An Online VSI Error Parameter Identification Method for Multiphase IM With Non-Sinusoidal Power Supply

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ABSTRACT Voltage-source-inverter (VSI) nonlinearity compensation is an important scheme for improve motor performance. The traditional off-line method is cumbersome and the compensation effect is poor because it does not simulate the real operation of the motor. To solve this problem, an on-line identification method for compensating multiphase machine based on non-sinusoidal power supply is proposed in this paper. First, a full order air-gap flux observer on harmonic plane was established. Then the voltage error is estimated based on a model reference adaptive system (MRAS) and compensated to the given voltage. In the process of identification, the adaptive law of dead-time voltage is designed, and a feedback gain matrix satisfying Lyapunov stability is also designed. Finally, the feasibility and the effectiveness for harmonic suppression of the proposed method are verified on an experimental platform of a seven-phase induction motor platform.

INDEX TERMS Multiphase induction motor, non-sinusoidal power supply, MRAS, VSI nonlinearity, air-gap flux orientation control (AFOC).

NOMENCLATURE

\[ \theta \] Rotation angle of the magnetic fields.
\[ \omega \] Electrical angular velocity.
\[ \psi \] Flux linkage.
\[ B_g \] Amplitude of flux density.
\[ R, L \] Resistance and inductance.
\[ \alpha \] Reciprocal of rotor time constant.
\[ \rho \] Number of pole pair.
\[ \psi_m \] Air-gap flux.

Subscripts

\[ s, r \] Stator and rotor component.
\[ s_l \] Slip component.
\[ \sigma \] Leakage component.

\[ d, q \] The axes of a rotation frame of reference by rotor flux orientation control (RFOC).
\[ x, y \] The axes of a rotation frame of reference by air-gap flux orientation control (AFOC).

I. INTRODUCTION

Multiphase motor has become a hot research area with the development of modern industrial application. At present, multiphase motor has been used in offshore wind power generation, rail transportation and ship electric propulsion systems and so on [1], [2], [3]. The main reason is that, compared with three-phase motors, multiphase motors have the advantages of high reliability, large power density, and more controllable degrees of freedom [4], [5], [6]. At the same time, with the increase of phase number, more inverters are needed to drive the motors.
The VSI nonlinearity problem is always an indispensable part of motor drive control. It is mainly caused by the dead time delay and the voltage drop of the insulated gate bipolar transistor (IGBT). This phenomenon results in that the phase voltage supplied to the motor differs from the given value. Due to the existence of the voltage distortion, the control performance of motor deteriorates, such as: distortional parameter identification, harmonic current increasing, intensified torque chattering and so on [7], [8], [9]. These adverse effects are particularly significant when the motor is operating at low speed. Therefore, VSI nonlinearity compensation plays a great important part in motor control.

As for multiphase machine studied in this paper, its voltage fluctuations at neutral points are affected by seven-phase currents. It leads to the presence of higher harmonic voltages rather than (6k ± 1)th order harmonic [10]. Third harmonic current injection is widely used in multiphase motor control to improve torque performance [11], [12], [13], [14]. An accurate flux observer is of great help to multiphase motors [14], [15]. However, voltage distortion has non-ignorable influence on flux observer at low speed. Therefore, the given voltage is badly in need of compensation for seven-phase.

At present, the methods of VSI nonlinear compensation can be divided into two categories: off-line measurement compensation and online identification compensation. The off-line compensation method consists of look-up-table (LUT) [16] and fitting curve [8], [17]. This kind of skills is easy to understand, but cumbersome in operation and not universal. When the bus voltage or dead time was changed, the off-line compensation method needs to be re-measured. Online compensation methods also mainly include two kinds. One of them is approximates off-line fitting curve compensation [18], [19], [20]. It can adjust the parameters of the fitting function online to achieve that actual voltage is equal to given voltage. However, this method did not take into account voltage fluctuations at neutral points. The other approach is to treat the VSI nonlinearity as a constant times the polarity of the current under synchronous coordinate frame [21], [22], [23]. Then, the constant can be estimated by Low-Pass Filter, Kalman-Filter and current injection [24], [25], [26]. The method proposed in [24] can only used in PMSM. The estimated voltage error is fed back only to the observer. It improves the accuracy of the observer, but it has no suppression effect on harmonic currents. Though [26] reduced the harmonic current, additional current injection will increase motor losses.

In this paper, a new VSI nonlinearity parameter identification and compensation algorithm based on non-sinusoidal power supply is proposed. Due to the harmonic injection will provide the extra torque on harmonic plane for the prototype, the advantages of this algorithm are particularly obvious in the multiphase motor we studied. At the same time, by using the degree of freedom of the harmonic plane, the parameters of voltage error can be identified on the harmonic plane with a MRAS and compensated to the given voltage. This identification method is mainly used to estimate speed and other motor parameters [15], [27], [28], [29]. A new adaptive law is designed for inverter voltage error identification. If the feedback gain matrix is not considered during the design process, the adaptive law cannot meet the stability requirement [30]. Therefore a feedback gain matrix satisfying Lyapunov stability is designed in this paper.

The outline of this paper is as follows. In section II, we explained how the nonlinearity of VSI produce and establish the voltage error model of traditional method. Then, we analyzed the shortcomings of the traditional model and establish the VSI model under synchronous coordinate frame. In section III, we built a full-order flux observer for third harmonic plane based on AFOC and identified the voltage error by MRAS method. Moreover, in the process of error voltage adaptive law calculation, we designed a set of feedback gain matrix satisfying Lyapunov’s stability. Next, the algorithm is validated on a self-designed seven-phase induction motor experimental platform. Finally, the conclusion is drawn.

![FIGURE 1. Seven-phase IM drive system.](image-url)
of upper and lower IGBT in phase A. After the introduction of dead time, \( S_1 \) should be opened later and shut down in advance with \( \frac{t_{\text{dead}}}{2} \). \( S_5 \) need to be shut down later and opened in advance at the same sample time, as shown in Fig. 2 (b). In practice, the inverter on and off time needs to be taken into account. When \( i_{\text{sa}} > 0 \), the potential difference waveform between point A and point \( N \) in a pulse width modulation (PWM) cycle is shown in Fig. 2 (c). Fig. 2 (d) shows the voltage waveform under \( i_{\text{sa}} < 0 \).

After considering dead-time effect and IGBT on/off time effect, A-phase voltage is expressed as follows:

\[
u_{\text{ua}} = (V_{dc} - V_{sat} + V_d) \left( T_a - \frac{1}{2} \right) - \frac{1}{2} (V_{sat} + V_d) \text{sgn}(i_{\text{sa}})\]

where: \( T_a = T_a^* - \text{sgn}(i_{\text{sa}}) (t_{\text{dead}} + t_{\text{on}} - t_{\text{off}}) \), \( T_s \) is the sample time.

**B. OFFLINE COMPENSATION METHOD**

Usually, a manual measurement of single phase voltage error method is used for feedforward compensating inverter voltage [8], [17]. The method is simple and easy to operate and can meet most industrial needs. The error voltage can be expressed as:

\[
u_{\text{aerr}} = u_{\text{an}}^* - u_{\text{an}}\]

(4)

The actual voltage of the phase A consists of the resistance voltage, the inductance voltage and the back electromotive force (EMF). The inductance voltage and the back EMF are both zero, because the rotor of the motor is stationary and the injecting current is direct current while using manual measurement method. So the actual voltage of phase A is equal to the voltage drop across the resistance. It is worth noting that since the other phases have no current, the potential at point \( N \) is equal to the potential at point \( O \) and both are 0. Thus, \( u_{\text{aerr}} \) can be expressed as:

\[
u_{\text{aerr}} = u_{\text{an}}^* - i_{\text{sa}} R_s\]

(5)

Fig. 3 shows the voltammetry curve after injecting direct current into a single phase. The other curve is the voltammetry curve of error voltage derived from (5). Generally, there are two kinds of off-line feed-forward compensation methods: LUT and curve fitting.

**III. FLUX MODEL OF THIRD HARMONIC PLANE**

For the multiphase induction motors, the third harmonic current injection can improve the distribution of the motor air gap flux, so that it could achieve increment of the torque density [31]. Furthermore, there are two mutually independent coordinate planes are formed after PARK transformation with third harmonic current injection. Fortunately, the extra plane can also be used to identity voltage error.

**A. HARMONIC CURRENT INJECTION SCHEME**

According to the principle of linear superposition, the airgap flux density can be expressed as:

\[
B_p(t, \theta_{s1}, \theta_{s3}) = k_1 B_1 \left[ \sin(\theta_{s1}) + k_3 \sin(\theta_{s3}) \right]
\]

(6)

where \( k_1 = \frac{2}{\sqrt{3}} \) and \( k_3 = \frac{1}{5} \) are the optimized proportionality coefficients using genetic algorithm [14], [32].

In order to obtain better flux distribution, this paper adopt a non-sinusoidal power supply control technique which is based on airgap flux orientation [13]. The rms current is reduced by 10.5% when the motor operates at rated load. The given currents on the third harmonic plane are shown in (7a) and (7b):

\[
i_{s3x} = \frac{k_1 k_3 L_{m1}}{3} \sqrt{\frac{l_{sd1}^2 + \left( 1 - \frac{l_{ml1}}{L_{r1}} \right)^2 l_{sq1}^2}{L_{m3} (X + Y)}} Y
\]

(7a)

\[
i_{s3y} = \frac{k_1 k_3 L_{m1}}{3} \sqrt{\frac{l_{sd1}^2 + \left( 1 - \frac{l_{ml1}}{L_{r1}} \right)^2 l_{sq1}^2}{L_{m3} (X + Y)}}
\]

(7b)

Where:

\[
X = -\frac{3 l_{sd1} L_{r3} \sigma_3}{R_3}
\]

\[
Y = \frac{9 l_{sd1}^2 L_{r3} \sigma_3 + R_3^2}{3 l_{sd1} L_{m3} R_3}
\]
B. AIR-GAP FLUX MODEL ON THIRD HARMONIC PLANE

Because of the injection of the third harmonic current, the third harmonic plane is formed which is independent of the fundamental plane. As the harmonic plane is based on airgap flux field orientation, \( x - y \) reference frame, its state equation variables should be the currents and the airgap flux, which is shown in:

\[
\dot{x}_3 = M_3 x_3 + N_3 u_3 \\
y_3 = C_3 x_3
\]

where

\[
x_3 = \begin{bmatrix}
i_{sx3} & i_{sy3} & \psi_{ms3} & \psi_{ms3}
\end{bmatrix}^T
\]

\[
u_3 = \begin{bmatrix}
u_{sx3} & \nu_{sy3}
\end{bmatrix}^T,
\]

\[
N_3 = \begin{bmatrix}
1 & 0 & 0 & 0
0 & 1 & 0 & 0
0 & 0 & 1 & 0
0 & 0 & 0 & 1
\end{bmatrix}
\]

and \( \omega_{sx3} \) is the synchronous angular velocity of the air-gap flux.

IV. ONLINE COMPENSATION METHOD USING THIRD HARMONIC CURRENT INJECTION

In this section, the extra degrees of freedom on third harmonic plane was used to identify the error of voltage and feedback to the given voltage on synchronous coordinate frame. The whole control block diagram of proposed online method is shown in Fig.4. This method uses MRAS on the third harmonic plane to estimate error voltage. Prior to this, the model VSI nonlinearity on synchronous coordinate frame needs to be rebuilt.

A. VSI NONLINEARITY MODEL ON SYNCHRONOUS COORDINATE FRAME

The offline measurement method is simple, but when the bus voltage or dead zone parameters need to be adjusted, the compensation curve needs to be manually measured off-line again. Moreover, manual measurement method injects direct current into only one phase and assumes that the potential at neutral point \( N \), in Fig.1, is the same as point \( O \). However, when the motor is running normally, the \( N \) point potential is affected by the current of each phase [23] and it is not equal to 0.

As shown in Fig.1, after considering the voltage fluctuation of \( N \) points, \( u_{no} \neq 0 \), the actual phase voltage is:

\[
u_{an} = u_{ao} - u_{no}
\]

The phase voltage of the other phases is expressed in the same way as phase A, which is expressed as following:

\[
\begin{bmatrix}
u_{bn} \\
u_{cn} \\
u_{dn}
\end{bmatrix} = \begin{bmatrix}
u_{bo} \\
u_{co} \\
u_{fo}
\end{bmatrix} - u_{no}
\]

In the case of symmetrical and stable operation of the motor, the voltage fluctuation \( u_{no} \) can be derived from (10) and (11).

\[
u_{no} = \frac{u_{ao} + u_{bo} + u_{co} + u_{do} + u_{eo} + u_{fo} + u_{go}}{7}
\]

\[
u_{no} = (V_{dc} - V_{sat} + V_d) \left( \frac{1}{7T_s} \sum_{\chi=a}^{g} \text{sgn}(i_{\chi}) \right)
\]

Substitute (3) into (12), then (13) can be obtain. According to the given phase voltage: \( u_{an} = V_{dc} \left( \frac{T_s}{T_s} - \frac{1}{7} \right) \) and by combining (5), (10) and (13), the voltage error of phase A inverter can be obtained as (14).

\[
u_{uerr} = \left( \frac{(V_{dc} - V_{sat} + V_d) \left( t_{dead} + t_{on} - t_{off} \right) + V_{sat} + V_d}{2} \right)
\]

\[
\times \left( \frac{\text{sgn}(i_{sa}) - \frac{1}{7} \sum_{\chi=a}^{g} \text{sgn}(i_{\chi})}{V_{dc} - \nu_{ian}} \right) + \frac{(V_{sat} + V_d)}{2}
\]

Due to \( V_{dc} \) is much larger than \( (V_{sat} - V_d) \), the second term in (14) can be neglected normally [25]. So (14) can be simplified as:

\[
u_{uerr} = \frac{1}{7} (6\text{sgn}(i_{sa}) - \text{sgn}(i_{sb}) - \text{sgn}(i_{sc}) - \text{sgn}(i_{sa}) - \text{sgn}(i_{sc}) - \text{sgn}(i_{sg}))
\]

where:

\[
u_{err} = \frac{1}{T_s} \left( \frac{(V_{dc} - V_{sat} + V_d) \left( t_{dead} + t_{on} - t_{off} \right) + V_{sat} + V_d}{2} \right)
\]
It can be seen that $u_{err}$ is mainly influenced by $V_{dc}$ and $i_{dead}$.

According to (15), when seven-phase sinusoidal current is injected into the seven-phase induction motor, the voltage error generated by the inverter of phase $A$ is shown in the Fig.5.

Fig.6 illustrates that VSI nonlinearity will bring $(14k \pm n)$th order harmonic voltage to the system, where $k$ is a non-negative integer, and $n = 1, 3, 5$ [10]. Correspondently, these harmonic voltages produce harmonic currents. Based on the characteristics of the seven-phase motor, the third harmonic and fifth harmonic currents are controllable. So that ninth harmonics, eleventh harmonics and thirteenth harmonics become the main suppression object of inverter compensation algorithm.

The inverter nonlinearity model of seven-phase platform can be extended from (15).

$$u_{err} = u_{err}C_7\text{sgn}(i_{sub})$$

where:

$$u_{err} = \begin{bmatrix} u_{aerr} & \cdots & u_{gerr} \end{bmatrix}^T, \quad i_{sub} = \begin{bmatrix} i_{sa} & \cdots & i_{sg} \end{bmatrix}^T,$$

$$C_7 = \frac{1}{7} \begin{bmatrix} 6 & -1 & -1 & -1 & -1 & -1 & -1 \\ -1 & 6 & -1 & -1 & -1 & -1 & -1 \\ -1 & -1 & 6 & -1 & -1 & -1 & -1 \\ -1 & -1 & -1 & 6 & -1 & -1 & -1 \\ -1 & -1 & -1 & -1 & 6 & -1 & -1 \\ -1 & -1 & -1 & -1 & -1 & 6 & -1 \\ -1 & -1 & -1 & -1 & -1 & -1 & 6 \end{bmatrix}$$

Usually, IMs is analyzed under synchronous rotation frame. Therefore, the VSI nonlinearity model can be calculated from (16).

$$\begin{bmatrix} u_{ad1err} \\ u_{sq1err} \\ u_{x33err} \\ u_{y33err} \end{bmatrix} = u_{err} \cdot \begin{bmatrix} f_d1 \\ f_q1 \\ f_{x3} \\ f_{y3} \end{bmatrix},$$

where:

$$\begin{bmatrix} f_{d1} \\ f_{q1} \\ f_{x3} \\ f_{y3} \end{bmatrix} = T(\theta_{r1}, \theta_{m3}) \cdot \begin{bmatrix} \text{sgn}(i_{sa}) \\ \vdots \\ \text{sgn}(i_{sg}) \end{bmatrix}$$

and $T(\theta_{r1}, \theta_{m3})$ is represents the Park transformation. It is noticed that fifth harmonic variables are not listed, because fifth harmonic current was controlled to zero.

**B. DESIGN OF ADAPTIVE LAW**

Once VSI nonlinearity is taken into consideration, the actual machine model in rotating reference frame on third harmonic plane can be written as:

$$\dot{x}_3 = M_3x_3 + N_3 \left( u_3^* - u_{3err} \right)$$
where: \( \mathbf{u}_{3}^{e} = \begin{bmatrix} u_{sx3}^{e} & u_{sy3}^{e} \end{bmatrix}^T \), are the given voltage of xy axis respectively on the harmonic plane, \( \mathbf{u}_{3err} = \begin{bmatrix} u_{sx3err} & u_{sy3err} \end{bmatrix}^T \) are the real error voltage of xy axis. In fact, the actual error voltage cannot be obtained. Thus, \( u_{err} \) needs to be estimated. After introducing the estimates, the observer state equation is obtained as follows:

\[
\dot{x}_3 = M_3 \dot{x}_3 + N_3 (u_{3}^e - \hat{u}_{3err}) + G_3 (y_3 - C_3 \dot{x}_3) \tag{19}
\]

where \( G_3 \) is the feedback gain matrix.

Substituting the estimation parameters into the observer state equation and error state equation can be obtained:

\[
\dot{e}_3 = (M_3 - G_3 C_3) e_3 - N_3 \Delta u_{3err} \tag{20}
\]

where: \( e_3 = x_3 - \hat{x}_3 \), \( \Delta u_{3err} = \Delta u_{err} \left[ f_{x3} f_{y3} \right]^T \).

A positive definite Lyapunov function which is containing parameter error space can be constructed [27], [28], [30]:

\[
V = \frac{1}{2} \mathbf{e}^T P \mathbf{e} + \frac{\lambda_u}{2} (u_{err} - \hat{u}_{err})^2 \quad \mathbf{P} \in \mathbb{R}^{4 \times 4} \tag{21}
\]

where: \( P \) is a symmetric positive definite matrix.

The derivative of the Lyapunov function (21) becomes:

\[
\dot{V} = \frac{1}{2} \mathbf{e}^T \left( (M_3 - G_3 C_1)^T P + P (M_3 - G_3 C_1) \right) \mathbf{e} - \mathbf{e}^T P N_3 \Delta u_{err} - \lambda_u \Delta u_{err} \frac{\hat{u}_{err}}{dt} \tag{22}
\]

According to the Lyapunov stability theorem, a sufficient condition for the asymptotic stability of the estimator is that the candidate function \( V \) must be decreased while the error is not zero [28], [30]. Therefore, the first-order time derivative of (21) must be negative when \( \mathbf{e} \neq 0 \). \( M_{3g}^T \mathbf{P} + \mathbf{P} M_{3g} \) can be guaranteed to be negative definite matrix through the design of \( G_3 \) and \( \mathbf{P} (M_{3g} = M_3 - G_3 C_1) \). Generally, \( \mathbf{P} \) is assumed to be the identity matrix, so that the gain matrix needs to be designed which will be discussed in the next part. If the gain matrix is properly designed, the Lyapunov condition will eventually be satisfied, if the sum of the last two terms in (22) is zero as:

\[
- \mathbf{e}^T P N_3 \Delta u_{err} - \lambda_u \Delta u_{err} \frac{\hat{u}_{err}}{dt} = 0 \tag{23}
\]

Since the flux can not be measured, the estimated error of the flux can be ignored, which would not affect the stability in practice [29]. By substituting (17), into (23), the adaptation law can be described by:

\[
\frac{\hat{u}_{err}}{dt} = - \frac{L_{r3} (e_{is3x f_{x3}} + e_{is3y f_{y3}})}{\lambda_u (L_{m3 L_{r3} + L_{r3} L_{r3})}} \tag{24}
\]

\section*{C. DESIGN OF GAIN MATRIX}

If the feedback gain matrix is not designed, the coefficient matrix of the first term in (22) cannot be guaranteed to be a negative definite matrix. It means that the above adaptive law (24) cannot satisfy the asymptotic stability defined by Lyapunov. Therefore, the gain matrix need to be designed according to Lyapunov theory. The design of feedback matrix becomes the requirement of this paper.

\[
G_3 = \begin{bmatrix} g_1 & g_2 & g_3 & g_4 \end{bmatrix} \tag{25}
\]

We assumed that:

\[
M_{3g}^T \mathbf{P} + \mathbf{P} M_{3g} = \mathbf{Q} \tag{26}
\]

Obviously, is a fourth order real symmetric matrix. \( |\mathbf{Q}_n| \) represents leading principal minor of \( \mathbf{Q} \), where \( n = 1, 2, 3, 4 \). Substituting (25) into (26) can we obtain:

\[
|\mathbf{Q}_1| = 2 (m_{11} - g_1) \tag{27a}
\]

\[
|\mathbf{Q}_2| = 4 (m_{11} - g_1)^2 \tag{27b}
\]

\[
|\mathbf{Q}_3| = (m_{13} + m_{31} - g_3)^2 - (m_{32} - m_{14} - g_4)^2 \tag{27c}
\]

\[
|\mathbf{Q}_4| = (4 (m_{11} - g_1) m_{33} - (m_{13} + m_{31} - g_3)^2 - (m_{32} - m_{14} - g_4)^2 \tag{27d}
\]

In order to the satisfy Lyapunov stability theorem, \( \mathbf{Q}_4 \) must be a negative definite matrix. \( G_3 \), a necessary not sufficient condition for the stability is easy to be designed as:

\[
g_1 = 2\delta
\]

\[
g_2 = 0
\]

\[
g_3 = m_{13} + m_{31}
\]

\[
g_4 = m_{32} - m_{14}
\]

where \( \delta > 0 \).

\section*{V. EXPERIMENT AND EVALUATION}

\subsection*{A. EXPERIMENTAL SYSTEM SETUP}

The proposed algorithm is validated on a seven-phase induction motor experimental platform. Fig.7 displays the complete experimental platform, which is composed of DC power supply, DSPACE-1005 controller, multiphase VSI and alternating current servo load. The multi-phase VSI is constructed by several groups of IGBT device, filter capacitors, current and voltage sensors, etc. In particular, the IGBT model is the INFINEON FF150R12ME3G. The sampling frequency of the platform is 10 kHz, and the dead time is set to 3 μs to ensure the safety of the system. Table 1 and Table 2 exhibit all parameters of the self-designed squirrel-cage seven-phase induction motor.
B. COMPARISON OF ONLINE ADAPTIVE METHODS

If the motor is running at high speed, the phase voltage of the motor is very large. At this point, the voltage distortion generated by the VSI has little effect on motor performance, or even can be ignored. However, when the motor speed is lower, the dead-time effect is more obvious. Thus, all experiments were carried out at 2% of rated speed.

In order to verify the superiority of the proposed method in parameter identification performance, the proposed method is compared with the traditional MRAS method which is described in detail in the Appendix. The seven-phase machine runs steadily at 30 rpm and rated-load condition. The algorithm starts at 0.5 s. Fig.8 (a) exhibits that $u_{err}$ estimated by traditional method can converge to 6.34 V and 2.49 V when the bus voltage is set 200 V and 100 V respectively. When the bus voltage up to 400 V, traditional method has poor performance. It can not converge to a constant. At $V_{dc} = 400$ V, when the deadtime is set to 6 $\mu$s, the system crashed after 1 second, which is illustrated in Fig.8 (b).

However, the proposed method can operate stably under all the above conditions and can converge stably to a constant. In Fig.9 (a), the estimated error voltage converges to 3.95 V when the bus voltage is 100 V. When $V_{dc}$ increase to 200 V and 400 V respectively, $u_{err}$ become 6.54 V and 11.52 V after 1 s. In Fig.9 (b), it can be seen that when $t_{dead}$ is adjusted to 4 $\mu$s and 6 $\mu$s under $V_{dc} = 400$ V, estimated value of voltage error converge to 15.25 V and 22.72 V correspondingly.
According to the IGBT parameters from data sheet, the reference value of $u_{err}$ can be calculated, which is list in Table 3. It can be seen that, under above conditions, the deviation between the estimated value and the reference value is about 5%. The errors are sufficient for most industry applications. Therefore, the evaluation of the proposed method is reliable.

C. ESTIMATE OF $\hat{u}_{err}$ AT VARIOUS CONDITIONS

During the actual operation, the system load is usually changed. In order to test the performance of the proposed algorithm in practical applications, the load variation experiments are given in Fig.10. Fig.10 (a) shows the running state of the motor. The load is reduced to half at around 0.8 s and increased to full load at 2 s. The motor speed also varies with the load. However, the VSI non-linear identified by the proposed method does not change much. The fluctuation factor of $\hat{u}_{err}$ is 5.1% which is shown as Fig.10 (b).

As we all know, the long-term operation will result in the temperature increasing of the motor and the winding resistance will change. Therefore, this paper also shows the performance of the proposed algorithm when the actual parameters of the motor deviate from the control parameters. Fig.11 shows the VSI error identification curve when resistance deviation occurs in the proposed method. In Fig.11, $R_{s\text{ref}}$ is the reference in the control system, and $R_{s\text{act}}$ is the actual value. When the reference resistance is 20% greater than the actual resistance, the estimate of $u_{err}$ will be an about 6.5% reduction. Otherwise, when the reference resistance is 20% smaller than the actual resistance, the estimate of $u_{err}$ will rise about 6.7%. It can be seen that the proposed method is affected by the variation of motor parameters. The reason is that the proposed method is based on the model of the air-gap flux observer. When the model itself has errors, the parameter identification results will also produce errors.
D. COMPARATIVE EXPERIMENT OF COMPENSATION EFFECT

The dead-time effect of seven-phase IGBTs mainly produce (14±n)th order harmonic current. Therefore, the performance of proposed method is mainly reflected in the suppression of these harmonics. As third harmonic is injected actively, third harmonic would be much larger than other harmonics. It leads to the the value of total harmonic distortion (THD) can not reflect the advantage of the algorithm. Thus, a selective harmonic distortions (SHD) was adopted, which is defined as:

\[
\text{SHD} \% = \frac{\sqrt{I_9^2 + I_{11}^2 + I_{13}^2 + I_{15}^2 + I_{17}^2 + I_{19}^2}}{I_1} \tag{28}
\]

where \(I_k\) is the magnitude of the \(k\)th order harmonic [33].

In Fig.12, Fig.13, Fig.14 and Fig.15 the prototype was operated with a rated load and 30 rpm. The bus voltage is 400 V, and dead time is adjusted to 3 \(\mu\)s. Fig.12 (a) displays the current of phase A without using any VSI nonlinearity compensation. It can be seen that the current has obvious distortion near zero point which is caused by the inverter dead-time effect. After FFT analysis of the current, the \(\text{SHD} = 4.95\%\) when there is no compensation for the drive system as shown in Fig.12 (b). It is worth noting that the 9th and 11th order harmonics dominate. These high order harmonics increase the loss of the motor and greatly harm the performance of the motor.

As for Fig.13 (a), when the offline feedforward method was adopted, the current distortion has been little improved on the zero crossing point. The value of SHD decrease to 3.22\%, which is shown as Fig.13 (b). Among them, there has been a noticeable decrease in 9th and 11th order harmonics. In Fig.14, after using the traditional MRAS method, as the compensation voltage is higher than the required voltage, the zero crossing of the current curve is not smooth. Fortunately, its SHD decrease to 2.55\%. The zero current clamping phenomenon is almost eliminated which is shown as Fig.15 (a). Moreover, the harmonic content decreases more significantly, \(\text{SHD} = 2.43\%\). These experiments are enough to conclude that the proposed method not only can adapt to various operating conditions, but also has obvious advantages in harmonic suppression.

VI. CONCLUSION

Non-sinusoidal power supply is a significant application for multiphase motors. In this paper, a VSI nonlinearity identification method based on third harmonic injection is proposed. In addition, this method utilizes third harmonic plane only, and the fundamental plane can be used for other parameter estimation purposes. In order to make the adaptive law of voltage error stable, we designed a feedback gain matrix of a full-order air-gap flux observer. Based on the above algorithm, a series of experiments are carried out on a seven-phase induction motor. By comparing the estimated values with the reference values which is calculated from IGBT data sheet, the errors are satisfactory under different conditions. It is important to note that the accuracy of the estimates is affected by parameters of the flux observer, especially the stator resistance. Finally, the proposed method can not only estimate the reliable voltage error of the inverter, but also effectively reduce the SHD.

APPENDIX

In this Appendix, the traditional VSI nonlinearity estimation method mentioned in the experiment is given [22], [23]. In [22], the authors obtained the VSI error by direct calculation which is expressed as:

\[
u_{err} = \frac{u_{sy3} - u_{sy3}}{f_y} \tag{A1}
\]

The actual voltage of the motor can be calculated from the flux equation and voltage equation. Then, (A1) can be
The denominator of (A2) contains $f_y$ and $f_y^*$ close to zero, so that the perturbation of $f_y$ is going to be amplified in the estimation. A low pass filter is used to reduce this disturbance [22], but the effect is not obvious. Therefore, after two years, they optimized the perturbation of the estimated value by using MRAS [23]. As for the induction machine in this article, there are two types of air-gap flux observer, voltage model and current model. For the current model does not contain voltage variables, it is used as the reference model.

$$\psi_{mx3} = L_{m3}i_{sx3} \times \frac{L_{m3}L_{sigma}}{\sigma_3L_3} (\omega_3 - \omega_{psi3}) L_{sigma3}i_{sx3} \quad (A3)$$

On the other hand, voltage model is regarded as adaptive model which is expressed as (A4). It is noticed that these two models are steady-state mathematical models of air-gap flux.

$$\psi_{mx3}^* = u_{sy3}^* - L_{sigma3} f_y^* L_{m3} i_{sy3} - L_3 i_{sy3} \times \frac{L_{m3}L_{sigma}}{\sigma_3L_3} (\omega_3 - \omega_{psi3}) L_{sigma3}i_{sx3} \quad (A4)$$

The $u_{err}$ can be estimated by the error of the two air-gap fluxes.

$$\hat{u}_{err} = \frac{k_{ps} + k_l}{s} (\psi_{mx3} - \psi_{mx3}^*) \quad (A5)$$

Fig. 16 shows the block diagram of traditional MRAS method.

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