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Voltage disturbance observer for dual three-phase PMSM system

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Abstract—In this paper, a new compensation strategy design of the dead-time effects and other VSI non-linearities is presented for a dual three-phase PMSM. The strategy is based on an estimation of compensation voltages for each motor sub-system by the proposed disturbance observer. This observer is based on the Kalman filter algorithm and the knowledge of the motor model structure and its parameters. The proposed approach is verified by experiments on the test bench which consists of the experimental dual three-phase PMSM and set of three-phase power stages controlled by one Tricore AURIX TC275 microcontroller. The achieved results are compared with the results of the standard compensation strategy.

Index Terms—Dead-time compensation, Kalman filter, dual three-phase Permanent Magnet Synchronous Motor (PMSM), Voltage Source Inverter (VSI)

I. INTRODUCTION

In the last decades, the permanent magnet synchronous motors (PMSM) have gained great popularity in industrial applications. Now, PMSMs are also starting to become popular in the automotive industry. The standard three-phase variants of the PMSM are most often applied. However, the multi-phase PMSMs have many advantages compared to standard three-phase PMSMs, especially in applications that are safety-critical and require a certain level of redundancy. High demands are placed on the multi-phase motor control algorithms. These algorithms require not only precise currents, position, and speed measurement but also precise application of the command voltages by the voltage source inverter (VSI). Unfortunately, the precise application of the command voltages is influenced by non-linearities of the VSI. As a result, the action values calculated by the control algorithm differ from the real applied voltages. These deformations of the inverter output voltages cause a harmonic distortion of the phase currents, the ripple of dq -axis currents and the torque pulsation [1], [2]. This issue has a negative effect on the quality of machine control. Therefore, it is advisable to compensate for inverter non-linearities.

The simplest solution of this issue is to use the so-called standard compensation strategy [3] that uses a feed-forward

approach. The average loss of the phase voltages during the PWM period is compensated depending on the phase currents polarity. The polarity detection near to zero-crossing region is critical [4]. In [2], this issue is fixed by the phase current observer that is based on the Kalman filter algorithm. Its disadvantage is the dependence on the accurate knowledge of the motor model structure and its parameters. The VSI model-based methods [5]–[7] are aimed at including the snubber and parasitic capacitance into the mathematical VSI model. The precise current sampling or high sampling current rate is required by these methods. In [8], the dead-time compensator for an asymmetrical six-phase machine using a resonant controller in the anti-synchronous (xy) reference frame is presented. The resonant PI controller is set to suppress only the dominant harmonic component. In [9], the compensation voltage in the motor model for the observer is composed of two parts. The first constant part represents the average voltage drop related to PWM period. The second part depends on the actual phase current value. The estimation of the two-component compensation voltages has a negative effect on the computational complexity of the observer.

This paper describes the design of the voltage disturbance observer for dual three-phase PMSM which is controlled by a single Tricore AURIX TC275 microcontroller. The PMSM is fully symmetrical. The motor sub-systems are separated electrically and also in the software layer. Both sub-systems have dedicated one microcontroller core and use the same software functions. The design of the compensation strategy is adapted to this concept. The voltage disturbance observer is designed for the reduction of the VSI output voltage deformation. It is based on the Kalman filter algorithm and knowledge of the dual three-phase PMSM model structure ([10], [11]) and its parameters. As a result, the observer is able to estimate the disturbance voltages for each sub-system. The proposed observer is validated on the real motor system. The experimental results confirm the usefulness of the proposed approach.

II. VOLTAGE DISTORTION ANALYSIS

The issue of dead-time and the reason for deformation of the output voltage can be explained on one leg of a three-phase inverter. The dead-time serves as a protection time,

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Fig. 1. (a) One leg of the three-phase VSI during the dead-time period. (b) Gating signal patterns and VSI output voltages.

which prevents the simultaneous conduction of both switching devices in one inverter leg. By inserting a dead-time period T_d into the PWM control signals, it is ensured that both switching devices S_H/S_L are turned off during the protection time. As a result, the output voltage depends on the actual direction of the phase current flow, which can only flow through the diodes D_H/D_L , as is shown in Fig.1(a). The deformation of the output voltage is caused by the protection time T_d as is shown in Fig.1(b).

Based on the behavior analysis, the voltage error ΔV_{an} can be defined with respect to the switching period T_{PWM} as

$$\Delta V_{an} = v_{dead} \cdot \text{sign}(i_a) = \frac{T_d + t_{on} - t_{off}}{T_{PWM}} V_{DC} \cdot \text{sign}(i_a) \quad (1)$$

where DC bus voltage is represented by V_{DC} . The switching delay time turn-on and turn-off are described by t_{on} and t_{off} . The sign function of the phase current can be represented as

$$\text{sign}(i_a) = \begin{cases} 1, & i_a > 0 \\ -1, & i_a < 0 \end{cases} \quad (2)$$

The magnitude of the voltage constant v_{dead} can be determined by the equation (3). This term depends on diodes forward voltage (V_d) and saturation voltages (V_{sat}) of the switching devices.

$$v_{dead} = \frac{T_d + t_{on} - t_{off}}{T_{PWM}} (V_{DC} - V_{sat} + V_d) + \frac{V_{sat} + V_d}{2} \quad (3)$$

III. OBSERVER DESIGN

The above analysis shows that the dead-time and other VSI non-linearities have an undesirable effect on the output voltage of the inverter. This deformation causes harmonic distortion of the phase currents and negatively affects the quality of the control algorithm. Therefore, it is advisable to compensate for the deformations of the VSI output voltage.

One possible solution and probably the easiest one is to use feed-forward compensation based on the calculation of compensation voltages by the equation (1). However, to achieve the required compensation precision, it is necessary to know the exact magnitude of the voltage constant v_{dead} and the polarity of each phase current during the current PWM period. The polarity detection of the phase currents is a challenging task especially in the area of the zero-crossing passage [2]. Further, the magnitude of the voltage v_{dead} varies depending on the

motor current operating conditions in practice. This voltage is very difficult to measure without the hardware intervention in the VSI. In addition to the above equations, (1)-(3) are only an approximation of the actual voltage errors.

For these reasons, the solution that uses the estimated compensation voltages obtained by disturbance observers is proposed. The voltage disturbance observers are designed on the basis of the Kalman filter and knowledge of the dual three-phase PMSM model.

A. Model of the dual three-phase PMSM

The model of the dual three-phase PMSM is derived in the rotor coordinate system. The derivation assumes a fully symmetrical PMSM system. It is further assumed that the subsystems are not phase shifted and their behavior is the same. When these conditions are satisfied, the system can be described by the voltage equation (4) that is extended by the disturbance voltage Δv_{dq} .

$$v_{dq} = R i_{dq} + \frac{dL i_{dq}}{dt} + \omega_e L i_{dq} + \omega_e \Lambda_M + \Delta v_{dq} \quad (4)$$

The voltage vector v_{dq} and current vector i_{dq} contain voltages and currents of the both sub-systems and can be written as follows

$$v_{dq} = (v_d^{S1}, v_q^{S1}, v_d^{S2}, v_q^{S2})^T \quad (5)$$

$$i_{dq} = (i_d^{S1}, i_q^{S1}, i_d^{S2}, i_q^{S2})^T \quad (6)$$

The vector Λ_M indicates back electromotive force (BEMF) constant λ_m for q -axis part of the sub-systems model.

$$\Lambda_M = (0, \lambda_m, 0, \lambda_m)^T \quad (7)$$

The electrical angular velocity is represented by ω_e . The symbol R denotes the matrix of stator resistances and symbol L represents the matrix of inductances.

$$R = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & 0 & R_s & 0 \\ 0 & 0 & 0 & R_s \end{bmatrix} \quad (8)$$

$$L = \begin{bmatrix} L_d & L_{dq} & M_{dd} & M_{dq} \\ L_{dq} & L_q & M_{dq} & M_{qq} \\ M_{dd} & M_{dq} & L_d & L_{dq} \\ M_{dq} & M_q & L_{dq} & L_q \end{bmatrix} \quad (9)$$

The matrix of inductances L contains direct and quadrature inductances L_d and L_q , mutual inductance L_{dq} between d - and q - axes within a single sub-system and mutual inductances between sub-systems M_d , M_q and M_{dq} . The mutual inductance $M_{dq} = 0$ for a fully symmetrical system. It is the result of the transformation from abc to dq - coordinate system. If the system is symmetrical with phase-shifted subsystems, then M_{dq} will be constant. In other cases, the mutual inductance M_{dq} is a function of the electrical position θ_e . In this case, it is better to create a PMSM model in the abc - coordinate system, the model in the dq - coordinate system is meaningless. In the

case of the symmetrical system, the mutual inductance L_{dq} equals zero even in the case of phase-shifted subsystems. In other cases, the mutual inductance L_{dq} is again a function of the electrical position θ_e .

$$\Delta \mathbf{v}_{dq} = (\Delta v_d^{S_1}, \Delta v_q^{S_1}, \Delta v_d^{S_2}, \Delta v_q^{S_2})^T \quad (10)$$

The vector $\Delta \mathbf{v}_{dq}$ represents the distortion of the output voltage which includes VSI non-linearities that are transformed into the dq -coordinate system. Subsequently, the equation (4) is adjusted to the form where derivatives of dq -currents are converted to the left side as follows

$$\frac{d\mathbf{i}_{dq}}{dt} = \mathbf{L}^{-1}(\mathbf{v}_{dq} - \Delta \mathbf{v}_{dq}) - \mathbf{L}^{-1} \mathbf{R} \mathbf{i}_{dq} - \mathbf{L}^{-1} \omega_e (\mathbf{L}_M + \mathbf{L}_c \mathbf{i}_{dq}) \quad (11)$$

where $\mathbf{L}_c$ represents a cross-coupling matrix. The mutual inductances $L_{dq} = 0$ and $M_{dq} = 0$.

$$\mathbf{L}_c = \begin{bmatrix} L_{dq} & L_q & M_{dq} & M_{qq} \\ -L_d & -L_{dq} & -M_{dd} & -M_{dq} \\ M_{dq} & M_{qq} & L_{dq} & L_q \\ -M_{dd} & -M_{dq} & -L_d & -L_{dq} \end{bmatrix} \quad (12)$$

B. Voltage disturbance observer for sub-system S_1

On the basis of equation (11), the state-space models for both sub-systems can be assembled that are identical in both cases. In this chapter, voltage disturbance observer equations for the subsystem S_1 are derived.

The state equations for the sub-system can be generally described as follows

$$\frac{d}{dt} \mathbf{x}^{S_1} = \mathbf{A}^{S_1} \mathbf{x}^{S_1} + \mathbf{B}^{S_1} \mathbf{u}^{S_1} \quad \mathbf{y}^{S_1} = \mathbf{C}^{S_1} \mathbf{x}^{S_1} \quad (13)$$

The state vector $\mathbf{x}^{S_1}$ contains $i_d^{S_1}$ and $i_q^{S_1}$. Further, the $\mathbf{x}^{S_1}$ is extended by disturbance voltages $\Delta v_d^{S_1}$ and $\Delta v_q^{S_1}$ which will be used for the compensation.

$$\mathbf{x}^{S_1} = [i_d^{S_1}, i_q^{S_1}, \Delta v_d^{S_1}, \Delta v_q^{S_1}]^T \quad (14)$$

The input vector $\mathbf{u}^{S_1}$ includes real voltages ($v_d^{*S_1}$, $v_q^{*S_1}$, $v_d^{*S_2}$, $v_q^{*S_2}$) of both sub-systems S_1 and S_2 that were applied to the VSI. Furthermore, the current of the sub-system S_2 and BEMF constant are contained.

$$\mathbf{u}^{S_1} = [v_d^{*S_1}, v_q^{*S_1}, v_d^{*S_2}, v_q^{*S_2}, \lambda_m, i_d^{S_2}, i_q^{S_2}]^T \quad (15)$$

The output vector $\mathbf{y}^{S_1}$ is given as

$$\mathbf{y}^{S_1} = [i_d^{S_1}, i_q^{S_1}]^T \quad (16)$$

The state matrix $\mathbf{A}^{S_1}$, input matrix $\mathbf{B}^{S_1}$ and output matrix $\mathbf{C}^{S_1}$ of the continuous state-space representation can be written as follows

$$\mathbf{A}^{S_1} = \begin{bmatrix} n_1 & n_3 & -\frac{1}{L_d} & 0 \\ n_2 & n_4 & 0 & -\frac{1}{L_q} \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \quad \mathbf{C}^{S_1} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \quad (17)$$

$$\mathbf{B}^{S_1} = \begin{bmatrix} L_5 & L_2 & L_3 & L_1 & n_9 & n_5 & n_7 \\ L_2 & L_6 & L_1 & L_4 & n_{10} & n_6 & n_8 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} \quad (18)$$

The substitutions L_{1-6} and n_{1-10} are introduced in matrices $\mathbf{A}^{S_1}$ and $\mathbf{B}^{S_1}$. The substitutions are defined as follows

$$n_1 = -L_5 R_s - L_2 L_d \omega_e - L_1 M_{dd} \omega_e \quad (19)$$

$$n_2 = -L_2 R_s - L_6 L_d \omega_e - L_4 M_{dd} \omega_e \quad (20)$$

$$n_3 = L_5 L_q \omega_e - L_2 R_s + L_3 M_{qq} \omega_e \quad (21)$$

$$n_4 = L_2 L_q \omega_e - L_6 R_s + L_1 M_{qq} \omega_e \quad (22)$$

$$n_5 = -L_3 R_s - L_1 L_d \omega_e - L_2 M_{dd} \omega_e \quad (23)$$

$$n_6 = -L_1 R_s - L_4 L_d \omega_e - L_6 M_{dd} \omega_e \quad (24)$$

$$n_7 = L_3 L_q \omega_e - L_1 R_s + L_5 M_{qq} \omega_e \quad (25)$$

$$n_8 = L_1 L_q \omega_e - L_4 R_s + L_2 M_{qq} \omega_e \quad (26)$$

$$n_9 = -\omega_e (L_2 + L_1) \quad n_{10} = -\omega_e (L_6 + L_4) \quad (27)$$

where

$$L_1 = 0 \quad L_2 = 0 \quad (28)$$

$$L_3 = -\frac{M_{dd}}{(L_d^2 - M_{dd}^2)} \quad L_4 = -\frac{M_{qq}}{(L_q^2 - M_{qq}^2)} \quad (29)$$

$$L_5 = -\frac{L_d}{(L_d^2 - M_{dd}^2)} \quad L_6 = -\frac{L_q}{(L_q^2 - M_{qq}^2)} \quad (30)$$

The elements L_1 and L_2 are equal to zero because the system is fully symmetrical ($L_{dq} = 0$ and $M_{dq} = 0$).

Subsequently, the state-space system (13)-(18) is converted using the Euler approximation from continuous to discrete-time assuming a suitable sampling period T_s . The noise of the process $\mathbf{w}_z^{S_1}$ and noise of the measurement $\mathbf{v}_z^{S_1}$ are included in the discrete state representation that can be described as follows:

$$\mathbf{x}_z^{S_1}(k) = \mathbf{A}_z^{S_1}(k) \mathbf{x}_z^{S_1}(k-1) + \mathbf{B}_z^{S_1}(k) \mathbf{u}_z^{S_1}(k-1) + \mathbf{w}_z^{S_1}(k-1) \quad (31)$$

$$\mathbf{y}_z^{S_1}(k) = \mathbf{C}_z^{S_1}(k) \mathbf{x}_z^{S_1}(k) + \mathbf{v}_z^{S_1}(k)$$

The discrete state matrix $\mathbf{A}_z^{S_1}(k)$ is given as

$$\mathbf{A}_z^{S_1}(k) = (\mathbf{I} + T_s \mathbf{A}^{S_1}) = \begin{bmatrix} 1 + n_1 T_s & n_3 T_s & -\frac{T_s}{L_d} & 0 \\ n_2 T_s & 1 + n_4 T_s & 0 & -\frac{T_s}{L_q} \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (32)$$

where $\mathbf{I}$ denotes the identity matrix. The discrete input matrix $\mathbf{B}_z^{S_1}(k)$ is given as

$$\mathbf{B}_z^{S_1}(k) = \mathbf{B}^{S_1} T_s = \begin{bmatrix} L_5 T_s & L_2 T_s & L_3 T_s & L_1 T_s & n_9 T_s & n_5 T_s & n_7 T_s \\ L_2 T_s & L_6 T_s & L_1 T_s & L_4 T_s & n_{10} T_s & n_6 T_s & n_8 T_s \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} \quad (33)$$

Fig. 2. Control algorithm integration

The discrete output matrix $C_z^{S_1}(k)$ determines which states are measured. In our case, these are dq - currents of the sub-system S_1 .

$$C_z^{S_1}(k) = C^{S_1} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \quad (34)$$

The assembled discrete state-space model (31)-(34) for a given sub-system can be used within the Kalman filter algorithm (35) to estimate the state vector $\hat{x}_z^{S_1}$. The symbol ' $\hat{\cdot}$ ' denotes the estimated state vector ($\hat{x}_z^{S_1}$). A priori and a posteriori estimates of the vectors and matrices are denoted by the upper index symbols ' $-$ ' and ' $+$ ', respectively. K and P represent the Kalman gains matrix and the covariance matrix.

$$\begin{aligned} \hat{x}_z^{S_1-}(k) &= A_z^{S_1}(k) \hat{x}_z^{S_1+}(k-1) + B_z^{S_1}(k) u_z^{S_1}(k-1) \\ P^{S_1-}(k) &= A_z^{S_1}(k) P^{S_1+}(k-1) A_z^{S_1}(k)^T + Q^{S_1} \\ K^{S_1}(k) &= P^{S_1-}(k) C_z^{S_1}(k)^T \\ &\quad \left[C_z^{S_1}(k) P^{S_1-}(k) C_z^{S_1}(k)^T + R^{S_1} \right]^{-1} \\ \hat{x}_z^{S_1+}(k) &= \hat{x}_z^{S_1-}(k) + K^{S_1}(k) \left[y_z^{S_1}(k) - C_z^{S_1}(k) \hat{x}_z^{S_1-}(k) \right] \\ P^{S_1+}(k) &= \left[I - K^{S_1}(k) C_z^{S_1}(k) \right] P^{S_1-}(k) \end{aligned} \quad (35)$$

The electrical velocity ω_e can be considered as constant because the ω_e is changing slowly with respect to the electrical quantities of the PMSM. On the basis of this assumption, the standard equation of the Kalman filter algorithm can be used.

The covariance matrix of the process noises Q^{S_1} and covariance matrix of the measurement noises R^{S_1} are chosen as diagonal.

$$Q^{S_1} = \text{diag} [q_{11}, q_{22}, q_{33}, q_{44}] \quad R^{S_1} = \text{diag} [r_{11}, r_{22}] \quad (36)$$

Subsequently, the estimated compensation voltages $\Delta v_d^{S_1}$ and $\Delta v_q^{S_1}$ are applied to the inputs of current controller blocks of the individual sub-systems where they are added to controller command voltages and decoupling voltages before the saturation.

IV. EXPERIMENTAL RESULTS

The test bench consists of the experimental dual three-phase PMS motor which is driven by a set of three-phase power stages controlled by a single Tricore AURIX TC275 microcontroller. Three standalone cores are located in this microcontroller. The sub-systems are separated not only electrically but also on the software layer. Each sub-system has dedicated one microcontroller core. Both cores use the same software functions. Communication between cores provides necessary

Fig. 3. The dual three-phase experimental motor System configuration

control variables for voltage disturbance observer. The last core is dedicated to data acquisition and communication. The whole platform is connected to the MATLAB/Simulink using UDP communication channel. The internal structure of the power inverter can be seen in Figure 2.

The core that was responsible for calculating the control algorithm and algorithm of the voltage disturbance observer of a given sub-system used a sampling rate of 10 kHz. The parameters of the PMSM are listed in Table III. The motor position is measured by the (GMR/AMR) sensor and is processed by position pre-processing block. Subsequently, an angle tracking observer (ATO) is used to reduce the measurement noise and to reconstruct motor speed ω_r^{ato} . The processed electrical speed $\omega_e = \omega_e^{ato} = p_p \omega_r^{ato}$ enters to the observer as a parameter. For this reason, all elements of the matrices $A_z^{S_i}$ and $B_z^{S_i}$ can't be pre-calculated. These matrices must be updated at each step of the KF algorithm. The adjustable elements of covariance matrices Q^{S_i} and R^{S_i} were determined experimentally. In the present configuration of the control algorithm, the same motor parameters and covariance matrices settings must be used for both sub-systems observers.

The dual three-phase experimental motor is connected to the dynamometer. The dynamometer operates as a speed reference during experiments. The experimental motor operates in torque mode. The dead-time effect is the most noticeable if the motor operates with low load torque. The torque setpoint was configured to 0.25 Nm for this reason. The dynamometer speed was configured to 50 rpm. This mechanical speed is equal to the electrical speed of 110 rad/s. The dead-time was set to 1 μ s. The configuration of the system is shown in Figure 3.

The experiments were performed for three different output voltage deformation compensation configurations. In the first case, no compensation strategy was applied. In the second case, the standard compensation strategy according to (3) was used. In the last case, the compensation voltages are estimated by the proposed voltage disturbance observer. The results are presented in steady-state under the aforementioned operating conditions. Differences between motor currents of both sub-systems without and with compensation of the output voltage

Fig. 4. Sub-system S_1 - phase currents and dq -axes currents without/with standard compensation and with the proposed voltage disturbance observer.

Fig. 5. Standard compensation voltages $\Delta v_{dead_dq}^{S1}$ and $\Delta v_{dead_dq}^{S2}$

TABLE I
HARMONIC DISTORTION OF PHASE CURRENTS

Method	5. harm.	7. harm.	11. harm.	13. harm.
	HRI_5 [%]	HRI_7 [%]	HRI_{11} [%]	HRI_{13} [%]
	S_1 / S_2	S_1 / S_2	S_1 / S_2	S_1 / S_2
Without comp.	2.27 / 2.21	1.38 / 1.46	0.74 / 0.64	0.37 / 0.43
Stand. comp.	0.27 / 0.60	0.20 / 0.10	0.45 / 0.14	0.10 / 0.09
Dist. observer	0.49 / 0.55	0.09 / 0.11	0.15 / 0.15	0.08 / 0.09

deformations caused by non-linearities of the VSI can be seen in Fig. 4 and Fig. 6. In the case without compensation, these voltage deformations cause the harmonic distortion of the phase currents waveforms as they are shown in Fig. 4(a) and Fig. 6(a). Through Clarke and Park transformations, the harmonic distortion of the phase currents is reflected by the ripple of the dq - currents as it is shown in Fig. 4(b)-(c) and Fig. 6(b)-(c). These current harmonic distortions are indicated by the presence of the parasitic odd harmonic components (5^{th} , 7^{th} , 11^{th} and 13^{th} harmonics are dominant) in the spectrum of the a - phase current and multiples of 6^{th} harmonic component in the spectrum of the dq - axis currents, as it can be seen in Fig. 8 and Fig. 9.

These undesirable phenomena can be partially suppressed by the proposed voltage disturbance observers. The output voltage distortions are compensated by estimated disturbance voltages that are shown for both sub-systems in Fig. 7 (For comparison, the compensation voltages of the standard compensation strategy are shown in Fig. 5). As a result, the dominant odd parasitic harmonic components of the phase

Fig. 6. Sub-system S_2 - phase currents and dq -axes currents without/with standard compensation and with the proposed voltage disturbance observer.

Fig. 7. Compensation voltages Δv_{dq}^{S1} and Δv_{dq}^{S2}

TABLE II
HD INDEX - EXPERIMENTAL RESULTS

Method	HD index [%]	
	Subsystem S_1	Subsystem S_2
Without compensation	2.7928	2.7552
Standard compensation	0.5680	0.6341
Disturbance observer	0.5283	0.5897

current are suppressed and the ripple of the dq - currents are reduced as it is evident from Fig. 4(g)-(i) and Fig. 6(g)-(i).

The parasitic harmonic components of the PMSM currents can be used as an indicator of the distortion rate. An index HRI_n (Harmonic Ratio of the Current) is defined as $HRI_n = I_n / I_1 \cdot 100\%$ where I_1 is the fundamental harmonic of the a -phase current and I_n represents n^{th} parasitic harmonic component. The resulting HRI_n index values are shown in Tab. I. The effectiveness of the proposed compensation approach is demonstrated by using harmonic distortion index of a - phase current that is defined as $(HD = \sqrt{I_5^2 + I_7^2 + I_{11}^2 + I_{13}^2} / I_1 \cdot 100\%)$. A comparison of the achieved HD index results for different compensation strategies is given in Tab. II. The proposed compensation strategy achieves a lower value of the HD index than the standard compensation method however the standard compensation strategy achieves a better suppression of the dominant 5^{th} harmonic component for the sub-system S_1 . This can be caused by minor differences in parameters of both sub-systems.

The existing fault-operation control algorithm for the dual three-phase PMSM has a computational cost 56 μs . The

Fig. 8. Sub-system S_1 - comparison of currents spectra without/with standard compensation and with the proposed voltage disturbance observer.

Fig. 9. Sub-system S_2 - comparison of currents spectra without/with standard compensation and with the proposed voltage disturbance observer.

proposed observer has a computational cost 21 μ s.

V. CONCLUSION

This paper describes the design of the voltage disturbance observer for dual three-phase PMSM that is based on the Kalman filter algorithm and knowledge of the motor model structure and its parameters. The proposed compensation algorithm is verified on the real system. The results show that the proposed compensation approach achieves a significant suppression of negative effects caused by the dead-time effect and the other non-linearities of the VSI. The advantage is the independence on the type of switching element and the other parameters of the inverter. The weakness of the presented approach is the dependence of the observer on accurate knowledge of motor parameters. The implementation complication

of the proposed approach is higher demand on the computing power of the control platform. Future work will be focused on testing the robustness of the proposed observer algorithm under various operating conditions. It is also planned to implement a control algorithm with a compensation algorithm on a six-core platform (AURIX TC397) with higher computing power. A higher sampling rate of the control algorithm and compensation strategy is expected.

REFERENCES

- [1] Z. Tang and B. Akin, "A New LMS Algorithm Based Deadtime Compensation Method for PMSM FOC Drives," *IEEE Transactions on Industry Applications*, vol. 54, no. 6, pp. 6472–6484, 2018.
- [2] W. Dafang, Y. Bowen, Z. Cheng, Z. Chuanwei, and Q. Ji, "A Feedback-Type Phase Voltage Compensation Strategy Based on Phase Current Reconstruction for ACIM Drives," *IEEE Transactions on Power Electronics*, pp. 5031–5043, 2014.
- [3] Sam-Young Kim, Wootaik Lee, Min-Sik Rho, and Seung-Yub Park, "Effective Dead-Time Compensation Using a Simple Vectorial Disturbance Estimator in PMSM Drives," *IEEE Transactions on Industrial Electronics*, pp. 1609–1614, 2010.
- [4] Y. Zhao, W. Qiao, and L. Wu, "Dead-Time Effect Analysis and Compensation for a Sliding-Mode Position Observer-Based Sensorless IPMSM Control System," *IEEE Transactions on Industry Applications*, pp. 2528–2535, 2015.
- [5] N. Bedetti, S. Calligaro, and R. Petrella, "Accurate modeling, compensation and self-commissioning of inverter voltage distortion for high-performance motor drives," in *2014 IEEE Applied Power Electronics Conference and Exposition - APEC 2014*. IEEE, 2014, pp. 1550–1557.
- [6] T. Morohoshi, N. Hoshi, and J. Haruna, "Dead-time compensation of adjustable dead-time controlled three-phase resonant snubber inverter for induction motor drive application," in *2013 International Conference on Electrical Machines and Systems (ICEMS)*, 2013, pp. 1766–1770.
- [7] Z. Zhang and L. Xu, "Dead-Time Compensation of Inverters Considering Snubber and Parasitic Capacitance," *IEEE Transactions on Power Electronics*, pp. 3179–3187, 2014.
- [8] H. S. Che, E. Levi, M. Jones, W.-P. Hew, and N. A. Rahim, "Current Control Methods for an Asymmetrical Six-Phase Induction Motor Drive," *IEEE Transactions on Power Electronics*, vol. 29, no. 1, pp. 407–417, jan 2014.
- [9] G. Pellegrino, R. I. Bojoi, P. Guglielmi, and F. Cupertino, "Accurate Inverter Error Compensation and Related Self-Commissioning Scheme in Sensorless Induction Motor Drives," *IEEE Transactions on Industry Applications*, vol. 46, no. 5, pp. 1970–1978, 2010.
- [10] E. Levi, R. Bojoi, F. Profumo, H. Toliyat, and S. Williamson, "Multi-phase induction motor drives - a technology status review," *IET Electric Power Applications*, vol. 1, no. 4, p. 489, 2007.
- [11] M. Kozovsky, P. Blaha, and P. Václavěk, "Verification of nine-phase PMSM model in d-q coordinates with mutual couplings," in *2016 6th IEEE International Conference on Control System, Computing and Engineering (ICCSCE)*. IEEE, 2016, pp. 73–78.

TABLE III
PARAMETERS OF EXPERIMENTAL MOTOR

Name	Symbol	Value	Unit
DC voltage	U_{DC}	55	V
Maximum motor current	i_{max}	6	A
Back-EMF constant	λ_m	0.00989	Vs/rad
Stator resistance	R_s	0.45	Ω
d-axis inductance	L_d	1.84	mH
q-axis inductance	L_q	1.98	mH
d-axis mutual inductance	M_d	75	μ H
q-axis mutual inductance	M_q	163	μ H
Nominal speed	ω_N	950	rpm
Maximum speed	ω_{max}	3500	rpm
Nominal power	P_N	270	W
Number of pole pairs	p_p	21	-