

Comparative Study on Harmonic Current Suppression of Dual three-phase PMSM Based on LMS Adaptive Linear Neuron and Resonant Controller

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Abstract—In this paper, the effects of the harmonic current suppression for dual three-phase permanent magnet synchronous motor (DTP-PMSM) by adaptive linear neuron (ALN) controller and resonant controller are systematically investigated. Although the ALN controller and the resonant controller have been extensively employed for the suppression of harmonic currents in the machine drive system, their suppression effect is not comparatively studied and the relationship between them is not revealed. The study shows that the transfer function of the least-mean-square (LMS)-based ALN controller in the frequency domain is equivalent to an ideal resonant controller, therefore, the ALN controller can be regarded as a resonant controller with ideal discretization method and has no resonant pole displacement and phase deviation to the desired resonant frequency. Compared with the conventional resonant controller with discretization methods, such as Tustin transformation and Tustin with pre-warping transformation, the ALN controller exhibits better harmonic suppression performance. Moreover, the design principle of the learning factor in the ALN controller is proposed by referring to the resonant gain in the conventional resonant controller. Finally, the effects of harmonic current suppression for DTP-PMSM by the ALN controller and the resonant controller are validated and compared by experiments.

Keywords—Adaptive linear neuron (ALN) controller, resonant controller, current harmonics suppression, dual three-phase PMSM (DTP-PMSM)

I. INTRODUCTION

In recent years, the multi-phase machine drive system has been widely studied due to its extra superiorities. For example, it can realize higher power applications with the same low-voltage switching devices and the lowest order of torque ripple frequency is higher. Moreover, its control degree of freedom and reliability is higher than that of its counterpart single three-phase machine [1]–[6].

Dual three-phase permanent magnet synchronous motor (DTP-PMSM) is an outstanding candidate for multi-phase machines. However, due to the small harmonic impedance of DTP-PMSM, there are large current harmonics even if the small voltage distortion [7], [8]. Those current harmonics will result in increased total harmonic distortion (THD).

Harmonic currents can be suppressed by the optimization of the machine's electromagnetic design. For example, the magnetomotive force of harmonics can be reduced by skew-slot, rotor-skewing, and pole shaping [9],

[10]. However, the 5th and 7th harmonics in the flux-linkage are inevitable, meanwhile, there are also 5th and 7th harmonic voltages resulting from the inverter non-linearity [11]. Therefore, the harmonic currents in the DTP-PMSM cannot be eliminated simply by the conventional PI controller due to limited bandwidth.

To suppress the harmonic currents in the machine drive system, the resonant controller is employed extensively in academic and industrial applications [11]–[14] for its infinite gain at the resonant frequency. In [11], the resonant controllers are employed for current unbalance compensation caused by system asymmetry and harmonic currents suppression caused by inverter nonlinearity. In [12], a generic decomposition model of the DTP-PMSM considering asymmetric impedances is established and the same strategy as [11]. In [13], the resonant controller is designed and optimized directly in the discrete domain to broaden the stability, but it faces the trade-off between tracking the target value and resisting system interference. In [14], the characteristics of the resonant controller using different discretization methods are evaluated, which may have a significant steady-state error due to resonant pole displacement at high frequencies.

As an alternative to the resonant controller, the adaptive linear neuron (ALN) controller based on the least-mean-square (LMS) algorithm can be adopted to suppress the current harmonics in the three-phase PMSM drives, which generates the corresponding harmonic voltage compensation according to the harmonic current of a specific order [15]–[17]. It has a simple structure, a small amount of calculation, and is easy to be implemented [18], [19]. In the ALN controller, the selection of the adaptive learning factor is very important. On the one hand, when the adaptive learning factor is chosen to be small, the learning rate is slow. On the other hand, when the adaptive learning factor is chosen to be large, it may cause the instability of the controller. However, the selection method of the adaptive learning factor is not given in [15]–[17], and the harmonic current effect might not be as expected if a decent adaptive learning factor is not applied.

Although the resonant controller and ALN controller have a satisfactory effect on the harmonic current suppression, the differences between them have not been analyzed. In this paper, the ALN controller and resonant controller are systematically investigated and it is found that they have an inherent relationship. Firstly, the mathematical model and

vector space decomposition (VSD) control of DTP-PMSM are introduced in Section II. After the brief introduction of the ALN controller and resonant controller, the relationship between them is analyzed in detail in Section III. Finally, the harmonic suppression effects of the two methods are compared on the drive platform of the prototype DTP-PMSM.

II. DTP-PMSM MODEL AND CURRENT HARMONICS

A. DTP-PMSM Model

The winding phasor diagram and circuit topology of the DTP-PMSM drive system is shown in Fig. 1. The DTP-PMSM consists of two sets of three-phase windings with two isolated neutral points n_1 and n_2 , the first set is called phase-ABC and the second set is called phase-XYZ. The phase shift electrical angle between two sets of single three-phase windings is 30° . The DTP two-level voltage source inverter (VSI) adopts a common DC bus with voltage U_{dc} for power supply.

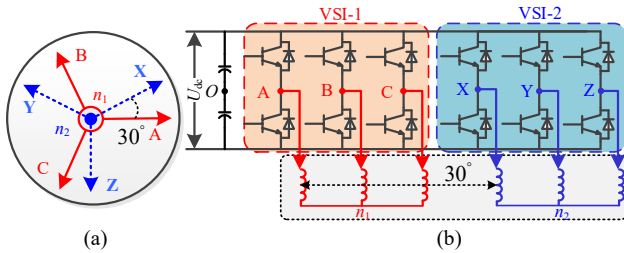


Fig. 1. DTP-PMSM drive system. (a) DTP-PMSM. (b) DTP two-level VSI.

According to the well-known VSD control theory [20], [21], the variables of DTP-IPMSM can be converted into three mutually orthogonal sub-planes [11]. If the zero-sequence components are not taken into account, the VSD transformation matrix T_{VSD} can be expressed as

$$T_{VSD} = \frac{1}{3} \begin{bmatrix} 1 & -1/2 & -1/2 & \sqrt{3}/2 & -\sqrt{3}/2 & 0 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 & 1/2 & 1/2 & -1 \\ 1 & -1/2 & -1/2 & -\sqrt{3}/2 & \sqrt{3}/2 & 0 \\ 0 & -\sqrt{3}/2 & \sqrt{3}/2 & 1/2 & 1/2 & -1 \end{bmatrix} \quad (1)$$

The variables in the $\alpha\beta$ -frame can be transformed into the dq -frame by the conventional Park matrix T_{dq} , and the variables in the $z1z2$ -frame can be transformed into the dqz -frame by the following matrix:

$$T_{dqz} = \begin{bmatrix} -\cos\theta_e & \sin\theta_e \\ \sin\theta_e & \cos\theta_e \end{bmatrix} \quad (2)$$

where θ_e is the electrical rotor position angle. Neglecting non-sinusoidal BEMF and cross-magnetic saturation effects, the voltage equations of DTP-IPMSM in dq - and dqz -frames can be modeled as

$$\begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} L_d p + R_s & -\omega_e L_q \\ \omega_e L_d & L_q p + R_s \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_f \end{bmatrix} \quad (3)$$

$$\begin{bmatrix} u_{dz} \\ u_{qz} \end{bmatrix} = \begin{bmatrix} L_{dz} p + R_s & -\omega_e L_{qz} \\ \omega_e L_{dz} & L_{qz} p + R_s \end{bmatrix} \begin{bmatrix} i_{dz} \\ i_{qz} \end{bmatrix} \quad (4)$$

where p is the differential operator; u , i , and L are the voltage, current, and inductance of the dq -axis and dqz -axis,

respectively; R_s is the stator resistance, ψ_f is the amplitude of fundamental PM flux linkage. ω_e is the electrical angular velocity. According to (3) and (4), the variables of the dq - and the dqz -frames are completely decoupled.

B. DTP-PMSM Current Harmonics

In the DTP-PMSM drive system, the harmonic currents are mainly caused by the inverter non-linearity, non-sinusoidal back-EMF of the DTP-PMSM, etc. If only the dominant 5th and 7th current harmonics are considered, the harmonic currents can be expressed as

$$i_h(\theta_e) = \begin{bmatrix} i_A \\ i_B \\ i_C \\ i_X \\ i_Y \\ i_Z \end{bmatrix} = \sum_{n=5,7} i_n \begin{bmatrix} \cos(n\theta_e + \varphi_n) \\ \cos(n(\theta_e - \frac{2\pi}{3}) + \varphi_n) \\ \cos(n(\theta_e + \frac{2\pi}{3}) + \varphi_n) \\ \cos(n(\theta_e - \frac{\pi}{6}) + \varphi_n) \\ \cos(n(\theta_e - \frac{5\pi}{6}) + \varphi_n) \\ \cos(n(\theta_e + \frac{\pi}{2}) + \varphi_n) \end{bmatrix} \quad (5)$$

where i_n is the amplitude of current harmonics, φ_n is the phase offset. The 5th and 7th current harmonics can be mapped into the dqz -frame by applying transformation matrices T_{VSD} , which can be expressed as

$$\begin{bmatrix} i_{dz} \\ i_{qz} \end{bmatrix} = \begin{bmatrix} -i_5 \cos(\varphi_5 + 6\theta_e) - i_7 \cos(\varphi_7 + 6\theta_e) \\ i_5 \sin(\varphi_5 + 6\theta_e) - i_7 \sin(\varphi_7 + 6\theta_e) \end{bmatrix} \quad (6)$$

According to (6), the 5th and 7th current harmonics are transformed into 6th harmonics in the dqz -frame, therefore, by suppressing the 6th harmonic currents in the dqz -frame, the 5th and 7th current harmonics in the ABC-XYZ-frame can be suppressed.

III. COMPARATIVE STUDY ON HARMONIC SUPPRESSION EFFECT OF DTP-PMSM

As detailed in Section II, the 5th and 7th harmonic currents are mapped to the 6th harmonic current in the dqz -frame, and the current harmonic suppression of DTP-PMSM is based on the 6th harmonic current suppression in the dqz -frame.

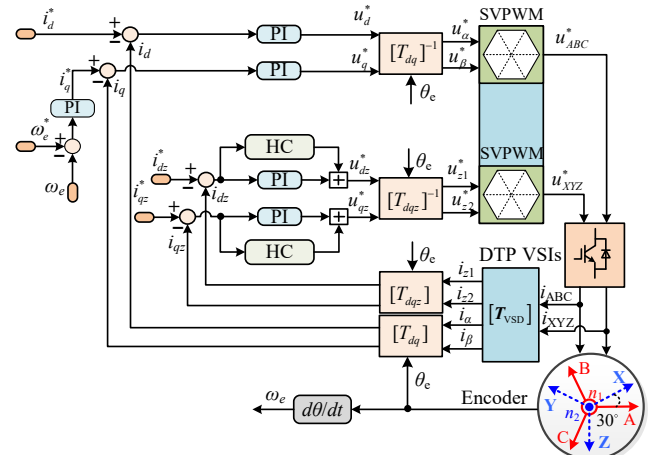


Fig. 2. Block Diagram of current harmonic suppression for DTP-PMSM.

The current control method of the DTP-PMSM can be illustrated in Fig. 2. For simplicity, the current control method with $i_d^* = 0$ is employed in the current loop of the dq -frame, and the pulse width modulation (PWM) strategy adopts the dual space vector PWM (SVPWM) strategy, where the conventional SVPWM strategy is applied to each single three-phase VSI individually. The current harmonics are suppressed by the harmonic controller (HC) superimposed on the conventional PI regulator. The HC can be ALN controller or resonant controller, which will be introduced in this Section.

A. Resonant Controller

The ideal resonant controller with gain K_r in the s -domain is [22]

$$G_R(s) = 2K_r \frac{s}{s^2 + (\omega_0)^2} \quad (7)$$

In the digital implementation of the resonant controller, the advantages and disadvantages of various discretization methods have been detailed in [14]. If the resonant controller is discretized into the z -domain by the conventional Tustin transformation, it can be expressed as

$$G_{R1}(z) = G_R(s) \Big|_{s=\frac{2}{T_s} \frac{z-1}{z+1}} = \frac{4T_s K_r z^2 - 4T_s K_r}{z^2 (T_s^2 \omega_0^2 + 4) + z(2T_s^2 \omega_0^2 - 8) + T_s^2 \omega_0^2 + 4} \quad (8)$$

If the resonant controller is discretized into the z -domain by the conventional Tustin transformation with pre-warping, it can be expressed as

$$G_{R2}(z) = G_R(s) \Big|_{s=\frac{\omega_0}{\tan(\omega_0 T_s/2)} \frac{z-1}{z+1}} = \frac{K_r \sin(\omega_0 T_s)}{\omega_0} \frac{z^2 - 1}{z^2 - 2z \cos(\omega_0 T_s) + 1} \quad (9)$$

B. ALN Controller

The structure of the LMS-based ALN controller can be illustrated in Fig. 3 [15]. In Fig. 3, the reference signals x_1 and x_2 are sinusoidal references, μ is the adaptive learning factor, k is the time index, and T_s is the sampling period in the discrete system, w_1 and w_2 are the weighting factors. d is the system input signal, y is the output signal, ε is the error signal, ω_0 is the frequency of the signal to be learned. The reference signals $x_1(k)$ and $x_2(k)$ can be expressed as

$$x_1(k) = \sin(\omega_0 k T_s), x_2(k) = \cos(\omega_0 k T_s) \quad (10)$$

The output signal $y(k)$ is

$$y(k) = x_1(k)w_1(k) + x_2(k)w_2(k) \quad (11)$$

The weighting factors $w_1(k)$ and $w_2(k)$ can be updated using the the following LMS algorithm

$$\begin{aligned} w_1(k+1) &= w_1(k) + 2\mu \cdot x_1(k)\varepsilon(k), \\ w_2(k+1) &= w_2(k) + 2\mu \cdot x_2(k)\varepsilon(k) \end{aligned} \quad (12)$$

The transfer function from ε to y can be expressed as

$$G_{ALN}(z) = \frac{y(z)}{\varepsilon(z)} = \frac{2\mu(z \cos(\omega_0 T_s) - 1)}{z^2 - 2z \cos(\omega_0 T_s) + 1} \quad (13)$$

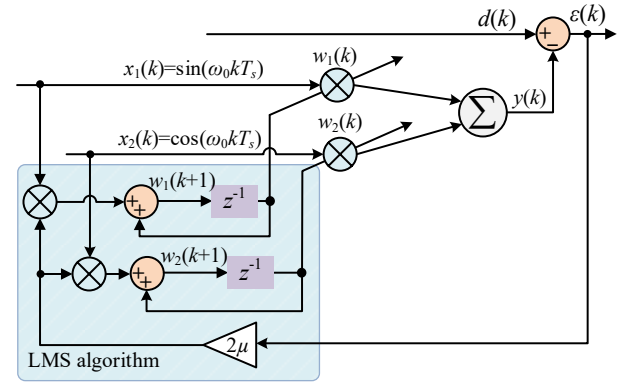


Fig. 3. The structure of the ALN controller based on LMS algorithm.

C. Relationship Between ALN Controller Resonant Controller

According to Fig. 3, the value of $y(k)$ at the k th sampling point is

$$y(k) = 2\mu \left[\sin(\omega_0 k T_s) \sum_{m=0}^k (\varepsilon(m) \sin(\omega_0 m T_s)) + \cos(\omega_0 k T_s) \sum_{m=0}^k (\varepsilon(m) \cos(\omega_0 m T_s)) \right] \quad (14)$$

(14) can be rewritten in the time-domain as

$$y(t) = \frac{2\mu}{T_s} \left[\sin(\omega_0 t) \int_0^t \varepsilon(\tau) \sin(\omega_0 \tau) d\tau + \cos(\omega_0 t) \int_0^t \varepsilon(\tau) \cos(\omega_0 \tau) d\tau \right] \quad (15)$$

The second-order ordinary differential equation between $y(t)$ and $\varepsilon(t)$ can be expressed as

$$y''(t) + (\omega_0)^2 y(t) = \frac{2\mu}{T_s} \varepsilon'(t) \quad (16)$$

(16) is rewritten in the s -domain as

$$s^2 y(s) + (\omega_0)^2 y(s) = \frac{2\mu}{T_s} s \varepsilon(s) \quad (17)$$

The transfer function from $\varepsilon(s)$ to $y(s)$ is

$$G_{ALN}(s) = \frac{y(s)}{\varepsilon(s)} = \frac{2\mu}{T_s} \frac{s}{s^2 + (\omega_0)^2} \quad (18)$$

Therefore, the ALN controller actually is equivalent to the resonant controller in the s -domain with a gain of μ/T_s . In the ALN controller, the selection of μ is very important. However, the design rules of selection of μ is not given in [15]–[17]. After the relationship between ALN controller and resonant controller is revealed, according to (18) and (7), the value of μ can be determined by referring to the integral gain K_r of the resonant controller.

$$\mu_i = T_s K_r \quad (19)$$

The bode diagrams of $G_{ALN}(z)$, $G_{R1}(z)$, $G_{R2}(z)$ are shown in Fig. 4 and Fig. 5, where the sampling period $T_s = 1e^{-4}$ s, and $\mu = 5e^{-3}$, $K_r = \mu/T_s$. The center resonant frequency $\omega_0 = 2\pi \times 300$ rad/s in Fig. 4, and the center resonant frequency $\omega_0 = 2\pi \times 600$ rad/s in Fig. 5. It can be seen that in Fig. 4, the center frequency of $G_{R1}(z)$ shifts left about 1 Hz; in Fig. 5, the center frequency of $G_{R1}(z)$ shifts left about 7 Hz. According to [14], a discrete-time implementation based on Tustin transformation produces a significant steady-state error due

to the resonant pole displacement, and this error increases with the increment of the sampling period and resonant frequency. However, this deviation can be negligible if the resonant frequency is low enough [14]. However, $G_{ALN}(z)$ and $G_{R2}(z)$ have no resonant pole displacement phenomena, and the gain of $G_{R2}(z)$ at the center frequency 600 Hz is slightly less than that of $G_{ALN}(z)$.

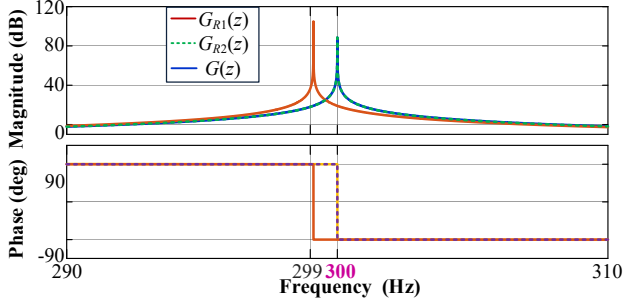


Fig. 4. Comparison of bode plots of $G_{R1}(z)$, $G_{R2}(z)$, $G_{ALN}(z)$ for $\omega_0=2\pi\times 300$ rad/s.

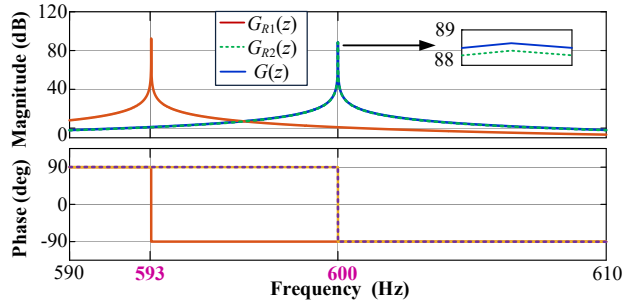


Fig. 5. Comparison of bode plots of $G_{R1}(z)$, $G_{R2}(z)$, $G_{ALN}(z)$ for $\omega_0=2\pi\times 600$ rad/s.

IV. EXPERIMENTAL RESULTS

A. Experimental Platform

To demonstrate the effect of harmonic current suppression of resonant controller and ALN controller, the DTP-PMSM test platform based on TI's DSP processor TMS320F28335 is constructed and shown in Fig. 6.

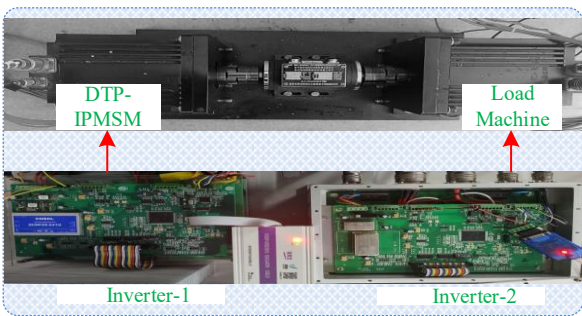


Fig. 6. Experimental setup for prototype DTP-PMSM drive system.

The inverter consists of two single three-phase VSIs with the same common DC bus. The operating frequency of the current loop is 10 kHz, which is the same as the PWM switching frequency, the dead time sets as $2\ \mu\text{s}$. The main parameters of the prototype DTP-PMSM are listed in TABLE I. For the convenience of discussion, the conventional resonant controller discretized by Tustin transformation, i.e., (8), it is designated as Method-I, the conventional resonant controller discretized by Tustin transformation with pre-

warping, i.e., (9), is designated as Method-II.

TABLE I
PARAMETERS OF PROTOTYPE DTP-PMSM

Parameters	Value
Resistance (Ω)	0.15
d -axis self-inductance in $\alpha\beta$ sub-plane (mH)	4.336
q -axis self-inductance in $\alpha\beta$ sub-plane (mH)	5.841
dz -axis self-inductance in z_1z_2 sub-plane (mH)	2.662
qz -axis self-inductance in z_1z_2 sub-plane (mH)	2.760
Flux linkage (Wb)	0.0785
Pole pairs	5
Rated current (A)	12
Rated speed (rpm)	600

B. Experimental Results

1) Comparison at the low-speed region

In this experiment, the speed reference sets to 600 r/min. The corresponding fundamental frequency is 50 Hz, and the corresponding frequency of current harmonics in the dqz -axis is 300 Hz. The load increases steadily until the root mean square current of the prototype DTP-PMSM increases to 12 A.

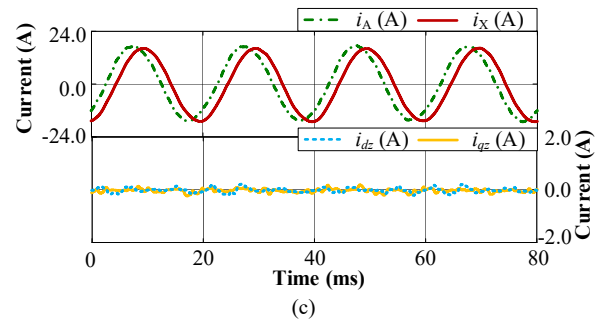
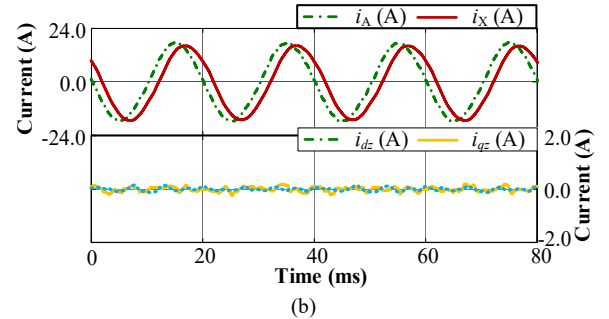
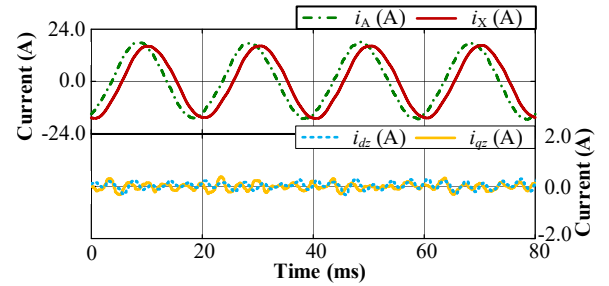


Fig. 7. Phase currents at low-speed operation. (a) Method-I. (b) Method-II. (c) ALN controller.

Fig. 7.(a)-(c) shows the phase currents and dqz -axis currents with Method-I, Method-II, and the ALN controller, respectively. From Fig. 7, the phase currents with these three methods are both very sinusoidal. However, the amplitude of the currents in the dqz -axis with Method-I is 0.17 A, the

amplitude of the currents in the dqz -axis with Method-II is 0.08 A, and the amplitude of the currents in the dqz -axis with ALN controller is 0.01 A.

The corresponding current harmonic contents and THD of Phase-A are shown in TABLE II. From TABLE II, in terms of the 6th harmonic current suppression effect in dqz -axis in low-speed region, Method-I is the worst, and Method-II is similar to ALN controller but slightly worse. Therefore, the ALN controller exhibits better performance than the conventional discretization methods in the low-speed region.

TABLE II
COMPARISON OF HARMONIC CONTENTS AND THD OF PHASE-A
CURRENT UNDER THE THREE METHODS

Order	Method-I	Method-II	ALN controller
5th	0.68%	0.51%	0.30%
7th	0.19%	0.41%	0.32%
THD	2.16%	2.24%	1.41%

2) Comparison at the high-speed region

In this experiment, the speed reference sets to 1200 r/min. The corresponding fundamental frequency is 100 Hz, and the corresponding frequency of current harmonics in the dqz -axis is 600 Hz. As the frequency is much higher than the rated frequency, the field weakening control is engaged for d -axis current reference generation [23]. The load increases steadily until the root mean square current of the prototype DTP-PMSM increases to 12 A.

Fig. 8(a)-(c) shows the phase currents and dqz -axis currents with Method-I, Method-II, and the ALN controller, respectively. From Fig. 8, as the harmonic components in the dqz -frame are 600 Hz and the corresponding carrier ratio is relatively low, the phase currents with Method-I deteriorate seriously at high-speed operation, which is due to the resonant pole displacement of the resonant controller. On the contrary, the phase currents in Fig. 8(b) with Method-II and Fig. 8(c) with the ALN controller are very sinusoidal, which is much better than the phase currents in Fig. 8(a). Among them, the amplitude of the currents in the dqz -axis with Method-I is 1.17 A, the amplitude of the currents in the dqz -axis with Method-II is 0.13 A, and the amplitude of the currents in the dqz -axis with ALN controller is 0.01 A.

The current harmonic contents and THD of Phase-A with Method-I, Method-II, and the ALN controller are detailed in TABLE III. It is worth noting that due to the gain attenuation of Method-II at the resonance point, the 5th and 7th harmonics and THD of Method-II are slightly higher than that of the ALN controller, so as the results shown in TABLE II. Therefore, the ALN controller exhibits even better performance than the Method-II.

It can be concluded from Fig. 7, Fig. 8, TABLE II, and TABLE II, the ALN controller exhibits even better performance than the Method-I and Method-II in the full-speed domain.

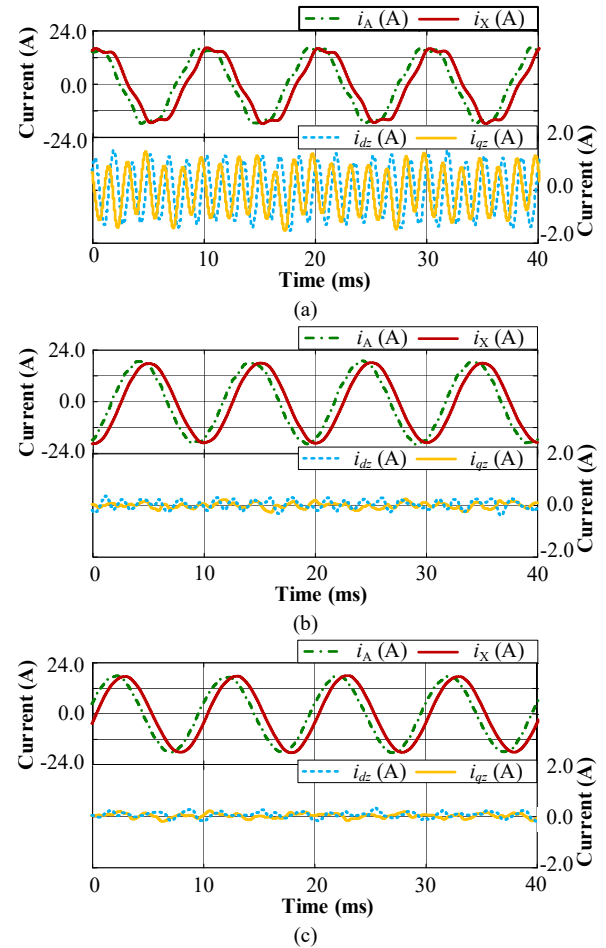


Fig. 8. Measured phase currents at high-speed operation. (a) Method-I. (b) Method-II. (c) The ALN controller.

TABLE III
COMPARISON OF HARMONIC CONTENTS AND THD OF PHASE-A UNDER
THE THREE METHODS

Order	Method-I	Method-II	ALN controller
5th	7.80%	0.41%	0.1%
7th	1.45%	1.22%	0.05%
THD	8.32%	2.28%	1.86%

V. CONCLUSION

In this paper, the harmonic current suppression effects of the resonant controller and ALN controller for DTP-PMSM are systematically investigated. As a candidate of the resonant controller for harmonic currents suppression, the ALN controller exhibits better performance than the conventional resonant controlled with discretization methods, such as Tustin transformation and Tustin with pre-warping transformation. The LMS-based ALN controller can be regarded as the resonant controller with the ideal discretization method and has no resonant poles displacement and phase deviation to the desired resonant frequencies. Moreover, the design principle of the learning factor in the ALN controller is proposed by referring to the resonant gain in the conventional resonant controller.

REFERENCES

- [1] L. Parsa and H. A. Toliyat, "Fault-tolerant interior-permanent-magnet machines for hybrid electric vehicle applications," *IEEE Trans. Veh. Technol.*, vol. 56, no. 4, pp. 1546-1552, 2007.

- [2] J. Dai, S. W. Nam, M. Pande, and G. Esmaeili, "Medium-voltage current-source converter drives for marine propulsion system using a dual-winding synchronous machine," *IEEE Trans. Ind. Appl.*, vol. 50, no. 6, pp. 3971-3976, 2014.
- [3] J. Xu, Y. Du, B. Zhang, H. Fang *et al.*, "Sensorless fault-tolerant control with phase delay compensation for aerospace FTPMSM drives with phase open-circuit and short-circuit faults," *IEEE Trans. Ind. Electron.*, vol. 68, no. 6, pp. 4576-4585, 2021.
- [4] L. de Lillo, L. Empringham, and P. W. Wheeler, "Multiphase power converter drive for fault-tolerant machine development in aerospace applications," *IEEE Trans. Ind. Electron.*, vol. 57, no. 2, pp. 575-583, 2010.
- [5] Z. Q. Zhu, S. Wang, B. Shao, L. Yan *et al.*, "Advances in dual-three-phase permanent magnet synchronous machines and control techniques," *Energies*, vol. 14, no. 22, 2021.
- [6] Y. Hu and X. Li, "Fault - tolerant control of dual three - phase PMSG wind turbine drive system with one channel of faulty rectifier," *IET Renew. Power Gener.*, vol. 14, no. 12, pp. 2135-2141, 2020.
- [7] Y. Xu, B. Zheng, G. Wang, H. Yan *et al.*, "Current harmonic suppression in dual three-phase permanent magnet synchronous machine with extended state observer," *IEEE Trans. Power Electron.*, vol. 35, no. 11, pp. 12166-12180, 2020.
- [8] J. Karttunen, S. Kallio, P. Peltoniemi, and P. Silventoinen, "Current harmonic compensation in dual three-phase PMSMs using a disturbance observer," *IEEE Trans. Ind. Electron.*, vol. 63, no. 1, pp. 583-594, 2016.
- [9] D. Wang, X. Wang, and S.-Y. Jung, "Cogging torque minimization and torque ripple suppression in surface-mounted permanent magnet synchronous machines using different magnet widths," *IEEE Trans. Magn.*, vol. 49, no. 5, pp. 2295-2298, 2013.
- [10] L. Guo, J. Xu, S. Wu, X. Xie *et al.*, "Analysis and design of dual three-phase fractional-slot permanent magnet motor with low space harmonic," *IEEE Trans. Magn.*, pp. 1-1, 2021.
- [11] H. Yashan, Z. Zi-Qiang, and L. Kan, "Current control for dual three-phase permanent magnet synchronous motors accounting for current unbalance and harmonics," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 2, no. 2, pp. 272-284, 2014.
- [12] J. Xu, M. Odavic, and Z. Q. Zhu, "A generalized decomposition model of dual three-phase permanent magnet synchronous machines considering asymmetric impedances and compensation capability," *IEEE Trans. Ind. Appl.*, vol. 57, no. 4, pp. 3763-3775, 2021.
- [13] M. Hu, W. Hua, W. Huang, and J. Meng, "Digital current control of an asymmetrical dual three-phase flux-switching permanent magnet machine," *IEEE Trans. Ind. Electron.*, vol. 67, no. 6, pp. 4281-4291, 2020.
- [14] A. G. Yepes, F. D. Freijedo, J. Doval-Gandoy, Ó. López *et al.*, "Effects of discretization methods on the performance of resonant controllers," *IEEE Trans. Power Electron.*, vol. 25, no. 7, pp. 1692-1712, 2010.
- [15] Z. Tang and B. Akin, "A new LMS algorithm based deadtime compensation method for PMSM FOC drives," *IEEE Trans. Ind. Appl.*, vol. 54, no. 6, pp. 6472-6484, 2018.
- [16] T. Qiu, X. Wen, and F. Zhao, "Adaptive-linear-neuron-based dead-time effects compensation scheme for PMSM drives," *IEEE Trans. Power Electron.*, vol. 31, no. 3, pp. 2530-2538, 2016.
- [17] L. Wang, Z. Q. Zhu, H. Bin, and L. M. Gong, "Current harmonics suppression strategy for PMSM with nonsinusoidal back-EMF based on adaptive linear neuron method," *IEEE Trans. Ind. Electron.*, vol. 67, no. 11, pp. 9164-9173, 2020.
- [18] Widrow, B., Glover, R. J. *et al.*, "Adaptive noise cancelling: Principles and applications," *Proceedings of the IEEE*, 1975.
- [19] Widrow, Bernard, Stearns, and S. D., *Adaptive signal processing*. Adaptive signal processing, 2008, pp. 316-324.
- [20] Y. Hu, Z. Q. Zhu, and M. Odavic, "Comparison of two-individual current control and vector space decomposition control for dual three-phase PMSM," *IEEE Trans. Ind. Appl.*, vol. 53, no. 5, pp. 4483-4492, 2017.
- [21] H. M. Eldeeb, A. S. Abdel-Khalik, and C. M. Hackl, "Dynamic modeling of dual three-phase IPMSM drives with different neutral configurations," *IEEE Trans. Ind. Electron.*, vol. 66, no. 1, pp. 141-151, 2019.
- [22] A. G. Yepes, F. D. Freijedo, Ó. Lopez, and J. Doval-Gandoy, "High-performance digital resonant controllers implemented with two integrators," *IEEE Trans. Power Electron.*, vol. 26, no. 2, pp. 563-576, 2011.
- [23] Y. Hu, Y. Li, X. Ma, X. Li *et al.*, "Flux-weakening control of dual three-phase PMSM based on vector space decomposition control," *IEEE Trans. Power Electron.*, vol. 36, no. 7, pp. 8428-8438, 2021.