUCTION

Permanent magnet synchronous motor (PMSM) is an attractive solution for the industrial drive applications, Electric and Hybrid Electric vehicles, Machine tools, Servo drives and domestic appliances, like washing machines, vacuum cleaners and so on where the torque to weight ratio, power density, dynamic performance as well as efficiency play a vital role [1,2]. PMSM possess inherent magnets to produce the rotor flux, unlike the Induction motor with a wound rotor. This makes the machine more compact and the rotor losses are also considerably reduced.

The applications demanding high dynamic performance usually employ field-oriented control (FOC) or direct torque control (DTC) techniques [3]. Even though FOC offers excellent steady-state performance, it requires modulators, complex rotor transformations as well as Proportional-Integral (PI) tuning. On the other hand, the DTC technique gives good dynamic performance without any co-ordinate transformations, modulators and tuning [4,5]. A pre-determined switching table based on the errors in flux and torque aids to achieve higher torque dynamics in DTC. The drawbacks of DTC are the higher torque and flux ripples as well as the difficulty to maintain the torque and flux within a definite range at very low speeds [6].

In recent years, with the introduction of fast and powerful processors, the computationally intensive model predictive control (MPC) techniques are drawing more attention [7]. MPC is conceptually similar to DTC and FOC techniques. MPC is a potential scheme for electric drives due to its intuitive features, like high dynamic control, easy inclusion of non-linearities and constraints, multi-variable operation and so on.

The MPC schemes as in Ref. [8] can be mainly classified as Continuous Control Set Model Predictive Control and Finite Control Set Model Predictive Control (FCS-MPC). The concept of FCS-MPC is to estimate the unknown variables, predict the future behaviour of the system and then find an optimal value of the actuating variable to reduce the cost function of the optimization problem. Thus, in every sampling period, the predicted value is compared with the reference value and the error is minimized by applying the best possible actuating variable.
Predictive torque control (PTC) and predictive current control (PCC) techniques are the most widely used model predictive control schemes. In the PTC technique, accurate estimators and observers are used to obtain the actual torque and flux. The voltage vector that decreases the torque and flux undulations is chosen as the optimal state for the two-level inverter. The priority of the specific variable in a multi-objective cost function is included in terms of the weighting factor [9]. Thus, to obtain the actuating vector, a cost function carrying the constraints, variables, specific targets and the associated weighting factors is optimized. In every control interval, the prediction and actuation take a considerable amount of time, leading to the limitation of sample frequency and thereby, degradation in the flux and torque response. Another issue involved in this technique is the tuning of the weighting factor.

In PCC, the current vector is considered as the control variable in the cost function [10]. This reduces the burden of estimation of the stator flux and torque. Moreover, the weighting factor tuning required in PTC can be avoided in the PCC scheme. The control is also easy to implement and more feasible. In conventional predictive current control (C-PCC), the control vector is applied for an entire sample time and hence the torque and flux response are degraded as the error is not controlled accurately to a minimum value. In Refs. [11–13], one active and one null vector is applied in a control interval, to enhance the steady-state performance in DTC and PTC schemes, as the null vector causes only a small variation in current. The same principle is applied to PCC to improve the steady-state response. Three vectors are applied in one control period in Ref. [14], based on an integrated switching table. The purpose of the null vector employed at the end of each sample is to obtain a reduction in flux and torque ripple and this is realized by varying the duty ratios of the initial active vectors.

However, this method is complex and requires more number of switching transitions. In Ref. [15], the torque ripple is dependent on the weighting factor and the duration of voltage vectors are calculated with the optimized weighting factor. In this method, the priority is to calculate the weighting factor and then the prediction of torque and flux. The computations are thus increased in the technique.

When the duty calculation is cascaded with the vector selection and optimization, the obtained machine performance is poor. Thus, the low-speed performance of the drive is improved by incorporating the duration of active voltage vector in the cost function in Ref. [16]. The stator flux slope and torque slope are used in Ref. [17] for the duty determination. However, the control becomes infeasible if the calculation time exceeds the control period. The control performance is also deteriorated with the parameter variation as the algorithm depends immensely on the machine parameters. The duty cycle is calculated based on the RMS torque ripple minimization principle, and the duty obtained is insensitive to parameter variations in Ref. [18]. The method is simple but requires additional tuning of two weighting factors.

In Ref. [19], a modulator based on the fuzzy logic is utilized to obtain the duty cycles of active and null voltage vectors. A Luenberger observer is deployed to obtain the state variables. In Ref. [20], two similar duty control methods are proposed by the authors. In the first method, the duty is calculated based on the current slopes. The second method is based on the deadbeat principle where the reference voltage vector is first obtained. The vector closer to the reference can be considered as the optimum vector. The current locus is predicted in Ref. [21], and a new reference frame is established. The zero vector is chosen based on the error between the reference frame and the origin of the developed frame. The optimum voltage vector is also obtained based on the current locus, unlike C-PCC where the cost function is used. This aids to reduce the computational burden.

A switching table is introduced in Ref. [22] to reduce the computational burden while selecting the optimum active vector and duty optimization is performed to reduce the ripples in steady-state responses. In Ref. [23], the parameter robustness of the PCC scheme is improved using a model based on incremental prediction. The performance given by the method against inductance mismatch disturbance is analysed and an inductance observer is used to reduce the parameter dependency. Zhang et al. [24] utilize a multi-vector based generalized MPC with voltage error cost function to reduce the control complexity and computational burden. A three vector-based duty method is adopted in Ref. [25], where an extended control set is developed to improve the torque and flux performance. An inverted triangular matrix is used for storing the duty information. Unlike C-PCC, this method uses voltage terms in the cost function.

In Refs. [26,27], a multi-vector based approach is considered for a two-level inverter, where the optimal vector, as well as duty, is obtained from the deadbeat control based on space vector modulation but the complexity is increased with more transformations. The PCC method adopted in Ref. [28–30] employs three vectors in a control interval. A second-order super-twisting algorithm based sliding mode observer, Luenberger observer as well as a disturbance feedforward compensation based on proportional integral observer are employed to estimate the lump disturbance and to reduce the computational burden in the aforementioned methods. Although these methods ensure robustness against parameter mismatch, the overall complexity is enlarged. In Ref. [31], the current tracking method is employed instead of the cost function for obtaining a quick response similar to Ref. [21].

The optimal voltage vector and duty is obtained using the Lagrange multipliers in Ref. [32], while in Ref. [33], vector modulation is used to make the voltage vector application more optimized and adaptive. The PCC duty methods are applied for five-phase machines in Refs. [34–36], where either virtual voltage vectors or sector determination is used to obtain good steady-state performance. However, all these methods have the limitation of the large computational burden.

A dual-vector based duty approach is used in Ref. [37] to improve the steady-state performance, where, the duty is obtained by projecting the current error onto the optimum voltage vector which minimizes the cost function. Moreover, the redundant active voltage vectors are eliminated to significantly reduce the computational burden. However, the method
is implemented in the rotor reference frame which increases the complexity of the control. In Ref. [38], modulated PCC is employed with virtual vectors to enhance performance. The duty can be determined by the optimization of cost function and non-linearity compensation is used to reduce the error in tracking input current. The duty cycle methods in Refs. [39,40] are based on the calculation of the current slopes. In Ref. [40], the calculation of quadrature-current slopes for each active voltage vector is required and the intersection points are obtained. The current errors are also calculated at these intersection points. The optimal vector which minimizes the error is chosen. This duty calculation method is complex and demands higher computational time.

The major drawbacks of PCC are the large ripples in flux and torque incurred due to the application of one active voltage vector in a control interval. The available literature indicates that the application of two (or more) voltage vectors can enhance the steady-state responses. However, the simplicity and robustness of the C-PCC are negated in the available duty methods as it involves complex calculations and relies mainly on the accuracy of the parameters used.

This paper proposes a modified PCC for a two-level voltage source inverter (VSI) based PMSM drive to improve the steady-state performance with two voltage vectors (one active and null) in a control interval. In the proposed scheme, the authors aim to retain the simplicity of the C-PCC method while addressing the demerits of large torque and flux ripples. The proposed duty ratio is obtained as the ratio of the error between the reference and predicted current values and a constant. The control of quadrature axis current error would minimize the torque variations and the direct-axis current error control aids to reduce the flux variations. In the proposed method, the calculations involving current slopes are eliminated, the steady-state performance is improved and the computational burden is reduced. Moreover, the proposed method establishes a parameter insensitive duty calculation. A comparison of the proposed method is made with C-PCC [20,37]. The results indicate that better steady-state performance is achieved in the proposed method. Moreover, dynamic performance is unaltered. The paper has the following organization: the dynamic model of the surface-mounted PMSM (SPMSM) is offered in Section 2. Section 3 describes the C-PCC while Section 4 presents the proposed PCC. Section 5 and 6 comprises of simulation and experimental results. Section 7 presents the conclusions.

2 | PMSM DYNAMIC MODELLING

In order to avoid the intricate rotor co-ordinate transformations, SPMSM in the stationary reference frame is considered in this paper. The mathematical model can be obtained using the following equations.

\[ u_s = i_s R_s + \frac{d\psi_s}{dt} \]  

(1)

\[ \psi_s = L_s i_s + \psi_r \]  

(2)

Substituting Equation (2) in Equation (1) and re-arranging gives,

\[ \frac{di_s}{dt} = \frac{1}{L_s} (u_s - R_s i_s - j\omega_s \psi_s) \]  

(3)

where, \( \psi_r = \psi_f e^{j\theta} \)

The electromagnetic torque can be obtained as:

\[ T_e = 3 \frac{p}{2} \text{imag}(\psi_s \times i_s) \]  

(4)

\[ T_e = 3 \frac{p}{2} i_q \psi_f \]  

(5)

where,

- \( R_s \)—Stator resistance (ohm)
- \( L_s \)—Synchronous inductance (Henry)
- \( \psi_f \)—Flux produced by the permanent magnet (Weber)
- \( \psi_e \)—Stator flux (Weber)
- \( \overline{\psi}_e \)—Complex conjugate of Stator flux (Weber)
- \( u_s \)—Stator voltage (Volt)
- \( i_s \)—Stator current (Ampere)
- \( p \)—No. of pole pairs
- \( i_q \)—Quadrature axis current (Ampere)
- \( \omega_s \)—Speed (electrical rad/s)

The mechanical equation is

\[ J \frac{d\omega_m}{dt} = T_e - T_l - B\omega_m \]  

(6)

where,

- \( J \)—Moment of inertia (kg·m²)
- \( T_l \)—Load torque (Nm)
- \( T_e \)—Developed torque (Nm)
- \( \omega_m \)—Mechanical speed (rad/s)
- \( B \)—Frictional co-efficient

3 | CONVENTIONAL PCC

PCC is evolving as a powerful scheme for controlling the PMSM drives, due to its inherent simplicity, the capability to include non-linearities as well as quick dynamic response [41]. In the PCC method, stator current for the next instant is predicted and the optimal switching state is obtained by minimizing a cost function. The performance of the method is based on the proper design of the cost function. The cost function, with a set of non-linear constraints, must be minimized to determine the optimal vector required to be applied in the next sampling period. The block diagram of PCC is
given in Figure 1. Unlike MPTC, the PCC technique does not require a weighting factor.

In a two-level VSI based PCC, eight basic voltage vectors as in Figure 2 are used. The torque varies linearly with stator quadrature axis current, $i_{qs}$ according to the Equation (3).

Any deviations in $i_{qs}$ will cause a similar change in the torque developed. Conversely, to reduce the ripples in torque, variations in $i_{qs}$ must be reduced. Therefore, the control strategy is to give the direct-axis current reference, $i_{ds}^*$ as zero. The output of the speed PI controller is given as the quadrature axis current reference, $i_{qs}^*$ as the speed dynamics is proportional to the motor torque which further depends on the current $i_{qs}$. The steps involved in C-PCC are as follows: (i) Prediction of current in stator, (ii) Cost Function estimation and (iii) One-step delay compensation.

### 3.1 | Prediction of the current in the stator

The stator currents, DC link voltage and speed of the PMSM at the $k^{th}$ instant are fed to the analogue to digital converter to digitalize the signals. The predicted current is obtained from the following equations. From Equation (3),

$$\frac{di_s}{dt} = \frac{1}{L_s}(u_r - R_s i_s - j\omega_s \psi_s) \quad (7)$$

Applying the first-order Euler’s discretization to Equation (7),

$$i_s(k+1) = i_s(k) + \frac{T_s}{L_s}(U_r(k) - i_s(k)R_s - j\omega_s \psi_s e^{j\theta}) \quad (8)$$

### 3.2 | Cost function estimation

The cost function ($G$), defines the deviation of predicted current from the reference value of current. This error should be minimized to get optimum actuating vectors. PCC cost function is given by,

$$G = |i_{ref} - i_s(k+1)|^2 \quad (9)$$

### 3.3 | One-step delay compensation

The real-time implementation of the control using a digital processor introduces a single-step delay between the voltage command and the actual voltage value [42].

This deteriorates the performance of the PCC. The delay can be compensated by predicting the control vectors at the $(k+2)^{th}$ instant instead of $(k+1)^{th}$ instant. The predicted current at the $(k+1)^{th}$ instant can be obtained from Equation (8). The final values of the currents are obtained by replacing $k$ by $(k+1)$ and $(k+1)$ by $(k+2)$ respectively. The stator current at the instant $(k+2)$ is:

$$i_s(k+2) = i_s(k+1) + \frac{T_s}{L_s}(U_r(k+1) - i_s(k+1)R_s - j\omega_s \psi_s e^{j\theta}) \quad (10)$$

The cost function can be modified as:

$$G = |i_{ref} - i_s(k+2)|^2 \quad (11)$$

The voltage vector which decreases the deviation in current error is opted as the optimum active voltage vector by the controller.

In PCC [20,37], two vectors are applied in one sample time, to control the torque variations in the steady-state response. The active voltage vector is chosen as in C-PCC, as per the Equation (11). The time ($t_{cep}$) required for the active and null vector is determined based on the difference between reference and predicted current. The time duration for active voltage vector application ($t_{cep}$) in Ref. [20] can be expressed as per the equation given below.
where, \( \mathbf{\cdot} \) gives the dot product of the two vectors, \( i_a(k + 1) \) and \( i_{ref} \) are the measured and reference value of stator currents.

\( m_0 \) and \( m_1 \) are the current slopes due to the null voltage vector and active voltage vector respectively. The duty ratio obtained in this method deals with the current slopes which are machine parameter dependent.

In Ref. [37], the duration of the active voltage vector (\( t_{ap} \)) is expressed as,

\[
 t_{ap} = \frac{c_{0}(k + 1) \cdot T_i U_i(k)}{T_i U_i(k) \cdot T_i U_i(k)}
\]

where, \( c_{0}(k+1) \) is the error between reference current and the current due to zero voltage vector. In this method, the redundant voltage vectors that are pointing in the opposite direction of the reference vectors are eliminated so that the total active voltage vectors for optimal vector selection is reduced to three. However, this method is established in the rotor reference frame which increases the control complexity.

### 4 | PROPOSED PCC

In C-PCC, the voltage vectors that reduces the error in the current according to the cost function \( G \) (Equation (11)) will be chosen for one control interval. In a two-level VSI, only eight voltage vectors are offered and the best vector which minimizes the error is chosen for actuation. But, this results in unwanted ripples in the steady-state response. The necessity to apply two voltage vectors (one active and one null) is hence to improve the steady-state performance. This method utilizes two voltage vectors during the control interval. The block diagram of the proposed method is shown in Figure 3.

In Equation (8), substituting \( U_i(k) = 0 \), we get,

\[
i_{s o}(k + 1) = i_e(k) + \frac{T_i}{L_s} (- i_e(k)R_s - jw_f w_e e^o)
\]

Thus, from Equation (8),

\[
i_e(k + 1) = i_e(k) - i_{s o}(k + 1)
\]

\[
e_e(k + 1) = e_e(k + 1) - i_{s o}(k + 1)
\]

where, \( e(k + 1) \) and \( e_o(k + 1) \) are the errors with respect to the reference current \( i_e(k+1) \).

Substituting value of \( i_e(k+1) \) from Equation (18) in Equation (19) gives,

\[
e_e(k + 1) = e(k + 1) + \frac{T_i}{L_s} (U_i(k))
\]

The entire sampling time is comprised of two intervals, one for active and other for null vector application. The Equation (20) (in Figure 4) proves that error is reduced when the null voltage vector is applied.

The proposed PCC method uses a deadbeat control of current components under the stationary reference frame for calculating the vector duration. The steps involved in the

**Figure 3** Block diagram of the proposed PCC method. PCC, predictive current control; PMSM, permanent magnet synchronous motor; VSI, voltage source inverter
method are shown in Figure 5. The one-step delay compensation is considered to recompense the performance of the control during the digital implementation.

According to the principle of deadbeat control,

\[ i_s(k + 2) = i_s(k + 1) + m_1 t_{op} + m_0 (T_s - t_{op}) \]  \hspace{1cm} (21)

where, \( m_o \) and \( m_1 \) are the current slopes as per Figure 6. \( m_o \) is the current slope due to the null vector and \( m_1 \) is the current slope due to the active vector, \( t_{op} \) is the active voltage vector duration and \( T_s \) is the sampling time (100 \( \mu s \)).

The current slopes can be obtained from the equations given below:

\[ m_1 = \frac{di_s}{dt}_{\text{active}} = \frac{u_s - R_s i_s - j \omega_s \psi_r}{L_s} \]  \hspace{1cm} (22)

\[ m_o = \frac{di_s}{dt}_{\text{null}} = \frac{-R_s i_s - j \omega_s \psi_r}{L_s} \]  \hspace{1cm} (23)

In any sample time,

\[ m_1 - m_o = \frac{|U_{opt}|}{L_s} \]  \hspace{1cm} (24)

Substituting Equation (21) in the cost function,

\[ G = |i_{ref} - i_s(k + 1) - (m_1 - m_0)t_{op} - m_0 T_s|^2 \]  \hspace{1cm} (25)

Applying the principle of minimization, \( \frac{\partial G}{\partial t} = 0 \) implies,

\[ d = \frac{|i_{ref} - i_s(k + 2)|}{(m_1 - m_o)T_s} - \frac{m_o}{m_1 - m_o} \]  \hspace{1cm} (26)

where, \( d = \frac{t_{op}}{T_s} \), the duty ratio.

From Equation (24), \((m_1 - m_0)T_s\) can be considered as a constant, \( C \).

Substituting values from Equations (22) and (23) we get,

\[ m_o = \frac{-R_s i_s - e(k)}{U_s(k) (m_1 - m_0) L_s} \]  \hspace{1cm} (27)

Discretizing Equation (1) gives,

\[ U_s(k) = R_s i_s(k) + \frac{L_s |i_{ref} - i_s(k)|}{T_s} + e(k) \]  \hspace{1cm} (28)

where, the back emf, \( e(k) = j \omega_s \psi_r \).

From Equation (28),
Figure 6 Stator current, $i_s$, in a sample time, $T_s$

$$R_s i_s(k) + e(k) = U_s(k) - \frac{L_s (i_{ref} - i_s(k))}{T_s}$$ (29)

Substituting for the numerator of Equation (27) from Equation (29),

$$\frac{m_o}{m_1 - m_o} = \frac{L_s (U_s(k)T_s - L_s (i_{ref} - i_s(k)))}{U_s(k)}$$

$$= -\left( L_s - \frac{L_s^2 (i_{ref} - i_s(k))}{U_s(k)T_s} \right)$$ (30)

As the squared term $L_s^2$ multiplied with an error gives a negligible value, this term can be neglected in the duty calculation. Thus from Equation (26), the duty ratio, $d$, can be obtained as,

$$d = \left| \frac{i_{ref} - i_s(k + 2)}{C} \right|$$ (31)

where, $C$ is a constant.

$$C = \left| \frac{U_{opt}}{L_s} \right| T_s$$ (32)

Thus, from Equation (31), it can be observed that the duty ratio can be obtained based on a simple current error. This eliminates the complex calculations involving the slopes and the consideration of both quadrature axis and direct-axis currents will offer good control over the torque and flux response. Hence, the number of calculations are reduced and parameter dependency is eliminated in the proposed duty calculation method under the steady-state condition. As a result, there is a reduction in torque and flux ripples, while the simplicity of C-PCC is retained.

5 | SIMULATION RESULTS

The proposed method is compared with the C-PCC PCC [20] and PCC [37] in MATLAB-Simulink. The parameters used for the simulation are given in Table 1. The parameters such as sampling time and load conditions are maintained the same for all the models for better comparison. For $V_{dc} = 415$ V, $T_s = 100$ μs, $L_s = 10.5$ mH, from Equation (32), the constant value, $C$ is obtained as 2.65 in this method. The modern high-performance applications require quick speed and torque dynamics. To study the effect of the sudden change in speed and torque, all four methods are subjected to similar speed and torque dynamic changes. The dynamic conditions for step changes in speeds as well as load torque are shown in Figures 7 and 8 respectively. The simulation results of obtained speed, torque and flux are shown for all the methods. The effect of the sudden change in speed from 300 to 700 to 1200 rpm under no-load condition can be analysed from Figure 7, where all the methods exhibit near-identical performance. However, the steady-state performance exhibited between each speed change shows that the proposed method has lower torque and flux ripples. Similarly, the effect of dynamic change in load torque is presented in Figure 8, where the load torque is suddenly increased from 0% to 50% of the rated torque. The simulation results show that the proposed method can quickly respond to the sudden torque change. In addition, the ripples in flux and torque of proposed PCC under steady load condition are lesser than the other three methods. Thus, it can be observed from the simulation results that the proposed PCC has a better performance under steady-state no-load and on-load operating conditions. The flux and torque ripples in C-PCC are higher than PCC [20], PCC [37], proposed PCC and all the methods exhibit almost identical dynamic performance.

6 | EXPERIMENTAL RESULTS

A 5 HP, 1500 rpm, 415 V, 4 pole PMSM motor is used for the experimentation. All three methods are implemented in real-time for comparison. The PMSM is operated using a three-phase two-level VSI. The LEM sensors are used to sense the

| Table 1 | Machine parameters |
|---------|---------------------|
| Parameters | Value |
| DC link voltage ($V_{dc}$) | 415 V |
| Stator resistance ($R_s$) | 1.12 Ω |
| Stator inductance ($L_s$) | 10.5 mH |
| Rated speed ($N_r$) | 1500 rpm |
| Permanent magnet flux ($\psi_p$) | 0.71 |
| No. of pole pairs ($p$) | 2 |
| Sampling time ($T_s$) | 100 μs |
| Rated torque ($T_o$) | 24 Nm |
| Moment of inertia ($J$) | 0.005 kg·m² |
DC link voltage and stator currents. The 1024-point incremental shaft encoder gives the position and is fed back to sense the speed. The control is given through the dSPACE 1104 interface. The PMSM motor shaft is coupled to a DC-generator. The electrical load is supplied to the machine using the DC-generator connected to a resistive bank. The one-step delay compensation is provided in all the methods and the sampling frequency is 10 kHz. For the proposed model, the calculated value of $\bar{C} = 2.65$ is used. The sampling time and load conditions are maintained the same for all the methods. The hardware setup is shown in Figure 9.

### 6.1 Steady-state and dynamic response

The steady-state response at low (300 rpm), medium (700 rpm) and high speeds (1200 rpm) are observed as in Figures 10–13. The dynamic performance is verified by providing the three-stage change in speed from 300 to 700 to 1200 rpm as in Figure 14. The sudden change in the load torque (0%–50% of rated torque) at a speed of 1000 rpm is given in Figure 15. The harmonics in the stator current are obtained in the steady-state. Figure 16 shows the steady-state response with a load of 12 Nm applied to the motor. Furthermore, the dynamics of the machine during the speed reversal from 1000 rpm to −1000 rpm is also confirmed in Figure 17. The waveforms of stator flux and torque for the corresponding speed are shown in all the above results.

In the C-PCC method, the application of a single voltage vector for an entire sample time leads to over-regulation and hence it results in large torque and flux ripples in the steady-state response as shown in Figure 10. However, the duty-modulated methods help to reduce the torque and flux ripples with the application of two (or more) voltage vectors in a sample time, as it is evident from Figures 11–13. The waveforms of torque and flux for the particular speed are also
included in Figures 10–14. The experimental results at different speeds also prove that the proposed method produces less flux and torque ripples compared to C-PCC, PCC [20] and PCC [37]. The response of the machine subjected to dynamic changes in speed, shown in Figure 14, affirms that all four methods have similar dynamic performance. Moreover, the waveforms obtained for the dynamic step changes in speed (Figure 14) and step change in torque (Figure 15) highlights the advantage of the PCC method for modern speed control applications. Whenever a step change in speed is applied, the control algorithm quickly responds to adjust the speed to the required reference value. The steady-state performance, with a load of 12 Nm in Figure 16, comprises of the obtained stator flux, torque, speed and current waveforms. The obtained
FIGURE 14  Step changes in speed from 300 to 700 to 1200 rpm. (a) Conventional PCC, (b) PCC [20], (c) PCC [37] and (d) Proposed PCC. PPC, predictive current control

FIGURE 15  Step-variation in torque from 0% to 50% of rated load torque. (a) Conventional PCC, (b) PCC [20], (c) PCC [37] and (d) Proposed PCC. PPC, predictive current control

FIGURE 16  Steady-state performance with a load torque of 12 Nm. (a) Conventional PCC, (b) PCC [20], (c) PCC [37] and (d) Proposed PCC. PPC, predictive current control

FIGURE 17  Speed reversal from −1000 rpm to 1000 rpm. (a) C-PCC, (b) PCC [20], (c) PCC [37] and (d) Proposed PCC. PPC, predictive current control
The proposed method yields better performance under no-load and on-load conditions based on experimental results. Moreover, Figures 18–21 presents stator current total harmonic distortion obtained based on the steady-state stator currents. Thus, it can be confirmed that the proposed method has less current harmonic distortion as compared to the C-PCC, PCC [20] and PCC [37].

The torque and flux ripples are calculated as:

$$T_{ripped} = \sqrt{\frac{1}{n} \sum_{i=0}^{n} (T_i - T^*_i)^2}$$  \hspace{1cm} (33)

$$\psi_{ripped} = \sqrt{\frac{1}{n} \sum_{i=0}^{n} (\psi_i - \psi^*_i)^2}$$  \hspace{1cm} (34)

The computational burden for the three methods is given in Table 2. From Table 2, it can be observed that the computational burden for the proposed method is less than the existing duty method PCC [20] as the slope calculation involved in PCC [20] is eliminated in the proposed method. However, the computational burden of PCC [37] is slightly less than the Proposed PCC. In PCC [37], the redundant voltage vectors are eliminated and only three active voltage vectors are used for the selection of optimal voltage vector. This reduces the computational time. The method still requires complex control and identification of the position of the voltage vector with respect to the reference vector. It can be observed from Table 3 that, compared to C-PCC, PCC [20] and PCC [37], the ripples in the steady-state response of proposed PCC are considerably less.

Although the switching frequency of the proposed method is slightly more than the C-PCC, the overall improvement in steady-state performance, as well as reduction in computational burden, override this slight increase. The flux ripples, torque ripples and switching frequency ($f_{sw}$) are calculated and are presented in Table 3.

**Table 2. Computational burden**

| C-PCC  | PCC [20] | PCC [37] | Proposed PCC |
|--------|----------|----------|--------------|
| 36 µs  | 50 µs    | 38 µs    | 39.5 µs      |

Abbreviations: C-PCC, conventional PCC; PCC, predictive current control.
### 6.2 Parameter sensitivity test

The proposed method utilizes a parameter independent duty calculation for improving steady-state performance. In order to verify the effectiveness of the proposed method against parameter variations, the parameter sensitivity test is conducted. This is achieved by increasing stator resistance as well as the inductance values by 20% in the PCC algorithm.

The effect of stator resistance change can be observed in Figure 22. The machine is operated at 75 rpm (5% of the rated speed). The stator resistance is increased by 20% after 10 s and the response shows that all the methods are insensitive to changes in the stator resistance.

To analyse the effect of stator inductance change at high speed, the machine is operated at rated speed and the inductance is increased by 20% after 10 s. The responses obtained are shown in Figure 23. The performance of PCC [20] and PCC [37] gets disturbed due to the change in inductance.

After the change in inductance, the ripples are increased in both methods. However, in C-PCC and proposed method, the effect of inductance change is very small. The methods regain stability without much difference in the steady-state response after the inductance change.

In PCC [20] and PCC [37], the duty calculation depends on the inductance value and hence the performance is affected by its variation. Thus, the proposed method offers a robust,
parameter independent duty operation with reduced torque and flux ripples.

6.3 Discussion of results

Table 3 presents the torque ripple, flux ripple and switching frequency at different speeds under the steady-state condition for C-PCC, PCC [20], PCC [37] and proposed method. It can be observed from the table that the proposed method removes at least 55%–65% of the torque ripples existing in the C-PCC under different speed conditions. Furthermore, the torque ripple in the proposed method is 9.8% and 6.3% less than PCC [20] and PCC [37], respectively. The flux ripples at high speed are 36.6%, 4.8% and 9.2% lesser than C-PCC, PCC [20] and PCC [37], respectively, in the proposed method. The switching frequency of the proposed method is comparable to the available duty methods, PCC [20] and PCC [37], but greater than the C-PCC. However, this increase is trivial as there is a significant reduction in torque and flux ripples compared to the C-PCC.

The reduction of torque and flux ripples in the proposed method is effective due to the application of one active and one null vector in a sample time. The duty calculation is proportional to the error between the reference and predicted current in the proposed method. The quadrature and direct-axis currents are considered in the calculation, which results in the reduction of torque and flux ripples. The change in torque will be reflected as a change in quadrature axis current and the change in flux will cause a change in the direct-axis current. When the error is large, the duty is increased proportionally to reduce the deviation in predicted and reference current. Similarly, when the error is small, duty is also correspondingly reduced. Moreover, the constant value in the denominator ensures the duty calculation to be parameter independent. However, in PCC [37], the duty is proportional to the stator inductance, while in PCC [20], the duty is proportional to current slopes. This affects the steady-state performance when the parameters are inaccurate.

7 CONCLUSION

The paper presents a modified PCC method to mend the shortcomings of a C-PCC scheme, that is, the ripples in the torque and flux responses. C-PCC scheme is simple and has robustness against parameter variations. However, it is confined to the use of only one active voltage vector for an entire control interval and this leads to poor steady-state performance. The method developed in this paper aims to control the torque and flux through the stator current. This retains the simplicity of the PCC scheme while improving performance. The comparison of the proposed method with the C-PCC scheme, PCC [20] and PCC [37] are performed both through simulation and real-time experimentation. The obtained results confirm the effectiveness of the method. Although the switching frequency of the inverter obtained is slightly more than that of C-PCC, owing to the advantages of torque and flux ripple reduction and reduced computational burden, this increase is trivial. Further, the proposed method is robust to parameter variations which are experimentally validated.

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