Predictive Torque Control of Three-level Sparse Neutral Point Clamped Inverter fed IPMSM Drives using Simplified Deadbeat Principle

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Abstract. This paper proposes a predictive torque control for interior permanent magnet synchronous machine (IPMSM) driven by three-level sparse neutral point clamped inverter. It contributes to greatly diminish the torque and flux ripples by using predictions and three voltage levels. To precisely synthesize the voltage vectors, deadbeat principle is employed. Different from existing works, the proposed control method is implemented in stationary reference frame, eliminating coordinate transformations. Furthermore, one cycle delay is compensated through predictions. In addition, to further reduce the torque and flux ripples, three-level sparse neutral point clamped inverter (3L-SNPCI) is employed. In comparison with other types of three-level inverters, it utilizes fewer power semiconductors and has alleviated neutral point voltage fluctuation problem. Space vector modulation is employed to generate the switching signals for the 3L-SNPCI. The validity of the proposed approach is verified by experimental results.

1. Introduction
The classical direct torque control (DTC) features simplicity, fast dynamics and robustness. However, it suffers from high torque ripples, variable switching frequency as well as large current harmonics [1]. In the past decades, research efforts have been made to overcome these problems. Unfortunately, most of them come with significantly increased computational complexity and the tedious controller tuning tasks. For instance, various space vector modulation (SVM) based DTC approaches have been proposed in recent years [2-4]. Despite of their excellent performances, most of these approaches are implemented in the rotating $d$-$q$ reference frame that requires coordinate transformation, leading to increased complexity. Moreover, predictive control is not used in these approaches, and hence, one cycle digital delay can not be compensated. In addition, controller tuning is still necessary in many controllers [2] [4]. Some recent research developments on predictive torque control (PTC) offer simplified controller tuning process while providing compensation for one cycle digital delay [5-6]. Nevertheless, these methods have high computational burden compared to the SVM based DTC due to their iterative evaluation of certain cost function. Besides, precise synthesis of the desired voltage vector (VV) is not realizable, limiting the improvements of torque and flux control. To synthesize the desired VV in a more accurate way, a deadbeat model predictive torque control with discrete SVM is proposed [7]. This approach utilizes deadbeat principle to determine the reference VV in $d$-$q$ frame. Then, three discretized virtual VVs that are adjacent to the reference VV are selected as candidates and are iteratively evaluated for torque and flux ripple minimization. Despite of its advantages, this
approach only approximates the reference VV by discretized virtual VVs. Thus, its performance is dependent on the accuracy of such approximations, which can be unsatisfying. Besides, its computational burden is still high with the presence of coordinate transformation and iterative optimization. A simplified predictive torque and flux control is implemented in the stationary $\alpha - \beta$ frame in [8]. Reduced computational burden and low torque ripples are reported. Nonetheless, the requirement of extra sliding mode observer and iterative optimization implies that its reduction of computational complexity is limited. Moreover, the basic assumption of $i_d = 0$ in [12] is not true for many applications that demand minimum copper loss.

To further reduce torque ripples and current harmonics, one of the most effective solutions is to employ multi-level inverters. For example, the conventional three-level inverter (3LI) is already widely used in commercial electric drive systems at the expense of increased semiconductor count. To decrease the number of power semiconductors, T-type 3LI is proposed for low voltage applications. It eliminates six diodes compared to the conventional 3LI. In practice, further reduction of semiconductor count is possible by using the three-level sparse neutral point clamped inverter (3L-SNPCI) with only ten switches [9]. The 3L-SNPCI offers the benefits of conventional 3LI with alleviated neutral point voltage (NPV) fluctuations. Therefore, it is a promising candidate for low voltage but high performance electric drive systems. In view of its merits, the 3L-SNPCI is employed in this paper to feed the interior permanent magnet synchronous machine (IPMSM).

To overcome the afore-mentioned difficulties, this paper proposes a PTC scheme for 3L-SNPCI fed IPMSM drives using simplified deadbeat principle. The proposed method is implemented in stationary reference frame without coordination transformation, extra observer or controller tuning. It also contributes to compensate for the one cycle digital delay via predictions. Space vector modulation is adopted in the proposed scheme to synthesize the reference VV while achieving NPV balance.

2. Model of IPMSM and proposed PTC using deadbeat principle

2.1. Mathematical Model of the IPMSM

Mathematically, the IPMSM can be modelled in the stationary $\alpha - \beta$ reference frame as

$$ L_q \frac{d i_s}{d t} = -R_s i_s + J \omega_r (\lambda_s - L_q i_s) + v_s $$

(1)

$$ \frac{d \lambda_s}{d t} = v_s - R_s i_s $$

(2)

$$ T_e = \frac{3}{2} P (\lambda_s \times i_s) $$

(3)

where $R_s$ stator resistance

$L_q$ stator inductance in q-axis

$\omega_r$ rotor electrical speed

$v_s$ stator voltage vector with $v_s = [v_{s\alpha} \ v_{s\beta}]^T$

$i_s$ stator current vector with $i_s = [i_{s\alpha} \ i_{s\beta}]^T$

$\lambda_s$ stator flux vector with $\lambda_s = [\lambda_{s\alpha} \ \lambda_{s\beta}]^T$

$T_e$ electromagnetic torque

$P$ number of pole pairs

$j$ +90° phase shift

2.2. Conventional Predictive Torque Control

By discretizing (1)-(3) via forward Euler method and re-writing them in scalar forms, equations (4)-(6) can be derived. where notations $k$ and $T_s$ represent the $k$-th sampling instant and the controller sampling period, respectively.

$$ [i_{s\alpha}(k + 1)] = \begin{bmatrix} 1 - \frac{R_s T_s}{L_q} & -\omega_r T_s \\ \omega_r T_s & 1 - \frac{R_s T_s}{L_q} \end{bmatrix} [i_{s\alpha}(k)] + \begin{bmatrix} 0 \\ \frac{\omega_r T_s}{L_q} \end{bmatrix} \left[ \frac{\lambda_{s\alpha}}{L_q} \right] + \begin{bmatrix} \frac{\omega_r T_s}{L_q} \\ \frac{T_s}{L_q} \end{bmatrix} \left[ v_{s\alpha}(k) \right] $$

(4)
\[
\begin{align*}
\frac{\lambda_{sa}(k+1)}{\lambda_{sb}(k+1)} &= \frac{\lambda_{sa}(k)}{\lambda_{sb}(k)} + T_s \left( \frac{v_{sa}(k)}{v_{sb}(k)} - R_s \frac{i_{sa}(k)}{i_{sb}(k)} \right) \\
T_e(k+1) &= \frac{3}{2} p \left[ \lambda_{sa}(k+1)i_{sb}(k+1) - \lambda_{sb}(k+1)i_{sa}(k+1) \right]
\end{align*}
\] (5)

In real-time digital control, the presence of one-cycle delay usually leads to increased torque and flux ripples. To compensate for this effect, the torque and flux should be predicted for one more step, i.e. the prediction of \( \lambda_s(k+2) \) and \( T_e(k+2) \) are needed. This can be done by following the flow chart in Fig. 1.

**Fig. 1.** Flow chart of predictive torque control.

2.3. Proposed Simplified PTC using Deadbeat Principle

In PTC, the reference torque \( T_e^* \) and flux amplitude \( |\lambda_e^*| \) are generally determined by speed regulator and the maximum torque per ampere trajectory, respectively, as illustrated by the block diagram in Fig. 2. Thus, when the deadbeat torque and flux regulation principle of \( T_e^* = T_e(k+2) \) and \( |\lambda_e^*| = [\lambda_{sa}^2(k+2) + \lambda_{sb}^2(k+2)]^{\frac{1}{2}} \) is applied, the analytical expressions of \( v_{sa}(k+1) \) and \( v_{sb}(k+1) \) can be derived correspondingly. This is because the values of two variables can be fixed with the presence of two governing equations. As a consequence, the desired reference VV can be directly computed, eliminating the necessity of iteratively searching for the optimal reference VV [8]. This contributes to greatly reduce the computational burden especially for multi-level inverter fed machine drives with numerous candidate voltage vectors.

**Fig. 2.** Reference torque and flux amplitude calculation.

Theoretically, the application of multiple VVs to precisely synthesize the reference VV in each sampling cycle will generate much smaller torque and flux ripples compared to the usage of only one or two VVs for reference VV synthesis in most PTC approaches. For better illustration, the derivation of expressions of \( v_{sa}(k+1) \) and \( v_{sb}(k+1) \) are given as follows, where notations \( M_1, M_2, N_1, N_2 \) are introduced for simplicity.

\[
\begin{align*}
M_1 &= \lambda_{sa}(k+1) - T_s R_s i_{sa}(k+1) \\
M_2 &= \lambda_{sb}(k+1) - T_s R_s i_{sb}(k+1)
\end{align*}
\]
\[
\begin{align*}
N_1 &= \left(1 - \frac{R_s T_s}{L_q}\right) i_{sa}(k + 1) - (\omega_r T_s) i_{sb}(k + 1) + \frac{\omega_r T_s}{L_q} \lambda_{sb}(k + 1) \\
N_2 &= (\omega_r T_s) i_{sa}(k + 1) + \left(1 - \frac{R_s T_s}{L_q}\right) i_{sb}(k + 1) - \frac{\omega_r T_s}{L_q} \lambda_{sa}(k + 1)
\end{align*}
\]  

(7)

By using the above notations, the predicted stator currents at \((k + 2)\)-th sampling instant is derived as
\[
\begin{align*}
i_{sa}(k + 2) &= N_1 + \frac{T_{psa}(k + 1)}{L_q} \\
i_{sb}(k + 2) &= N_2 + \frac{T_{psb}(k + 1)}{L_q}
\end{align*}
\]  

(8)

Then, the deadbeat torque and flux governing rules are re-expressed as (9) and (10), respectively.
\[
T_e^* = \frac{3P}{2} \left\{ (M_1 + T_s v_{sa}(k + 1)) i_{sb}(k + 2) - (M_2 + T_s v_{sb}(k + 1)) i_{sa}(k + 2) \right\}  
\]  

(9)

\[
|\lambda_s^*| = \left[ (M_1 + T_s v_{sa}(k + 1))^2 + (M_2 + T_s v_{sb}(k + 1))^2 \right]^{\frac{1}{2}}  
\]  

(10)

From (9), the relation between \(v_{sa}(k + 1)\) and \(v_{sb}(k + 1)\) can be figured out as
\[
v_{sb}(k + 1) = \mu_1 + \mu_2 \cdot v_{sa}(k + 1)  
\]  

(11)

where \(\mu_1, \mu_2\) are constants given by
\[
\mu_1 = \frac{2 T_e^*}{3P} - M_1 N_2 + M_2 N_1, \quad \mu_2 = \frac{M_2}{L_q} - \frac{N_2}{L_q}
\]

Substituting (10) into (9) and re-arranging the terms, the VV governing equation is derived as
\[
\begin{align*}
(1 + \mu_2^2)v_{sa}^2(k + 1) + 2(M_1 + M_2 \mu_2 + \mu_1 \mu_2) v_{sa}(k + 1) + (\mu_1^2 + 2M_1 \mu_1 + M_1^2 + M_2^2 - |\lambda_s^*|^2) = 0
\end{align*}
\]  

(12)

Obviously, there are two solutions for \(v_{sa}(k + 1)\). The smaller solution is chosen since the larger one generally results in a synthesized voltage vector that exceeds the voltage limit of DC/AC inverter as illustrated by Fig. 3. After the determination of \(v_{sa}(k + 1), v_{sb}(k + 1)\) is obtained via (10). Thus, the reference VV is determined and will be subsequently synthesized by SVM.

3. Fundamentals of the 3L-SNPCI and SVM Scheme

3.1. Fundamentals of the 3L-SNPCI

For better illustration, the circuit topology and space voltage vector distribution of the 3L-SNPCI are given in Fig. 4 and Fig. 5, respectively. The 3L-SNPCI can be divided into a front-end dual buck converter, which generates three voltage levels \((+V_{dc}/2, 0, -V_{dc}/2)\), and a back-end two-level voltage source inverter as shown in Fig. 4. The switching combinations of \(S_{A1}\) to \(S_{A4}\) that generate three voltage levels. Compared to the conventional three-phase 3LI and T-type 3LI that possess 27 VVs, the 3L-SNPCI has only 21 VVs with the absence of 6 medium VVs. This is illustrated in Fig. 5, where the numbers 2, 1, and 0 denote the voltage levels \(+V_{dc}/2, 0, \text{and} -V_{dc}/2\), respectively. The unavailability of medium VVs is advantageous from the perspective of NPV balancing. Nonetheless, this also means the synthesis of reference VV using only the large \((V_i, i \in [1,6])\) and small \((V_i, i \in [7,18])\) VVs requires base VV combinations that are different from those in the SVM schemes for conventional three-phase 3LI or T-type 3LI. Hence, the selection of base VV combinations under various operating conditions is an important issue.
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### 3.2. SVM Scheme

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space voltage vector in Fig. 5 is divided into six equivalent triangular sectors, i.e. sector $S_1$ to $S_6$. As an example, sector $S_1$ is illustrated in Fig. 6. It can be a further divided into four regions $R_1$ to $R_4$. The other sectors can be divided in the same manner using the enclosing VVs as summarized in Table I. Such a division satisfies the smooth VV switching constraint of 3LIs, i.e. three-phase line-line voltages are only allowed to be changed by at most one voltage level at each VV transition.

By referring to Fig. 6, it is seen that regions $R_1, R_2, R_3, R_4$ in sector $S_1$ are enclosed by the VVs tabulated in the first column of Table 1. Thus, for a given reference VV $V_{ref}$, its region can be determined on the basis of its magnitude $|V_{ref}|$ and phase angle $\theta$. Define the modulation index as $m = \frac{|V_{ref}|}{\frac{2}{3}V_{dc}}$. Then, the region location of $V_{ref}$ is determined by (13)-(16).

$$R_1: m \leq \frac{\sqrt{3}}{2(\sqrt{3}\cos\theta + \sin\theta)} \quad \text{&} \quad 0^\circ \leq \theta < 60^\circ \quad (13)$$
$$R_2: \frac{\sqrt{3}}{2(\sqrt{3}\cos\theta + \sin\theta)} < m \leq \frac{1}{\sqrt{3}\sin\theta + \cos\theta} \quad \text{&} \quad 0^\circ \leq \theta < 30^\circ \quad (14)$$
$$R_3: \frac{\sqrt{3}}{2(\sqrt{3}\cos\theta + \sin\theta)} < m \leq \frac{1}{2\cos\theta} \quad \text{&} \quad 30^\circ \leq \theta < 60^\circ \quad (15)$$
$$R_4: \begin{cases} \frac{\sqrt{3}}{\sqrt{3}\sin\theta + \cos\theta} < m \leq \frac{1}{\sqrt{3}\cos\theta + \sin\theta} & \text{&} \quad 0^\circ \leq \theta < 30^\circ \\ \frac{1}{2\cos\theta} < m \leq \frac{\sqrt{3}}{\sqrt{3}\cos\theta + \sin\theta} & \text{&} \quad 30^\circ \leq \theta < 60^\circ \end{cases} \quad (16)$$

After choosing the suitable base VV combination in each sector, the duty ratios of different VV can be computed by following the fundamental volt-second balance in each switching cycle, which is similar to the conventional SVM schemes. Hence, it is not further discussed here for conciseness. Noticeably, there are redundant VVs in Table 1, such as $V_7/V_8$ and $V_9/V_{10}$. The choice of redundant VVs is made based on NPV balancing requirements.

4. Experimental results and discussion

To verify the effectiveness of the proposed method, experiments are carried out on a laboratory set-up shown in Fig. 7. The digital signal processor (DSP) Texas Instrument TMSF28335 is used to implement the proposed algorithm. The IPMSM is mechanically coupled to a DC machine and results are sampled using dSPACE DS1106. Firstly, to study its transient performance, speed reversal test is carried out. The result is illustrated in Fig. 8. It is seen from this figure that the speed, torque, flux as well as the capacitor voltages are well regulated during the transient of speed reversal. Meanwhile, small torque and flux ripples are obtained. Secondly, the superiority of the proposed method over the conventional PTC is investigated with the same sampling cycle. The results of load test with the conventional PTC and the proposed deadbeat based PTC are shown in Fig. 9 and Fig. 10, respectively. By comparing the first and second subplots of the two figures, it is apparently seen that the torque and flux ripples obtained by the proposed method are much smaller than those of the conventional PTC. Moreover, the comparison of the third subplot reveals the attenuation of current harmonics and THD by employing the proposed method. A further comparison of NPV balancing in subplots 4 also show the better NPV balancing capacity of the proposed method.

5. Conclusion

This paper proposes a simplified predictive torque control approach for three-level sparse neutral point clamped inverter fed interior permanent magnet synchronous motor drives. It utilizes deadbeat principle to reduce computational burden and compensates for digital delay via predictions. It obtains superior torque and flux regulations as well as significantly reduced current harmonics.
Fig. 7. Experimental set-up in lab.

Fig. 8. Experimental results of IPMSM drive with speed reversal from -1000 rpm to 1000 rpm using the proposed method: (1) rotor speed; (2) electromagnetic torque; (3) stator flux; (4) capacitor voltages $V_{c1}$ (blue) and $V_{c2}$ (red).

Fig. 9. Experimental results of steady-state response of IPMSM drive with a rated load of 6 Nm at (a) 1000RPM (b) 200RPM using the conventional PTC with one cycle delay compensation: (1) rotor speed; (2) electromagnetic torque; (3) stator flux; (4) capacitor voltages $V_{c1}$ (blue) and $V_{c2}$ (red).
Fig. 10. Experimental results of steady-state response of IPMSM drive with a rated load of 6 Nm at (a) 1000RPM (b) 200RPM using the proposed method: (1) rotor speed; (2) electromagnetic torque; (3) stator flux; (4) capacitor voltages $V_{c1}$ (blue) and $V_{c2}$ (red).

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