A Deadbeat Current and Flux Vector Control for IPMSM Drive with High Dynamic Performance

That-Dong Ton and Min-Fu Hsieh

Department of Electrical Engineering, National Cheng Kung University, Tainan 70101, Taiwan;
ttdong.hueic@gmail.com
* Correspondence: mfhsieh@mail.ncku.edu.tw

Abstract: In this paper, a new deadbeat stator current and flux linkage vector control (DB-CFVC) scheme for interior permanent magnet synchronous machines (IPMSM) is proposed. The control structure is simplified by implementing the proposed flux linkage vector control method in the \( \alpha-\beta \) stationary coordinate. Unlike conventional deadbeat methods, the dynamic performance of the proposed DB-CFVC can be enhanced while voltage command saturation and over output current are avoided. This is achieved with a “reinforced” phase angle reference of stator flux linkage vector by considering rotor speed error and maximum voltage to properly enhance the quality of the calculated flux phase angle command. By predicting stator flux linkage and current in the stationary coordinate, the deadbeat direct flux linkage vector control based on the one-step delay compensation strategy becomes straightforward and exhibits low sensitivity to motor parameters compared to conventional methods performed in the rotating frame. Then, by developing a practical and robust hybrid flux linkage observer, the proposed DB-CFVC method can work more reliably and effectively than conventional methods. Simulations and experiments are conducted in a drive system for an IPMSM to evaluate the effectiveness and reliability of the proposed method.

Keywords: permanent magnet synchronous machine (PMSM); deadbeat current and flux vector control (DB-CFVC); flux linkage observer

1. Introduction

Permanent magnet synchronous machines (PMSMs) have been attractive for industrial applications owing to their excellent capability in high torque density and efficiency, rich configuration diversity, and low maintenance requirement compared to conventional types of motors [1,2]. However, high-performance motor drives are needed to maximize the performance of PMSMs. The field-oriented control (FOC) has been widely used for its facility to fit various types of PMSMs [3]. Nevertheless, its dynamic performance is restricted by indirect torque and flux control. The dynamic response can be enhanced by the classic direct torque control (DTC) [4], yet the hysteresis behavior causes high torque ripple. This can be mitigated by employing space vector modulation (SVM) [5], but a trade-off between high dynamics and low torque ripple should be considered.

Model predictive control (MPC), known as an attractive solution for high dynamic performance and high accuracy [6–10], is used to replace conventional vector control methods [5]. For classic MPC [9,10], an optimal active voltage vector is selected by cost function and imposed on a voltage source inverter (VSI) in a sampling period. Despite the strong dynamic performance obtained, high torque ripple and current harmonic still exists. Thus, a combination of more than one voltage vector in a fixed period with optimal duty cycles, called the duty-cycle MPC, was proposed to improve output torque quality [9]. However, this results in a rising computation burden by the cost function and optimization. In the deadbeat predictive control (DB-PC) [11–17], the aforementioned problems are resolved by precisely determining the voltage reference vector and then delivering it to
SVM to generate an output voltage at VSI. In the DB-PC, the desired values can be attained by one beat control, faster than conventional vector control methods, and it can neglect the complexity of controller parameter tuning. However, like MPCs, the DB-PCs highly depend on PMSM model, and its control quality is affected by processing delays, i.e., sampling and computing time. With the one-step delay compensation algorithm adopted [11,14–19], the delay problem has been resolved. Furthermore, the disturbance observer [12,15,16] or model-free [20] techniques were also developed to overcome the effects of motor parameter variations/mismatch and unknown disturbances.

The combination of the DB-PC methods with current control (CC), flux vector control (FVC), or direct torque and flux control (DTFC) has recently been investigated [13–17] to enhance control performance. In previous studies [15,16], the robust deadbeat predictive current control (DB-PCC) was presented with a sliding mode disturbance observer to obtain high-accuracy output current. Although accurate estimation of flux linkage and torque is not necessary in these DB-PCCs, the torque and flux linkage that are controlled indirectly via the rotor coordinate (d-q)’s current elements may produce a weaker dynamic response than a direct one. A modified DB-PCC for high-speed PMSM was presented [17], where the voltage command in the stationary frame (α-β) is calculated from the current elements in d-q frame indirectly via flux linkage elements in the α-β frame. However, this seems to be another way of employing conventional DB-PCC in α-β frame via the PMSM mathematical model. The deadbeat direct-flux vector control (DB-DFVC) developed in [18,19] and the DB-DTFC in [13,14] possess excellent dynamic performance, where both flux linkage and torque calculations are performed in the rotating frames of rotor or stator flux coordinate. Nonetheless, the predicted stator current and flux linkage in the one-step delay algorithm require more computation effort than those in the DB-PCC and relate much to motor parameters. In addition, coordinate transformation is necessary for most of the calculations in the rotating frame. For high-speed operations, the output voltage commands may likely exceed the allowable limit when large errors between command and feedback are combined with the high-bandwidth characteristic in the BD-PC. The voltage saturation would then restrict the dynamic performance. Overmodulation is often used to overcome this problem, but it also generates a non-sinusoidal output voltage resulting in increased torque ripple.

A hybrid DB-PCC with classic MPC [21] was developed to provide output current command delivered from an optimal geometrical solution. This avoids saturation of current and voltage and the dynamics are improved. A DB-DTFC method based on torque and flux vector trajectories in d-q frame with an improvement on a minimum-time ramp trajectory method was presented to mitigate the voltage limit problem [22], wherewith fast and stable transient dynamics for IPMSM drives can be achieved. Although the aforementioned studies [21,22] addressed the voltage and current limit issues to improve dynamics, they were still heavily linked to motor parameters. In the advanced hybrid DB-DTFC proposed in [23], the voltage command and flexible time duration were directly calculated based on torque and flux linkage errors considering voltage saturation. Having improved dynamic performance and torque ripple, compensation to the time delay and current limitation are not considered. A combination of deadbeat predictive torque control (DB-PTC) and duty ratio calculation for voltage, based on stator flux linkage vector in the stationary frame, was introduced with benefits for reducing computation and switching loss [24]. Nonetheless, these methods have been mostly applied to surface-mounted PMSM (SPMSM) to reduce complexity and it may still be a challenge for an implementation in IPMSM. On the other hand, a control algorithm for high efficiency, such as maximum torque per ampere (MTPA) [25], has usually been implemented for IPMSM drive. Since electromagnetic torque is typically used as an input command for the DB-DTFC or DB-DFVC schemes, it would be complex to determine flux linkage and/or load angle references from torque and to restrict the output current.

In this work, a deadbeat current and flux vector control (DB-CFVC) method is proposed, where a simple DB flux vector control in the stationary frame with the “reinforced
the reinforced phase angle command” concept is introduced. A practical and robust hybrid flux linkage observer is also developed. As a result, the dynamic performance is enhanced, the complexity of tuning the control/observer parameters is mitigated, and the quality of torque and flux is improved. Unlike other DB-DTFC or DB-DFVC, the input command of the proposed method is current magnitude, and the voltage command is determined by a deadbeat control based on stator flux linkage vectors in the stationary frame with improved current prediction for the one-step delay compensation algorithm. Therefore, the current magnitude can easily be restricted in an allowable range, resulting in a robust control strategy. Furthermore, differing from a previous study [17], the proposed method exploits the advantages of directly controlling the stator flux linkage vector. By using the reinforced phase angle command, the phase angle reference of the stator flux linkage vector can be accurately calculated based on the enhanced quality of load angle command considering actual speed error and voltage limit. They contribute to boosting dynamic response and simultaneously avoiding voltage command saturation. Finally, to confirm the effectiveness and reliability of the proposed method, simulations and experiments on an IPMSM drive are conducted to validate the proposed method, where with a comparison with other deadbeat control methods is made.

2. IPMSM Fundamental Model

The space vectors of voltage, current, and stator and rotor flux linkages of an IPMSM on different coordinate systems in steady state are shown in Figure 1, where the $\alpha$-$\beta$ frame is the stationary coordinate and the $d$-$q$ and $f$-$t$ frames are the rotor and stator flux linkage rotating coordinates, respectively.

![Figure 1. Phasor diagram used in analysis of IPMSM drive.](image)

The PMSM voltage equation in the stationary coordinate can be given as:

$$v_{\alpha\beta} = R_s i_{\alpha\beta} + \frac{d\psi_{\alpha\beta}}{dt}$$  \(1\)

where $v_{\alpha\beta} = [ v_\alpha \ v_\beta ]^T$, $i_{\alpha\beta} = [ i_\alpha \ i_\beta ]^T$ and $\psi_{\alpha\beta} = [ \psi_\alpha \ \psi_\beta ]^T$ are the stator voltage, current, and flux linkage vector in $\alpha$-$\beta$ frame, respectively; $R_s$ is stator winding resistance.

From Equation (1), the stator flux linkage in $\alpha$-$\beta$ frame can be computed as:

$$\psi_{\alpha\beta} = \int (v_{\alpha\beta} - R_s i_{\alpha\beta}) \, dt$$  \(2\)
\[
\theta_s = \tan^{-1}\left(\frac{\Psi_\beta}{\Psi_\alpha}\right) \tag{3}
\]

\[
\Psi_s = \sqrt{\Psi^2_\alpha + \Psi^2_\beta} \tag{4}
\]

where \(\theta_s\) and \(\Psi_s\) are phase angle and magnitude of the stator flux linkage vector, respectively.

The stator flux linkage can be calculated via the current model in \(d-q\) frame as:

\[
\begin{cases}
\Psi_d = \Psi_m + L_d i_d \\
\Psi_q = L_q i_q
\end{cases} \tag{5}
\]

\[
\Psi_s = \sqrt{\Psi^2_d + \Psi^2_q} \tag{6}
\]

\[
\delta = \tan^{-1}\left(\frac{\Psi_q}{\Psi_d}\right) \tag{7}
\]

\[
\theta_s = \delta + \theta_e \tag{8}
\]

where \(\Psi_d\) and \(\Psi_q\) are \(d-q\)-axis stator flux linkage elements, respectively; \(\Psi_m\) is flux linkage due to permanent magnet (PM) in rotor; \(\theta_e\) is rotor electrical angle; and \(\delta\) is load angle formed between stator and PM flux linkage vectors.

The electromagnetic torque \(T_e\) can be expressed as:

\[
T_e = \frac{3}{2} p \left| \vec{\Psi}_s \times \vec{i}_s \right| \tag{9}
\]

\[
T_e = \frac{3}{2} p (\Psi_\alpha i_\beta - \Psi_\beta i_\alpha) \tag{10}
\]

where "]\ denotes vectors, \(p\) is the number of pole pairs. Discretizing Equation (1) by the first-order Euler rule with a sampling period \(T_s\) yields:

\[
v_{\alpha\beta}(k+1) = R_s i_{\alpha\beta}(k) + \frac{\Psi_{\alpha\beta}(k+1) - \Psi_{\alpha\beta}(k)}{T_s} \tag{11}
\]

Equation (11) can be further used to determine voltage command in the stationary frame [5], where the desired \(\alpha-\beta\)-axis flux linkage elements can be calculated by:

\[
\Psi_{\alpha\beta}(k+1) = \begin{bmatrix}
\Psi_\alpha(k+1) \\
\Psi_\beta(k+1)
\end{bmatrix} = \Psi_s(k+1) \begin{bmatrix}
\cos \theta_s(k+1) \\
\sin \theta_s(k+1)
\end{bmatrix} \tag{12}
\]

with

\[
\theta_s(k+1) = \theta_s(k) + \Delta \theta_s(k+1) \tag{13}
\]

\[
\Delta \theta_s(k+1) = \delta(k+1) - \delta(k) + \omega_e(k) T_s \tag{14}
\]

where \(\Delta\) denotes deviation value and \(\omega_e\) is the angular speed of rotor flux linkage.

3. Proposed DB-CFVC Method for IPMSM Drive

This section introduces the principle of the proposed DB-CFVC in the stationary coordinate based on the one-step delay compensation algorithm. In addition, the stator flux linkage information plays a critical role in determining the performance of the proposed DB-CFVC method; hence, a modified hybrid flux observer strategy, which can yield accurate estimation, is developed in this study.

3.1. Principle of the Proposed DB-CFVC

The proposed DB-CFVC directly regulates the stator flux linkage vector with the current vector to achieve the expected dynamic response in each control beat. Unlike other direct flux vector control methods, the voltage command of the proposed method
is directly determined in the stationary frame by the predicted and desired flux linkage vectors without coordinate transformation. Moreover, the desired current delivered from the speed controller is employed as an input command. By adopting MTPA control, the current vector reference in the $d$-$q$ frame is identified and then used for the calculation of stator flux linkage magnitude and load angle reference.

For pure deadbeat (pure-DB) control, the voltage command can easily be obtained from Equations (11) and (12) by the measured/estimated and desired values of current and flux linkage. However, the disadvantage of the pure-DB is the inevitable delay caused by factors such as sampling, computation, and voltage modulation in a digital control system. Thus, this causes a delay for the modulated voltage command forwarded onto the VSI and leads to inaccuracies in control, especially in applications with high switching frequency. To avoid the performance drop caused by such a delay, the one-step delay compensation algorithm [14,15] is utilized for the deadbeat control in this work. In this method, the instant of updating the voltage command is delayed by one beat, and simultaneously, the measured/estimated feedback values used in calculating the voltage command are replaced by the predicted ones. By so doing, the calculated voltage command at the present control period will be updated at the beginning instant of the next sampling period. Hence, this resolves the delay problem without much computation effort added. The explanation of the control process and the phasor diagram analysis are illustrated in Figure 2.

![Figure 2](image_url)

**Figure 2.** (a) Control timeline and (b) phasor diagram analysis of proposed DB-CFVC in digital implementation based on the one-step delay compensation.

As presented by the graphical timeline shown in Figure 2a, it should be noted that the sampling frequency is synchronous with the switching frequency. Moreover, the superscripts "\(^n\)" and "\(^\text{ref}\)" denote the estimation, command/reference, and predicted values, respectively. Assuming at sampling period, \(kT_s\), the voltage command, \(v_{\alpha\beta}(k - 1)\), is calculated in the preceding sampling, \((k - 1)T_s\); then, they are modulated and updated onto VSI at the \(k\)th period corresponding to output voltage, \(v_{\alpha\beta}(k)\). Thus, the voltage \(v_{\alpha\beta}(k)\) can be regarded as constant within a control period \(T_s\) and it is estimated from \(v_{\alpha\beta}(k - 1)\) with deadtime compensation [26] by:

\[
v_{\alpha\beta}(k) = v_{\alpha\beta}^*(k - 1) + \Delta v_{\alpha\beta}^{\text{dt}}(k)
\]  

(15)
where $\Delta v_{\alpha\beta}^d(k) = \begin{bmatrix} \Delta v_{a}^d(k) \\ \Delta v_{b}^d(k) \end{bmatrix}^T$ is voltage error in $\alpha$-$\beta$ frame due to the deadtime effect of inverter switching at the $k$th period. The prediction moment $P_k$ in Figure 2a illustrates that this is the moment to predict the response values at the ending moment of period $kT_s$ after the voltage, $v_{\alpha\beta}(k)$, has been applied during this period. The complete control algorithm is presented in the block diagram shown in Figure 3. The detailed design of each step will be explained in the following sections.

![Figure 3. Complete block diagram of the proposed DB-CFVC.](image)

### 3.2. Proposed Hybrid Flux Linkage Observer

In this work, a modified hybrid flux linkage observer (MO-FOB) is developed by simplifying the manipulating process of the Gopinath observing strategy [27]. The complete observer scheme is shown in Figure 4. Unlike other observers, the proposed MO-FOB uses a closed-loop low-pass filter (LPF) estimator [28] for the voltage model and an open-loop linear calculating approach to mix the estimated results from the voltage and current models. The dominance of each model varies linearly according to the operating speed. This hybrid method can replace the conventional model-switching methods (i.e., using PI regulator). The detailed design will be explained later.

As reported in a previous study [29], the conventional Gopinath-model-based closed-loop flux linkage observer (CL-FOB) is a hybrid observer capable of combining both voltage and current models with closed-loop feedback for flux linkage estimation to overcome their individual disadvantages. The flux estimator by the conventional Gopinath model can be written as follows:

$$\hat{\Psi}_{\alpha\beta} = \frac{s^2}{s^2 + K_p s + K_i} \Psi_{\alpha\beta} + \frac{K_p s + K_i}{s^2 + K_p s + K_i} \hat{\Psi}_{\alpha\beta}$$

where $\Psi_{\alpha\beta}$ is stator flux linkage estimated by Gopinath method; $\hat{\Psi}_{\alpha\beta}$ = $[\hat{\Psi}_a \hat{\Psi}_b ]^T$ and $\Psi_{\alpha\beta}$ = $[\Psi_a \Psi_b ]^T$ are output stator flux linkage vectors and their elements in $\alpha$-$\beta$ frame estimated by the voltage model and current model, respectively; $K_p = \sqrt{2} \omega_C$ and
\[ K_1 = (\omega_b^G)^2 \] are PI gains of the observer, which are determined based on \( \omega_b^G \) which is the transition speed for switching between the current and voltage models. However, the conventional Gopinath observer is highly sensitive to the estimation errors at medium speed (around the model-switching region) since the voltage model uses a pure integrator \([29]\) and requires careful tuning for the PI coefficients of the observer. Therefore, some studies have modified the voltage model and improved the model-switching method to be more precise in the transition. However, this can make the observers more complicated.

**Figure 4.** (a) Modified hybrid flux linkage observer scheme and (b) a proposed hybrid strategy for flux linkage calculation.

In this study, the closed-loop LPF estimator \([28]\) is used to replace the pure integrator in the voltage model. The flux linkage of the voltage model adopting the closed-loop LPF estimator \([28]\) can be expressed by:

\[
\dot{\Psi}_{s}^V = \frac{1}{s + \omega_c} (V_{s} - R_s i_{\alpha \beta}) + \frac{\omega_c}{s + \omega_c} \begin{bmatrix} \cos \hat{\theta}_s \\ \sin \hat{\theta}_s \end{bmatrix}
\]

with

\[
\Psi_{\text{cor}}^s = \begin{cases} 
\Psi_{\text{m}}, & \Psi_{\text{m}} > \Psi_s^V \\
\Psi_{\text{m}}, & \Psi_{\text{m}} < \Psi_s^V 
\end{cases}
\]

where \( \Psi_{\text{cor}}^s \) is amplitude of corrective flux linkage, \( \Psi_{s}^V \) is magnitude of output stator flux linkage estimated by the voltage model, \( \Psi_{\text{m}} \) is magnitude of stator flux linkage reference, \( \hat{\theta}_s \) is flux linkage phase angle deduced from Equation (3) by \( \Psi_s^V \) and \( \omega_c \) is cut-off frequency of the LPF flux linkage estimator. As reported in \([28]\), this estimation method can significantly resolve the problem of phase drift and DC offset caused by pure integrator at medium and high speeds. As can be seen in Equation (17), the flux linkage is independent of rotor angle and is non-sensitive to most of the motor parameters, excluding stator resistance \( R_s \). This is the reason why it cannot work accurately at low speed when the resistance voltage drops are significant compared to the output voltage. Hence, as in the Gopinath model, the current model is predominantly conducted in low-speed regions. The flux linkage estimated by the current model is written as:

\[
\dot{\Psi}_{\alpha \beta} = T^{-1}(\theta_e) \left\{ L_{dq} \begin{bmatrix} \cos \theta_e \\ -\sin \theta_e \end{bmatrix} i_{\alpha \beta} + \Psi_{\text{pm}} \right\}
\]

where \( L_{dq} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \) is stator inductance matrix in \( d-q \) frame, \( \Psi_{\text{pm}} = \begin{bmatrix} \Psi_m \\ 0 \end{bmatrix}^T \) is PM flux linkage vector and its elements in \( d-q \) frame, and \( T(\theta_e) = \begin{bmatrix} \cos \theta_e & -\sin \theta_e \\ \sin \theta_e & \cos \theta_e \end{bmatrix} \) is
Park transformation with rotor angle \( \theta_c \), and \( T^{-1}(\theta_c) \) is inverse Park transformation. In contrast to the voltage model, the current model can perform well from low to high speeds; however, it is highly sensitive to motor parameters and requires rotor angle information, \( \theta_c \).

As mentioned above, the PI regulators of the switching strategy are replaced by a modified hybrid strategy based on speed for flux linkage calculation in the proposed observer. The output flux linkage estimation equals the combined result of the voltage model and current model through a linear selection gain defined by operating speed, as shown in Figure 4. The formula to calculate the output flux linkage is described as:

\[
\hat{\Psi}_{\alpha\beta}(k) = f\left(\hat{\Psi}^v_{\alpha\beta}(k), \hat{\Psi}^i_{\alpha\beta}(k)\right) = K_p(k)\hat{\Psi}^v_{\alpha\beta}(k) + (1 - K_p(k))\hat{\Psi}^i_{\alpha\beta}(k)
\]

with

\[
\left\{
\begin{array}{l}
K_p(k) = (\omega_c(k) - \omega_0) / (\omega_{base} - \omega_0) \\
0 \leq K_p(k) \leq 1 \\
0 \leq \omega_0 < \omega_{base}
\end{array}
\right.
\]

where \( \hat{\Psi}_{\alpha\beta}(k) = [\hat{\Psi}_a(k) \hat{\Psi}_\beta(k)]^T \) is output stator flux linkage vector and its elements in \( \alpha-\beta \) frame estimated by the proposed MO-FOB at \( k \)th instant, \( K_p(k) \) is the linear selection gain between \( \hat{\Psi}_{\alpha\beta}(k) \) and \( \hat{\Psi}_{\alpha\beta}(k) \), \( \omega_0 \) is a minimum speed to active mixing estimated values, and \( \omega_{base} \) is base speed of motor. Note that if \( \omega_c(k) < \omega_0 \), then the output flux linkage vector \( \hat{\Psi}_{\alpha\beta}(k) \) is set equal to the result estimated by the current model \( \hat{\Psi}_{\alpha\beta}(k) \). Thus, \( \omega_0 \) should be chosen close to the cutting frequency, \( \omega_c \), of the LPF estimator. If \( \omega_c(k) > \omega_{base} \), then \( \hat{\Psi}_{\alpha\beta}(k) \) is set equal to the estimated result by the voltage model \( \hat{\Psi}_{\alpha\beta}(k) \).

As shown in Figure 4a, the flux linkage correction component, \( \hat{\Psi}_{\alpha\beta} \), in the closed-loop LPF estimator used for the voltage model is delivered from the previous output flux linkage, which has been intermixed with the current model’s result. The greater the descent of the speed, the more dominant the current model and the more robust to resistance variation; it can thus assist in improving accuracy in self-correction for error of the LPF scheme at low and medium speed. At higher speed, the higher the signal-to-noise ratio of back electromotive force (BEMF), the more potent and dominant the voltage model. As a result, the observed flux linkage can ensure robustness to the impact of resistance variation and disturbances in severe load conditions for the entire speed range even though the PI regulators are not applied.

### 3.3. Stator Flux Linkage and Current Predictions

As presented above, the predicted values at the end moment of the \( k \)th period are the response values obtained after the \( \alpha-\beta \)-axis voltages, \( \mathbf{v}_{\alpha\beta}(k) \), are applied during period \( kT_s \). Thus, the stator flux linkage vector and current in \( \alpha-\beta \) axes predicted at the ending instant of the control period \( kT_s \) can be calculated by the measured current, i.e., \( \mathbf{i}_{\alpha\beta}(k) \), the updated voltage, \( \mathbf{v}_{\alpha\beta}(k) \), and estimated stator flux linkage, i.e., \( \hat{\Psi}_{\alpha\beta}(k) \), based on the voltage Equation (11) that is transformed into:

\[
\Psi^p_{\alpha\beta}(k) = \hat{\Psi}_{\alpha\beta}(k) + (\mathbf{v}_{\alpha\beta}(k) - R_i\mathbf{i}_{\alpha\beta}(k))T_s
\]

\[
\theta^p_{\alpha}(k) = \tan^{-1}\left(\frac{\Psi^p_\beta(k)}{\Psi^p_\alpha(k)}\right)
\]

where \( \Psi^p_{\alpha\beta}(k) = [\Psi^p_a(k) \Psi^p_\beta(k)]^T \) is the predicted stator flux linkage vector and its elements in \( \alpha-\beta \) frame at the ending instant of period \( kT_s \), and \( \theta^p_{\alpha}(k) \) is phase angle of
predicted stator flux linkage vector. Note that \( \hat{\Psi}_{\alpha\beta}(k) \) is estimated at the beginning of the \( k \)th period from the voltage \( v_{\alpha\beta}(k-1) \) and current \( i_{\alpha\beta}(k) \) using the proposed MO-FOB.

Since the control period \( T_s \) is sufficiently short compared to mechanical time constant of the motor, the rotor speed can be considered constant during one control period, resulting in the rotor angle being predicted by periodically adding an incremental angle \( \omega_c T_s \) over each control period. Thus, the load angle prediction can be deduced from Equation (8) by:

\[
\delta^P(k) = \theta^*_{d}(k) - (\theta_c(k) + \omega_c(k) T_s)
\]

(24)

where \( \delta^P(k) \) is the predicted load angle of flux linkage vector.

It is known that the voltage drops of resistance in Equation (11), i.e., \( R_s i_{\alpha\beta}(k) \), are insignificant to the output voltage, i.e., \( v_{\alpha\beta}(k) \), in operation at medium and high speed but should not be neglected at low speed. Nevertheless, the change of current magnitude, \( i_s \), is relatively small in each control period. Thus, the magnitude of the current vector can be considered constant in one control period. Its phase angle after each sampling period is approximately equal to the incremental angle produced by the rotor movement during one control period, i.e., \( \omega_c T_s \). As a result, the \( \alpha\beta \)-axis currents can be approximately predicted by a simple method, given by:

\[
i^P_{\alpha\beta}(k) = T(\omega_c(k) T_s) i_{\alpha\beta}(k)
\]

(25)

As observed in (22) and (25), the prediction of \( \alpha\beta \)-axis stator flux linkage and current in this work is simplified and is nearly independent of motor parameters (excluding \( R_s \)). This is different from other methods [14,15,17].

3.4. Stator Flux Linkage and Current Reference Calculations

To gain high efficiency in IPMSM drive operation, the MTPA algorithm is usually considered. Concerning MTPA implemented in the conventional DTFCs or DFVCs, the calculations of the reference/command values of the stator flux linkage magnitude and load angle from the torque reference are a complicated computation [30]. Hence, the look-up-table solution is usually used instead. Moreover, it is difficult to limit the output current directly by these methods. Conversely, it would be more straightforward to conduct the MTPA computation using current reference rather than torque one.

Therefore, in the proposed method, the current reference is chosen as input for the control loop. The current magnitude reference, \( i^*_d(k) \), is determined by a PI controller for the speed loop with output limited by the rated current. The gains of the PI controller are designed based on a method presented in [31].

According to the MTPA algorithm [32], the \( d-q \)-axis current references, \( i^*_{dq}(k) = \begin{bmatrix} i^*_d(k) & i^*_q(k) \end{bmatrix}^T \), are computed as given by:

\[
\begin{align*}
i^*_d(k) &= -M - \sqrt{M^2 + \frac{i^{2q^2}(k)}{2}} \\
i^*_q(k) &= \sqrt{i^{2q^2}(k) - i^{2q^2}(k)}
\end{align*}
\]

(26)

with

\[
M = \frac{\Psi_m}{4(L_d - L_q)}
\]

(27)

Then, the magnitudes of stator flux linkage vector reference, \( \Psi^*_s(k) \), and load angle reference, \( \delta^*(k) \), are calculated by (5), (6), and (7) using \( i^*_d(k) \) and \( i^*_q(k) \), respectively, as rewritten to be:

\[
\Psi^*_s(k) = \sqrt{(\Psi_m + L_d i^*_d(k))^2 + (L_q i^*_q(k))^2}
\]

(28)

\[
\delta^*(k) = \tan^{-1} \left( \frac{L_q i^*_q(k)}{\Psi_m + L_d i^*_d(k)} \right)
\]

(29)
3.5. Voltage Command Determination with “Reinforced Flux Phase Angle” Concept

As stated in Section 3.1 and depicted schematically in Figure 2a, for the one-step delay compensation implemented on digital devices [14–17], the voltage commands calculated at the \( k \)th period will be updated at the beginning of interval \((k+1)T_s\), but they are determined during the \( k \)th sampling period. These voltage commands will force the flux linkage errors between predicted and desired quantities at the beginning and ending instants of period \((k+1)T_s\), respectively, into zero. Consequently, the voltage commands at the \( k \)th instant are calculated by (11) based on control timeline shown in Figure 2a, expressed as follows:

\[
v_{\alpha\beta}^*(k) = R_s i_{\alpha\beta}(k + 1) + \frac{\Psi_{\alpha\beta}^*(k + 1) - \Psi_{\alpha\beta}(k + 1)}{T_s}
\]  

(30)

where \( v_{\alpha\beta}^*(k) = \begin{bmatrix} v_{\alpha}^*(k) & v_{\beta}^*(k) \end{bmatrix}^T \) is voltage vector command in \( \alpha-\beta \) frame determined at sampling step \( k \), \( \Psi_{\alpha\beta}^*(k + 1) = \begin{bmatrix} \Psi_{\alpha}^*(k + 1) & \Psi_{\beta}^*(k + 1) \end{bmatrix}^T \) is stator flux linkage vector reference in \( \alpha-\beta \) frame at instant \((k + 1)T_s\), \( i_{\alpha\beta}(k + 1) \) and \( \Psi_{\alpha\beta}(k + 1) \) are current and stator flux linkage vectors, respectively, in \( \alpha-\beta \) frame at the beginning instant of the sampling period \((k + 1)T_s\). The elements in Equation (30) will be determined in the following.

Since the \( i_{\alpha\beta}(k + 1) \) and \( \Psi_{\alpha\beta}(k + 1) \) are the future quantities that are inaccessible at the present instant \( kT_s \), they will be predicted from the measured/estimated values at the present, as presented in Section 3.3. Consequently, the quantities at the beginning instant of the \((k + 1)\)th period are taken to be equal to the predicted quantities at the \( k \)th period, as listed in the following:

\[
\Psi_{\alpha\beta}(k + 1) = \Psi_{\alpha\beta}^P(k)
\]  

(31)

\[
\theta_s(k + 1) = \theta_s^0(k)
\]  

(32)

\[
\delta(k + 1) = \delta^0(k)
\]  

(33)

\[
i_{\alpha\beta}(k + 1) = i_{\alpha\beta}^P(k)
\]  

(34)

As previously mentioned, the control period, \( T_s \), can be considered sufficiently short compared to the mechanical response time. Hence, the desired values of current and flux linkage can be reasonably retained for two successive sampling steps, which means that the demanded flux linkage magnitude and load angle at the \((k + 1)\)th instant can be acquired by:

\[
\Psi_{\alpha\beta}^P(k + 1) = \Psi_{\alpha\beta}^P(k)
\]  

(35)

\[
\delta^*(k + 1) = \delta^*(k)
\]  

(36)

According to the phasor diagram shown in Figure 2b, the flux linkage phase angle reference, \( \theta_s^*(k + 1) \), can be identified by:

\[
\theta_s^*(k + 1) = \theta_s(k) + \Delta\theta_s^P(k) + \Delta\theta_s^i(k + 1)
\]  

(37)

with

\[
\Delta\theta_s^P(k) = \delta^P(k) - \delta(k) + \omega_e(k)T_s
\]  

(38)

and

\[
\Delta\theta_s^i(k + 1) = \delta^*(k + 1) - \delta(k + 1) + \omega_e(k + 1)T_s
\]  

(39)

As mentioned earlier, the rotor speed is assumed to be constant over one control period, i.e., \( \omega_e(k) \approx \omega_e(k + 1) \). Then, substituting Equations (31), (32), (36), (38), and (39) into Equation (37), \( \theta_s^*(k + 1) \) is obtained as:

\[
\theta_s^*(k + 1) = \delta^*(k) - \delta(k) + \theta_s(k) + 2\omega_e(k)T_s
\]  

(40)

\[
\therefore \theta_s^*(k + 1) = \delta^*(k) + \theta_s(k) + 2\omega_e(k)T_s
\]  

(41)
In this study, a concept to “reinforce” the phase angle reference of stator flux linkage is proposed. According to the conclusion in [5], dynamic response is decided by the quality of load angle increment, and the flux phase angle increment can be restricted by the maximum output voltage. It can be noted from Equation (41) that the flux phase angle reference depends on load angle reference. Therefore, to achieve high dynamic performance, the flux phase angle reference should be appropriately reinforced by considering the actual speed error ratio and existing load angle reference. At the same time, this reinforced angle should be bound to ensure the output voltage is confined within an allowable value. Then, the reinforced flux phase angle, during control period $(k+1)T_s$, can be proposed as follows:

$$\theta_s^*(k+1) = \theta_s^*(k+1) + F_\theta(k+1)$$  \hspace{1cm} (42)

$$F_\theta(k+1) = K_\delta \frac{\omega_s^*(k) - \omega_s(k)}{\omega_s^*(k)} \delta^*(k+1)$$  \hspace{1cm} (43)

$$\Delta \theta_s^*(k+1) = \theta_s^*(k+1) - \theta_s^*(k)$$  \hspace{1cm} (44)

where $\theta_s^*(k+1)$ is reinforced phase angle reference of stator flux linkage vector at the $(k+1)^{th}$ instant, $F_\theta(k+1)$ is the amount of reinforced angle, $K_\delta$ is the percentage reinforcement of load angle, and $\Delta \theta_s^*(k+1)$ is new phase angle increment reference of flux linkage. Note that $F_\theta$ can influence the quality of current and torque generation. Thus, to prevent overshoot of current and high torque ripple, $K_\delta$ in Equation (43) should be chosen properly (here selecting $K_\delta = 0.1$).

Further, to ensure that the final flux phase angle increment reference is within a permissible value, the phase angle reference selection, $\theta_s^*$, should be concerned with the maximum phase angle increment, i.e., $\Delta \theta_s^\text{max}$. This maximum angle is constrained by the voltage limitation and magnitudes of the predicted and desired flux linkage vectors. By applying the cosine theorem to the phasor diagram shown in Figure 2b, the maximum phase angle increment calculated at the $(k+1)^{th}$ instant, $\Delta \theta_s^\text{max}(k+1)$, can be defined by:

$$\Delta \theta_s^\text{max}(k+1) = \cos^{-1} \left[ \frac{(\Psi_s^*(k+1))^2 + (\Psi_s(k))^2 - (V_s^\text{max}T_s)^2}{2\Psi_s^*(k+1)\Psi_s(k)} \right]$$  \hspace{1cm} (45)

where $V_s^\text{max} = \frac{V_{dc}}{\sqrt{2}}$ is a maximum voltage of VSI.

Then, the final desired phase angle, e.g., $\theta_s^*(k+1)$, is determined as follows:

$$\theta_s^*(k+1) = \begin{cases} \theta_s^*(k) + \Delta \theta_s^\text{max}(k+1), & \Delta \theta_s^\text{max}(k+1) \geq \Delta \theta_s^\text{max}(k+1) \\ \theta_s^*(k+1), & \Delta \theta_s^\text{max}(k+1) < \Delta \theta_s^\text{max}(k+1) \end{cases}$$  \hspace{1cm} (46)

According to Equations (12)–(14), the $\alpha$-$\beta$-axis stator flux linkage, $\Psi_{\alpha\beta}^*(k+1)$, can be derived based on its magnitude and phase angle reference, as given by:

$$\Psi_{\alpha\beta}^*(k+1) = \begin{bmatrix} \Psi_s^*(k+1) \\ \Psi_{\beta}^*(k+1) \end{bmatrix} = \Psi_s^*(k+1) \begin{bmatrix} \cos \theta_s^*(k+1) \\ \sin \theta_s^*(k+1) \end{bmatrix}$$  \hspace{1cm} (47)

Finally, the stator flux reference, $\Psi_{\alpha\beta}^*(k+1)$, is then combined with the predicted results of current and flux linkage from Equations (31) and (34) to compute the voltage vector command in $\alpha$-$\beta$ frame at the $k$th period, $\mathbf{v}_{\alpha\beta}^*(k)$, of the proposed DB-CFVC using (30).

### 3.6. Study on Motor Parameter Variation

The motor parameter variation can significantly influence the accuracy and performance of the control method. In this work, several simple estimation methods of parameter variations are conducted to secure the robustness and reliability of the proposed drive system.
When IPMSMs are in operation, stator inductances and resistance usually vary with various working conditions while PM flux linkage variation rarely occurs under normal operations. Hence, the PM flux linkage is assumed to be constant in this work. Regarding stator resistance variation, it only affects the flux linkage estimation in the proposed DB-CFVC scheme; however, this problem has been resolved by the proposed MO-FOB.

According to the analysis in [33–35], it can be realized that the effect of $d$-axis inductance error, $\Delta L_d$, on driving performance has been insignificant compared to that of the $q$-axis stator inductances, $\Delta L_q$. Thus, only the error $\Delta L_q$ is considered in this study. By the approximate calculation based on produced electromagnetic torque proposed in [36], the $q$-axis stator inductance can be estimated by:

$$\hat{L}_q(k) = \frac{L_q}{1 + K_T \hat{T}_{e,\text{rated}}(k)}$$

(48)

where $\hat{L}_q(k)$ is the estimated $q$-axis stator inductance, $\hat{T}_e(k)$ and $T_{e,\text{rated}}$ are the estimated and rated torques, respectively, and $K_T$ is torque constant. The estimated parameter (i.e., $\Delta L_q$) at the present instant will be updated in all the related calculations at the next control period.

4. Test Bench Configuration

A standard test bench system for the IPMSM drive to verify the effectiveness of the proposed DB-CFVC scheme is shown in Figure 5a. The specifications of the target IPMSM are given in Table 1. The core controller is a digital signal processor (DSP) of Texas Instrument, TMS320F28379D, with a sampling and PWM switching frequency of 10 kHz. An IGBT VSI is used to feed the IPMSM with current limited at 10 A and DC bus voltage supplied at 100 V. The rotor speed of the PMSM is calculated by taking the derivative of rotor position (i.e., $\theta_m$), in which $\theta_m$ is measured by an encoder with 5000 pulse/rev. The phase current and output torque are also measured independently by current and torque sensors. For data collection, the estimated values are exported via DAC ports of the DSP; then, the data are measured and analyzed by a Teledyne Leroy MDA803A oscilloscope. Furthermore, a real-time hardware-in-the-loop (HIL) MR series II toolkit, as shown in Figure 5b, is used to evaluate tests as motor parameters change instead of the real-test platform [3] (since the HIL allows conveniently changing the motor parameters).

The PI gains of the speed loop are chosen as the method reported in [31], i.e., $K_{p\omega} = 0.1256$ and $K_{I\omega} = 0.3933$.

Table 1. Target IPMSM specifications.

| Specification                        | Value | Unit (SI) |
|--------------------------------------|-------|-----------|
| DC voltage, $V_{dc}$                 | 100   | V         |
| Rated current, $i_{\text{rated}}$    | 10    | A         |
| Maximum current, $i_{\text{max}}$    | 20    | A         |
| Base speed, $N_{\text{base}}$        | 2300  | rpm       |
| Rated torque, $T_{e,\text{rated}}$   | 2.9   | N.m       |
| Pole-pairs, $p$                      | 4     |           |
| Stator resistance, $R_s^0$           | 315   | m$\Omega$ |
| Permanent magnet flux linkage, $\Psi_m$ | 48.2  | mWb       |
| Torque constant, $K_T$               | 0.2892| N.m/A     |
| $d$-axis inductance, $L_d^0$         | 2.03  | mH        |
| $q$-axis inductance, $L_q$           | 2.84  | mH        |
5. Evaluation Results

5.1. Assessment of Proposed MO-FOB Scheme

To evaluate the effectiveness of the proposed flux linkage observer, the simulation results of conventional CL-FOB and proposed MO-FOB are generated using MATLAB-Simulink with the IPMSM given in Table 1. There are two scenarios regarding parameter mismatch by considering magnetic saturation and temperature variation to evaluate the robustness of the two flux observer methods discussed above. In Scenario A, the stator resistance, i.e., $R_s^A$ is increased to 150% of the original $R_s^0$ (i.e., $R_s^A = 1.5 R_s^0$), and simultaneously the $d$- and $q$-axis inductances are both reduced to 75% of the original ones (i.e., $L_d^A = 0.75 L_d^0$ and $L_q^A = 0.75 L_q^0$). In Scenario B, the $q$-axis inductance is assumed to be unchanged (i.e., $L_q^B = L_q^0$) while the resistance and $d$-axis inductance vary as in Scenario A (i.e., $R_s^B = 1.5 R_s^0$ and $L_d^B = 0.75 L_d^0$). This aims to assess the sensitivity level of the closed-loop hybrid observers to resistance and $d$-axis inductance variations. The error percentages of the flux linkages estimated by the two methods based on the actual flux linkage are summarized in Table 2.

Table 2. Error percentages of the estimated flux linkages (simulation result).

| Operation Condition | Conventional Closed-Loop Hybrid Flux Observer (CL-FOB) | Modified Hybrid Flux Observer (MO-FOB) |
|---------------------|-----------------------------------------------------|---------------------------------------|
| Speed (rpm)         | Load (Nm)                                           | Scenario A | Scenario B | Scenario A | Scenario B |
| 100                 | 1.25                                                | 4.39%      | 0.95%      | 2.09%      | -0.41%     |
| 700                 | 1.25                                                | 7.48%      | 2.32%      | 2.75%      | 1.03%      |
| 2300                | 2.5                                                 | 8.37%      | 3.15%      | 3.28%      | 2.39%      |

As can be seen in Table 2, the error percentages of the estimation by the proposed MO-FOB are much lower than those of the conventional CL-FOB under various operating conditions. Moreover, the influence of parameter variations (especially changes in resistance...
and $d$-axis inductance) on the application of the proposed hybrid flux observer can be considered insignificant, even when operating under rated conditions.

5.2. High Dynamic Performance Verification

To examine the dynamic performance of the proposed DB-CFVC, the experiments are performed, and the results are compared with two benchmark methods, i.e., the DB-PCC [11] and DB-DFVC [18]. All candidates are configured with the same sampling and control periods and speed loop bandwidth for a fair comparison. In these experiments, rotor speed, output torque, stator flux linkage, and current are considered in the evaluation of dynamic performance and output quality such as speed response, operating current (as shown in Figure 6a), torque ripple, and flux linkage ripple (as shown in Figure 7). Furthermore, the voltage commands of the three methods are analyzed and compared (as shown in Figure 6b) to track the voltage saturation problem. Two test scenarios include: (a) speeding up from standstill to the base speed (2300 rpm) under heavy load (2.5 Nm), and (b) speeding down gradually with a few steps to a low speed (100 rpm) to assess the driving performance of the three candidates (i.e., the proposed DB-CFVC, the conventional DB-PCC, and DB-DFVC) in the transient and steady states.

Figure 6. Experiment results of dynamic performance comparison between DB-PCC, DB-DFVC, and proposed DB-CFVC: (a) comparison of response time as sped up to 2300 rpm at load torque 2.5 Nm; (b) zoom-in at the points (1–3) in (a) to analyze voltage command waveforms.
Figure 7. Comparison of speed, torque, flux linkage, and current between conventional DB-PCC, DB-CFVC, and proposed DB-CFVC at steady state under load torque of 2.5 Nm at various speed references of (a) 2300 rpm, (b) 1300 rpm, and (c) 100 rpm.
The results displayed in Figure 6a compare the dynamic performance of the proposed DB-CFVC with the other methods in terms of speed response and output currents under the same load conditions and speed commands. As can be seen, the conventional DB-PCC shows the lowest dynamic response with large current spikes in the transient states when the speed descends. With the conventional DB-DFVC, the output speed cannot attain the base speed under the heavy load condition. In the two conventional methods, their voltage commands both exceed the allowable limit (and then they saturate at 58 V), as seen in Figure 6b-1,b-2, or the voltage commands fluctuate while the current exceeds the limit during speeding down, as seen in Figure 6b-1. The possible reasons are that, in the transient state when the motor speed is close to the base speed, the generated BEMF would be large. When this is combined with a large difference (or error) between the reference and feedback values of the variables calculated in the synchronous frame, e.g., the torque and flux linkage in the DB-DFVC or the \(d-q\)-axis current in the DB-PCC, the voltage command obtained is likely to exceed the allowable range after being computed with a high bandwidth deadbeat controller. In practice, in the conventional DB-PCC case (as shown in Figure 6b-1), the actual and reference \(d-q\)-axis currents’ direction may alter rapidly during fast speeding down and affect each other. Consequently, the voltage commands may be out of range. This would further result in rapid voltage fluctuations and overcurrent due to the reaction from the \(d-q\)-axis current regulated by the deadbeat control to follow the step-falling speed command. Regarding the conventional DB-DFVC case (as displayed in Figure 6b-2), the \(f\)-axis voltage saturates at the voltage limit while the \(f\)-axis voltage is nearly maintained. Thus, the magnitude of the output voltage vector is insufficient as truncated by modulation of the SVPWM, resulting in output torque shortage and fluctuation (causing output speed (2200 rpm) to be lower than the reference value (2300 rpm)). For the proposed DB-CFVC result (as shown in Figure 6b-3), the voltage limit issues have been included in the computations of stator flux linkage reference using the “reinforced” phase angle concept (as depicted in Section 3.5). Thus, the \(a-b\)-axis voltage commands are constantly ensured to be lying in the allowable range while the motor dynamic performance is enhanced and output speed effectively tracks the reference values, as seen in Figure 6a.

From the comparisons presented in Figure 7a–c, it can be observed that the proposed DB-CFVC method has slightly improved the quality of outputs compared to the conventional ones since its current, torque, and flux linkage waveforms have seemed to be smoother than the others under the same operating conditions. The conventional DB-PCC appears to have the lowest performance, working at low speeds. Meanwhile, in the conventional DB-DFVC method, output speed cannot reach the base speed under heavy load conditions (since the voltage command saturates early, as explained above).

5.3. Parameter Sensitivity Analysis for Proposed DB-CFVC

In this section, the robustness of the proposed DB-CFVC scheme in regard to motor parameter variation is evaluated via experiments on the HIL device (Figure 5b). The examinations are conducted under the scenario of simultaneous resistance and inductance mismatches during operation. These tests are used to thoroughly investigate the robustness of the proposed DB-CFVC to variations of the winding resistance and \(d\)-axis inductance and the effectiveness of using the estimated \(q\)-axis inductance, \(\hat{L}_q\).

In this case, it is assumed that the winding resistance is varied to 150% of the original one (i.e., \(R'_s = 1.5R_0^s\)), and the \(d-q\)-axis inductances are varied to 75% of the original ones simultaneously (i.e., \(L'_d = 0.75L_0^d\) and \(L'_q = 0.75L_0^q\)). Then, the IPMSM is driven at low and high speeds under heavy load (2.5 Nm). There are two test cases: without and with updating the value of the estimated \(q\)-axis inductance, \(\hat{L}_q\). The comparison between the two cases (i.e., without and with updating \(\hat{L}_q\)) is shown in Figure 8a (at base speed, 2300 rpm) and Figure 8b (at low speed, 100 rpm), where the left side of each figure shows the result without updating \(\hat{L}_q\).
In particular, the torque and flux ripple increase when the actual stator flux linkage (HIL result).

Moreover, the flux linkage estimation error increases due to variation of \( q \)-axis inductance; however, it seems to be insignificant and independent of the operating speed.
The results on the right side of Figure 8a,b show that the drive system improves as the estimated $q$-axis inductance variation is updated. This demonstrates the usefulness of the $q$-axis inductance estimation. It also confirms the robustness of the proposed DB-CFVC method to mismatch of resistance and $d$-axis inductance.

6. Conclusions

In this article, a novel DB-CFVC for IPMSM drives has been proposed. Overall, the proposed method relies on stator flux linkage vectors, i.e., magnitude and load angle, to directly determine the voltage commands in the stationary coordinate by utilizing the one-step delay computation algorithm. Compared to conventional methods, the proposed one can minimize the computation cost and reduce sensitivity by omitting coordinate transformation and simplifying the strategies of flux linkage estimation and current prediction strategies. Furthermore, by applying the reinforced phase angle reference which takes into account the rotor speed error and the voltage limit, saturation of the voltage vector command can be avoided in transient states. This improves the dynamic performance under heavy loads and high speeds. With the estimated and updated $q$-axis inductance, the sensitivity of the proposed method to PMSM nonlinearity during operation can be significantly reduced. The proposed method has also been verified to be reliable and effective by experiments on the IPMSM drive test bench.

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