Voltage Control of IPMSM Servo Drive in Constant Power Region With Intelligent Parameter Estimation

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ABSTRACT A novel voltage control scheme for an interior permanent magnet synchronous motor (IPMSM) servo drive in the constant power region with intelligent estimation of the motor parameter is proposed in this study. In the novel voltage control scheme, a feedforward voltage angle controller is proposed where an intelligent parameter estimation method by using a wavelet fuzzy neural network (WFNN) is developed to estimate the $q$-axis inductance online. In this study, in order to minimize the copper loss, a flux-weakening (FW) control scheme under maximum phase voltage is developed first. Then, an adaptive backstepping based nonlinear controller (ABNC) considering nonzero $d$-axis current is developed to improve the robustness of the speed control. The Lyapunov stability theorem is used to derive the adaptive law of the online estimation of the lumped uncertainty to ensure the asymptotical stability of the ABNC. Moreover, a feedforward voltage angle controller is developed for the voltage control where the $q$-axis current controller is retained in order to ensure the steady-state performance of the control system. Furthermore, the WFNN is adopted to estimate the actual $q$-axis inductance value online for the feedforward voltage angle controller to improve the dynamic response. In addition, some experimental results are demonstrated to verify the effectiveness of the proposed voltage control scheme with ABNC in the constant power region.

INDEX TERMS Interior permanent magnet synchronous motor (IPMSM), adaptive backstepping based nonlinear controller (ABNC), maximum torque per ampere (MTPA) control, flux-weakening (FW) control, online parameter estimation, wavelet fuzzy neural network (WFNN).

I. INTRODUCTION
There are many attractive characteristics of the interior permanent magnet synchronous motors (IPMSMs) including wide speed operating range, superior power density, high efficiency, and high torque-to-inertia ratio. These features permit the IPMSMs to be operated not only in the constant torque region but also in the constant power region up to a high speed by using flux weakening. Thus, IPMSMs has been adopted in many industrial applications [1], [2], [3], [4]. In order to improve the control performance and efficiency of the IPMSM servo drives, optimal control methods such as maximum torque per ampere (MTPA) control and flux-weakening (FW) control have been proposed [1], [2], [3], [4]. In order to utilize the advantages of the reluctance torque term of the IPMSMs in the constant torque region, the MTPA control has been developed to improve the torque output. Furthermore, the FW control is an important issue in the range of high speed with the back electromotive force (EMF) increasing along with the rising speed, which will reach the limit of dc-link voltage in the constant power region.

There are mainly two strategies to achieve the FW control which are the current control methods [4], [5], [6], the voltage control methods [7], [8], [9], [10], [11]. In [4], to limit the inverter output voltage to the maximum phase voltage of the inverter at high speed, a voltage control loop was designed for the current control in the constant power region. A FW control was achieved by using current control with the flux level
being adjusted automatically by the outer voltage regulation loop to prevent saturation of the current regulator in [5]. In [6], an improved FW control algorithm of IPMSMs based on torque feedforward technique was proposed to generate large and stable torque in the FW region. Moreover, traditional FW control using current control methods, which are developed from the vector control methods, usually adopt two proportional-integral (PI) regulators to control the \( d \)-axis and the \( q \)-axis current separately. However, the stability of the control system of traditional FW control may be affected by the undesired current saturation phenomenon. Though many anti-windup strategies for the permanent magnet synchronous motors (PMSMs) were presented in the literature [6], the conflict between the two current regulators still lead to the slow dynamic response or even the instability of the control system in high speed [5]. On the other hand, the voltage control methods can make fully use of dc-link voltage and eliminate the conflict between the two current regulators [7], [8], [9], [10], [11]. In [7], a novel torque control based on voltage phase angle control was proposed to provide more accurate torque for the IPMSMs in the constant power region. A voltage angle based FW control scheme appropriate for the operation of surface PMSMs over a wide range of speed was proposed in [8]. A single current regulator which controls the \( d \)-axis current actively with a fixed \( q \)-axis voltage command was proposed in [9]. In [10], a single \( q \)-axis current regulator with variable voltage angle control method was proposed. In [11], a voltage control scheme was developed to control the torque with the voltage angle in the FW region when the inverter output voltage amplitude is saturated. Furthermore, an online stator flux estimation method was proposed in [12], in which a real-time torque control was proposed to satisfy both torque control accuracy and high-efficiency operation in consideration of flux linkage variations in both the MTPA and FW regions. Both stator flux linkages and dynamic inductances are estimated online in [12] to consider the flux variations in real time.

Although the control performance and dynamic response of the IPMSMs are excellent, the motor parameters are easily varying at different operating conditions. Therefore, the development of parameters estimation of IPMSMs has been a popular research topic in the past two decades [13], [14], [15], [16], [17], [18]. In [13], an online parameter estimation method based on a discrete-time dynamic model for the IPMSMs was proposed. The proposed method consists of two affine projection algorithms and has adopted the difference in dynamics of motor parameters. Moreover, an estimation method of the spatial inductance map by spatially scanning the motor using the sinusoidal voltage injection was developed in [14]. In [15], a real-time method to estimate the inductances of an IPMSM by using the measured dc-link voltage of the inverter and the derivatives of the stator currents at certain voltage vectors during each pulse width modulation cycle was proposed. Furthermore, an adaptive online parameter estimation method for the high-speed control of an IPMSM drive was developed in [16]. In [17], the Adaline NN algorithm was employed to design the estimators for the rotor flux linkage and stator winding resistance. In addition, an online parameter estimation methodology using \( d \)-axis current injection, which can estimate the distortion voltage of the current-controlled voltage source inverter (CCVSI), the varying \( dq \)-axis inductances, and the rotor flux, was proposed in [18]. However, most of the parameter estimation methods mentioned above were only developed for the control of the IPMSMs in the constant torque region.

The PI speed controller is largely adopted in many control applications due to its simplicity. Nevertheless, the disadvantages of the PI controller, such as sensitive to parameter variations and external disturbances, is well known. On the other hand, the backstepping control, as a systematic and nonlinear recursive design method, has attracted much attention for the nonlinear feedback control [19], [20], [21], [22], [23]. Moreover, the backstepping control is based on Lyapunov stability theory. Its control law is derived by constructing the Lyapunov function, and the global asymptotic stability is ensured. However, the sign function in the backstepping control may cause undesired chattering phenomena. Therefore, to reduce the chattering phenomena and improve the control performance, some control methods such as adaptive control [19], [20], intelligent control [21], and sliding mode control [22], [23], have been proposed to merge with the backstepping control. Therefore, one of the objectives of this study is to replace the conventional PI speed controller by an adaptive backstepping based nonlinear controller (ABNC).

To improve the control performance of an IPMSM servo drive in the constant power region, a novel voltage control scheme with a feedforward voltage angle controller is developed in this study. Moreover, an intelligent parameter estimation method by using a wavelet fuzzy neural network (WFNN) [24], [25], [26] is proposed to estimate the \( q \)-axis inductance online for the feedforward voltage angle controller. This study is organized into six sections. The dynamic analysis of a field-oriented control (FOC) IPMSM servo drive with traditional FW control is studied in Sec. II. Then, an ABNC speed controller considering nonzero \( d \)-axis current with adaptive online estimation of the lumped uncertainty is discussed in Sec. III. Furthermore, the proposed voltage control by using a feedforward voltage angle control with the \( q \)-axis current controller is introduced in Sec. IV. In addition, the experimentation based on a TMS320F28075 32-bit floating-point digital signal processor (DSP) with some experimental results to verify the effectiveness of the proposed voltage control scheme is presented in Sec. V. Finally, some conclusions are addressed in Sec. VI.

The main contributions of this study are listed as follows:

1. An ABNC speed controller considering nonzero \( d \)-axis current is proposed to improve the robustness of the speed control.
2. A feedforward voltage angle controller, in which an intelligent parameter estimation method using WFNN is adopted to estimate the \( q \)-axis inductance online, is developed.
3. (3) The current control and voltage control modes with ABNC speed controller to achieve the current and voltage
control of an IPMSM servo drive at different speeds and load torque conditions are successfully implemented.

II. TRADITION FW CONTROL OF IPMSM

The voltage model at steady state of an IPMSM in the dq reference frame can be expressed as follows:

\[
\begin{align*}
v_d &= r_s i_d - \omega_e L_q i_q \\
v_q &= r_s i_q + \omega_e (L_d i_d + \lambda_m)
\end{align*}
\]

where \(v_d\) and \(v_q\) are the \(d\)-axis and \(q\)-axis voltages; \(i_d\) and \(i_q\) are the \(d\)-axis and \(q\)-axis currents; \(r_s\) is the stator resistance; \(L_d\) and \(L_q\) are the \(d\)-axis and \(q\)-axis inductances; \(\lambda_m\) is the permanent-magnet (PM) flux of the rotor; \(\omega_e\) is the electrical speed. Moreover, voltage-fed inverters are widely used in the IPMSM servo drive system. The maximal phase voltage limit \(V_{s,\text{max}}\) of the inverter is defined by the dc-link voltage \(V_{dc}\), which equals to \(V_{dc}/\sqrt{3}\), for the space vector pulse width modulation (SVPWM). Furthermore, the current limit \(i_{s,\text{max}}\) is defined by the machine rated current \(i_{\text{rated}}\). The following two constraints must be fulfilled for the control of an IPMSM:

\[
\begin{align*}
v_d^2 + v_q^2 &= v_e^2 \leq v_{s,\text{max}}^2 \\
i_d^2 + i_q^2 &= i_s^2 \leq i_{s,\text{max}}^2 = (\sqrt{2} i_{\text{rated}})^2
\end{align*}
\]

For the speeds above the rated speed, where the stator resistance can be neglected, (1) and (2) can be simplified as follows:

\[
\begin{align*}
v_d &= -\omega_e L_q i_q \\
v_q &= \omega_e L_d i_d + \omega_e \lambda_m
\end{align*}
\]

Combining (3), (5) and (6), the voltage limit ellipse of an IPMSM can be obtained as follows:

\[
\frac{(i_d + \frac{\omega_e L_d}{L_q})^2}{(i_{s,\text{max}}/L_d)^2} + \frac{i_q^2}{(i_{s,\text{max}}/L_q)^2} = 1
\]

According to (4) and (7), the operation limit of IPMSM with the current limit circle and voltage limit ellipses in \(dq\)-axis current plane with four different electrical speeds and electromagnetic torques are shown in Fig. 1 where \(T_e\) is the electromagnetic torque. In addition, the MTPA curve is obtained by using (26) and will be discussed in Sec. IV. In the constant torque region, the operating points are located on the MTPA curve for minimizing the copper loss. Additionally, the overlapping between the current limit circle and voltage limit ellipse defines the FW operating region. The operating region becomes smaller along with the shrinking of the voltage limit ellipse when the electrical speed increases [6], [11].

Fig. 2 is the block diagram of an FOC IPMSM servo drive by using the SVPWM CCVSI with the conventional MTPA and FW control. In Fig. 2, first, the actual rotor position \(\theta_m\) of the motor is determined by the encoder, and the mechanical speed \(\omega_{rm}\) is obtained by differentiating \(\theta_m\). The electrical angle \(\theta_e\) can be obtained by multiplying \(\theta_m\) with the pole pairs \(P/2\). Next, the mechanical speed \(\omega_{rm}\) is subtracted from the mechanical speed \(\omega_{rm}\) to obtain the mechanical speed \(\omega_{s, \text{err}}\). Then, \(\omega_{s, \text{err}}\) is inputted into PI speed controller to get the \(q\)-axis current command \(i_{q, \text{err}}\). Moreover, \(i_{q, \text{err}}\) is substituted into the MTPA formula to derive the \(d\)-axis current command \(i_{d, \text{MTPA}}\) for the MTPA control. While the FW control is proceeded, the motor speed will be increased above the rated speed. The input of the MTPA block shown in Fig. 2 will be switched to FW and the \(i_{d, \text{MTPA}}\) will be kept constant during the FW control. Furthermore, to make sure that the stator voltage command \(v_s^*\) will not exceed the phase voltage limit \(V_{s, \text{max}}\) of the inverter, \(V_{s, \text{max}} - v_s^*\) is inputted into a PI controller to generate the variation of \(d\)-axis current command \(\Delta i_d^*\) which is a negative value. Only when the inverter output voltage \(v_s^*\) exceeds the maximum phase voltage \(V_{s, \text{max}}\) of the inverter, the input of the PI controller will be nonzero. As shown in Fig. 2, the three-phase currents \(i_q, i_b\) and \(i_c\) of the CCVSI are transformed to the corresponding \(q\)-axis current \(i_q\) and \(d\)-axis current \(i_d\) by using the coordinate transformation. In addition, \(i_q\) and \(i_d\) are subtracted respectively from \(i_{q, \text{err}}\) and \(i_{d, \text{MTPA}}\), and then the \(dq\)-axis voltage commands \(v_{d, \text{err}}^*\) and \(v_{q, \text{err}}^*\) are obtained through the PI controllers of the current loop with the decouple control shown in Fig. 2. After \(v_{d, \text{err}}^*\) and \(v_{q, \text{err}}^*\) are obtained, \(v_{d, \text{err}}^*\) and \(v_{q, \text{err}}^*\) are derived by using coordinate transformation. Additionally, the switching signals of the insulated-gate bipolar transistors (IGBTs) of the CCVSI are generated.
through SVPWM. Finally, the switching signals are sent to the IGBTs to achieve the conventional MTPA and FW control. However, when the IPMSM servo drive is operated above the rated speed, the undesired current saturation phenomenon caused by the integrator windup problem of the PI current controllers is the main disadvantage of the conventional FW control system.

### III. ADAPTIVE BACKSTEPPING BASED NONLINEAR CONTROLLER

Fig. 3 is the block diagram of the proposed voltage control IPMSM servo drive where the ABNC is adopted as the speed controller. In Fig. 3, first, $e_1$ is inputted into the ABNC to get the $q$-axis current command, and $i_{d}^*$ is substituted into the MTPA formula to derive the $d$-axis current command $i_{d,MTPA}^*$ for the MTPA control in the constant torque region. While the FW control is proceeded, the motor speed will be increased above the rated speed. As the stator voltage command reaches the maximum phase voltage $V_{s,max}$, the mode selector shown in Fig. 3 will be switched to the voltage control mode for the constant power region. Moreover, the WFNN is adopted to estimate the correct $q$-axis inductance value $\hat{L}_q$ online. The $\hat{L}_q$ is used to obtain the feedforward voltage angle command $\theta_{vf}^*$ as shown in the block of voltage control mode of Fig. 3. Furthermore, the $q$-axis current error $e_q$ is inputted into PI controller to generate the compensating value of voltage angle $\theta_{v}^*$. Owing to the voltage magnitude is fixed in voltage control mode, the $dq$-axis voltage commands can be obtained from the voltage angle command $\theta_{vf}^*$.

The mechanical dynamic equation of the IPMSM servo drive system can be represented as follows:

$$T_e = J\dot{\omega}_{rm} + B\omega_{rm} + T_L \quad (8)$$

where $J$ is the inertia coefficient; $B$ is the damping coefficient; $T_L$ is the load torque. By neglecting the load torque, (8) can be modified as

$$\dot{\omega}_{rm} = - \frac{B}{J}\omega_{rm} + \frac{T_e}{J} \quad (9)$$

The developed electromagnetic torque $T_e$ can be represented by the following equation:

$$T_e = \frac{3P}{2}\frac{2}{2}\left[\lambda_m i_q + (L_d - L_q)i_d i_q\right] \quad (10)$$
The IPMSM servo drive system can be formulated by rewriting (9) and (10) as follows:

\[
\dot{\omega}_{rm} = -\frac{\ddot{\theta}}{J_{rm}} + \frac{3p\rho_{rm}}{4J} i_q^s + \frac{3p(\dot{L}_d - \dot{L}_q)}{4J} i_d^s + A_m\dot{\omega}_{rm} + B_m i_q^s + C_m i_d^s + E_m
\]

where \( A_m = -\ddot{\theta}/J; B_m = \frac{3p\rho_{rm}}{4J}; C_m = \frac{3p(\dot{L}_d - \dot{L}_q)}{4J}; \ddot{\theta}, J, \dot{\omega}_{rm}, \dot{L}_d \) and \( \dot{L}_q \) are the nominal values of damping coefficient, moment of inertia, PM flux, \( d \)-axis inductance and \( q \)-axis inductance, respectively. By considering the uncertainties including the existence of parameter variations and external disturbances of the IPMSM servo drive system, (11) can be rewritten as

\[
\dot{\omega}_{rm} = (A_m + \Delta A_m)\omega_{rm} + (B_m + \Delta B_m)i_q^s + (C_m + \Delta C_m)i_d^s + (D_m + \Delta D_m)T_L
\]

where \( D_m = -1/J; \Delta A_m, \Delta B_m, \Delta C_m \) and \( \Delta D_m \) are the time-varying parameter variations. Then, \( E_m \) is named the lumped uncertainty and defined as follows:

\[
E_m = \Delta A_m\omega_{rm} + \Delta B_m i_q^s + \Delta C_m i_d^s + (D_m + \Delta D_m)T_L
\]

And \( E_m \) is assumed to be bounded

\[
|E_m| \leq \rho
\]

where \( \rho \) is a given positive constant.

Define the speed tracking error and its derivative term as follows:

\[
e_1 = \omega^*_m - \omega_{rm}
\]

\[
\dot{e}_1 = \dot{\omega}^*_m - \dot{\omega}_{rm}
\]

where \( \dot{\omega}_{rm} \) can be viewed as a virtual control input. Then, define the following stabilizing function as:

\[
\lambda_1 = -c_1 e_1 - \dot{\omega}^*_m
\]

where \( \lambda_1 \) is a positive constant. Moreover, a virtual control error is defined as:

\[
e_2 = \dot{\omega}_{rm} + \lambda_1 = \dot{\omega}_{rm} - c_1 e_1 - \dot{\omega}^*_m
\]

Furthermore, a Lyapunov function is chosen as follows:

\[
V = \frac{1}{2} e_1^2 + \frac{1}{2\delta} |\dot{E}_m|^2 > 0
\]

where \( V \) is a positive definite function; \( \dot{E}_m = E_m - \dot{E}_m \) and \( \dot{E}_m \) is the estimated value of \( E_m \); \( \delta \) is a positive constant. Taking the time derivative of the Lyapunov function and using (12) and (18), one can obtain:

\[
\dot{V} = e_1\dot{e}_1 + \frac{1}{\delta} \dot{E}\dot{E} = e_1(\omega^*_m - \omega_{rm}) - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= e_1(-c_1 e_1 - e_2) - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= -c_1 e_1^2 - (\omega^*_m - \omega_{rm})e_2 - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= -c_1 e_1^2 - [\omega^*_m - (1/A_m)(\dot{\omega}_{rm} - B_m i_q^s - C_m i_d^s - E_m)]
\]

\[
\times e_2 - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= -c_1 e_1^2 - [\omega^*_m - \omega_{rm} + (1/A_m)(B_m + C_m i_d^s)]
\]

\[
\times e_2 - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= -c_1 e_1^2 - [\omega^*_m - \omega_{rm} + (1/A_m)(B_m + C_m i_d^s)]
\]

\[
\times e_2 - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

\[
= -c_1 e_1^2 - \frac{1}{\delta} \dot{E}_m \dot{E}_m
\]

Substituting (21) and (22) into (20), the following equation can be obtained:

\[
\dot{V} = -c_1 e_1^2 - c_2 e_2^2 \leq 0
\]

Therefore, according to (20), the ABNC control law \( U_{ABNC} \) and adaptive law \( \dot{E}_m \) are designed as follows:

\[
U_{ABNC} = i_q^s = \frac{1}{B_m + C_m i_d^s}(\dot{\omega}_{rm} - A_m \omega_{rm} - \dot{E}_m + A_m c_2 e_2)
\]

\[
\dot{E}_m = -\frac{\delta}{A_m} e_2
\]

Since \( \dot{V}(t) \leq 0 \) is negative semidefinite and \( V(t) > 0 \), it implies that \( e_1, e_2, \text{ and } \dot{E}_m \) are nonincreasing and bounded. According to Lyapunov Theorem and Barbalat’s Lemma, \( e_1 \) and \( e_2 \) will converge to zero as \( t \to \infty \). Thus, considering the dynamic equation of the IPMSM servo drive system represented by (12), if the ABNC control law and adaptive law are designed as (21) and (22), the asymptotically stable of the IPMSM servo drive system using the ABNC speed controller can be guaranteed.

IV. NOVEL VOLTAGE CONTROL SCHEME OF IPMSM

To achieve wide speed range operation, two control modes are proposed for the IPMSM servo drive and the switch signal is obtained by using the mode selector as shown in Fig. 3. In the current control mode, the switching signal of the mode selector turns to the voltage control mode when the stator voltage command reaches the maximum phase voltage (\( V_{s,\text{max}} \approx 179V \)) of the CCVSI. On the other hand, in the voltage control mode, the switching signal of the mode selector turns to the current control mode when the \( d \)-axis current \( i_d \) is larger than \( d \)-axis current command of MTPA \( i_d^{\text{MTPA}} \). The above description can be summarized as:

\[
v_s^* \geq V_{s,\text{max}}:
\]

Current Control Mode \( \rightarrow \) Voltage Control mode
A PI controller is adopted to generate the compensating value of voltage angle \( \Delta \theta_v \) by using \( e_q \). Then, the voltage angle command can be expressed as:

\[
\theta_v^* = \theta_{\text{cf}} + \Delta \theta_v^*
\]  

(32)

However, since the \( q \)-axis inductance value \( \hat{L}_q \) shown in (30) varies significantly in FW region, a WFNN [24], [25], [26] is adopted to estimate the \( q \)-axis inductance value \( L_q \) online as shown in Fig. 3 to improve the dynamic response of voltage control.

### C. ESTIMATION OF DISTORTION VOLTAGE

Owing to the nonlinearity and time delay of the CCVSI, the distortion voltage errors between the \( dq \)-axis voltages command and the real \( dq \)-axis voltages of the IPMSM servo drive are inevitable. Thus, considering the distortion voltage errors, the \( dq \)-axis voltages command can be expressed as follows in the FW operating region:

\[
v_d^* = -\omega_e L_q i_q + D_d V_{\text{dead}}
\]

(33)

\[
v_q^* = \omega_e L_d i_d + \omega_e \lambda_m + D_q V_{\text{dead}}
\]

(34)

where \( V_{\text{dead}} \) is the distortion voltage and \( D_d, D_q \) are the distorted coefficients [18]. To estimate the distortion voltage \( V_{\text{dead}} \), the instantaneous measurement of the electrical speed, \( dq \)-axis voltages and currents are required. However, when the motor is operated at high speed, these signals will be contaminated by various noises. Therefore, the smoothing values of the measured signals are adopted to improve the estimation accuracy. Eq. (33), (34) can be rewritten as follows:

\[
\bar{v}_d^* = -\tilde{\omega}_e L_q \bar{i}_q + \bar{D}_d V_{\text{dead}}
\]

(35)

\[
\bar{v}_q^* = \tilde{\omega}_e L_d \bar{i}_d + \tilde{\omega}_e \lambda_m + \bar{D}_q V_{\text{dead}}
\]

(36)

where the smoothing values of \( dq \)-axis voltage commands \( \bar{v}_{dq}^* \), \( dq \)-axis current \( \bar{i}_{dq} \) and electrical speed \( \tilde{\omega}_e \) are averaged every ten times. Moreover, the smoothing values of distorted coefficients \( \bar{D}_d, \bar{D}_q \) are defined as follows [18]:

\[
\bar{D}_d = \frac{1}{N} \sum_{k=1}^{N} 2 \sin[\theta_{\text{rmk}} - \text{int}\left(\frac{3(\theta_{\text{rmk}} + \gamma + \frac{\pi}{6})}{\pi}\right) \times \frac{\pi}{3}]
\]

(37)

\[
\bar{D}_q = \frac{1}{N} \sum_{k=1}^{N} 2 \cos[\theta_{\text{rmk}} - \text{int}\left(\frac{3(\theta_{\text{rmk}} + \gamma + \frac{\pi}{6})}{\pi}\right) \times \frac{\pi}{3}]
\]

(38)

where \( N = 360 \) for 360 degrees; \( \gamma \) is the current angle; \( \theta_{\text{rmk}} \) is the \( k \text{th} \) rotor angle. Besides, \( \bar{D}_d \) and \( \bar{D}_q \) are only functions of the current angle and are independent of rotor position. Since the calculation of the smooth values needs to substitute and accumulate the rotor angle 360 times according to different current angles, it will take a large amount of calculation and is impossible to obtain the smooth values of the distortion coefficients in one interrupt sampling time. For
Thus, the distortion voltage $V_{\text{dead}}$ can be estimated as follows:

$$
\hat{V}_{\text{dead}} = \frac{\hat{V}_d^2 + \hat{V}_q^2 - \omega_c T_e}{(\hat{D}_d \hat{d}_d + \hat{D}_q \hat{q}_q)}
$$

where $\hat{V}_{\text{dead}}$ is the estimated value of $V_{\text{dead}}$.

### D. ONLINE ESTIMATION OF Q-AXIS INDUCTANCE

The block diagram using the intelligent parameter estimation method to estimate the $q$-axis inductance $\hat{L}_q$ is shown in Fig. 5. There are two mostly adopted neuron networks, the fuzzy neural network (FNN) and the wavelet neural network (WNN), in the intelligent control systems. The FNN possesses the characteristic of fuzzy reasoning in handling uncertain information and the characteristic of artificial neural networks in learning from processes. Moreover, the WNN with reduced network size has the ability of converging quickly with high precision owing to the time–frequency localization properties of wavelets. Since the WFN as shown in Fig. 5. The $q$-axis current $i_q$ obtained from the IPMSM servo drive, i.e. the voltage model of the IPMSM servo drive, is subtracted from the estimated $q$-axis current $\hat{i}_q$ of the estimated voltage model of the IPMSM servo drive to obtain the error $e$. Then, and its derivative $\dot{e}$ are inputted into the WFN to obtain the adaptation value of the $q$-axis inductance $\Delta \hat{L}_q$ online.

The coupling terms of both the real and estimated $q$-axis voltage model are required as shown in Fig. 5. Therefore, after some mathematical manipulations using (3), (33) and (34), one can obtain:

$$
\omega_c L_d i_d + \lambda_m = \sqrt{v_{s,max}^2 - (\omega_c L_q i_q + D_d V_{\text{dead}})^2} - D_q V_{\text{dead}}
$$

Substituting $\Delta d$, $\Delta q$, $\dot{V}_{\text{dead}}$, $\alpha_p$ and $i_q$ into (44), the coupling term of the estimated voltage model can be expressed as:

$$
\omega_c (L_d i_d + \lambda_m) = \sqrt{(v_{s,max})^2 - (\omega_c \hat{L}_q i_q + \hat{D}_d \hat{V}_{\text{dead}})^2} - \hat{D}_q \hat{V}_{\text{dead}}
$$

Moreover, the output of the WFN is the adaptation value of the $q$-axis inductance $\Delta \hat{L}_q$. Thus, the estimated value of the $q$-axis inductance can be obtained by using the following equation:

$$
\hat{L}_q(N + 1) = \hat{L}_q(N) + \Delta \hat{L}_q(N)
$$

The adaptation is processed recursively until $\hat{i}_q = i_q$, which is the same as the traditional model following control scheme. Then, the steady-state estimation value of the $q$-axis inductance $\hat{L}_q$ at a specific high-speed operating condition can be obtained. Furthermore, since the coupling term shown in (45) is replaced by the right hand side of the equality, the variations of $L_d$ and $\lambda_m$ are effectively taken into account during the adjustment of the $q$-axis inductance.

### V. EXPERIMENTATION

The experimental setup of the IPMSM servo drive system are shown in Fig. 6. The IPMSM test platform is composed
of an IPMSM, a gearbox (with gear ratio 4:1), a torque meter and a magnetic powder brake. The detailed information of the magnetic powder brake and IPMSM is listed in Table 1. A TMS320F28075 32-bit DSP with 120 MHz is adopted in this study to develop the DSP-based servo drive system. Moreover, a torque meter with 100 Nm/7000 rpm is utilized to measure the load torque. The resolution of the adopted encoder is 2500 pulses/rotation and multiplied by 4 by the DSP. The maximum speed of the IPMSM in the experimentation is 4000 rpm in the constant power region. Therefore, by using the gearbox with a gear ratio of 4:1 to reduce the speed, the resulted maximum speed 1000 rpm is smaller than the rated speed of the magnetic powder brake 1800 rpm. Furthermore, the ratings of the adopted CCVSI is 5 kW/220 V/14 A. The switching frequency 10 kHz is controlled by the SVPWM technology. In addition, the dc-link voltage provided by an adjustable DC power supply is set at 311 V. According to the SVPWM, the inverter maximum phase voltage \( V_{s,\text{max}} \) was set to be \( 311/\sqrt{3} \approx 179 \text{V} \). In general, the bandwidth of the current and voltage control loops is 10 times of the speed control loop. Additionally, the operation cycles and execution time of the WFNN estimator is 12261 cycles/102.175 \( \mu \text{s} \). Though the proposed WFNN estimator is more complicated, the execution time is still within 1 ms. Thus, the sampling time of the speed, current, voltage control loops are 1 ms, 0.1 ms, 0.1 ms, respectively.

The experimental results are presented to verify the effectiveness of the designed voltage control for the operating speed above the rated speed. The parameters of the proposed ABNC are given as follows:

\[
c_1 = 6.25, \quad c_2 = 5.5, \quad \delta = 130
\]

Moreover, the parameters of the PI controller of the proposed voltage angle controller are given as \( K_p = 0.04 \) and \( K_I = 5 \) where \( K_p \) and \( K_I \) represent the proportional and integral gains. These parameters are tuned by trial and error to achieve the best response. Furthermore, to demonstrate the wide speed control range of the proposed voltage control scheme, the torque-speed curve with four experimental cases are presented in Fig. 7. The first case is the operating condition with rated speed 2000 rpm and rated torque 9.5 Nm. In addition, three operating conditions transforming from the current control mode to the voltage control mode to achieve the constant power control at rated output power 2kW, which are 3500 rpm under 5.43 Nm load torque, 3750 rpm under 5.07 Nm load torque and 4000 rpm under 4.75 Nm load torque, are tested.

For the comparison of the control performance, the experimental results of PI speed controller with proportional gain 2 and integral gain 0.14 are also given. The gains of the PI speed controller are obtained by trial and error to achieve acceptable transient and steady-state responses. To demonstrate the control performance of the PI and ABNC speed controllers, the IPMSM is operated at rated speed 2000 rpm and rated torque 9.5 Nm with 100 rpm step command, and the results are shown in Fig. 8. Figs. 8(a) and 8(b) present the mechanical speed command \( \omega_m^p \), the mechanical speed \( \omega_m \) and the mechanical speed error \( e_1 \). Figs. 8(c) and 8(d) present the \( q \)-axis current command \( i_q^p \), \( q \)-axis current \( i_q \), \( d \)-axis current command \( i_d^p \) and \( d \)-axis current \( i_d \). The stator voltage command \( V_s^p \), the \( q \)-axis voltage command \( V_q^p \) and the \( d \)-axis voltage command \( V_d^p \) are shown in Figs. 8(e) and 8(f). In Fig. 8, since the stator voltage command is less than the maximum phase voltage \( (V_{s,\text{max}} \approx 179 \text{V}) \) of the CCVSI, the

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**TABLE 1. Parameters of magnetic powder brake and IPMSM.**

| Items                      | Units | Quantities |
|----------------------------|-------|------------|
| Magnetic Powder Brake      |       |            |
| Torque                     | Nm    | 50         |
| Rated current              | A     | 2.15       |
| Coil resistance            | Ohm   | 11.14      |
| Mass of powder             | g     | 60         |
| Maximum rotating speed     | rpm   | 1800       |
|                            |       |            |
| IPMSM                      |       |            |
| Pole number                |       | 8          |
| Rated power                | W     | 2000       |
| Rated line voltage         | V     | 220        |
| Rated current              | A     | 10.6       |
| Rated torque               | Nm    | 9.5        |
| Rated speed                | rpm   | 2000       |
| \( d \)-axis inductance    | mH    | 3.48       |
| \( q \)-axis inductance    | mH    | 6.16       |
| Magnetic flux              | Wb    | 0.143      |
| Resistance                 | Ohm   | 0.57       |
| Viscous damping            | Nm/(rad/sec) | 2.69×10^3 |
| Inertia                    | Nm/(rad/sec^2) | 4.07473×10^3 |

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**FIGURE 6. Photo of experimental setup.**

**FIGURE 7. Torque-Speed Curve and four experimental cases.**
switching signal of the mode selector remains at the current control mode with MTPA control shown in (26). Moreover, it can be seen from Figs. 8(a) and 8(b) that the overshoot of the PI speed controller is 31.8%, and the overshoot of the ABNC speed controller is 13.3%. Furthermore, the peak value of speed tracking error of PI and ABNC are 101.31 rpm and 97.63 rpm, respectively. In addition, the speed tracking error of PI and ABNC converge to zero in 0.31 sec and 0.24 sec, respectively. Therefore, comparing with the PI speed controller, the ABNC speed controller possesses more robust control characteristic.

Figs. 9-11 depict the experimental resulted of the current and voltage control for the mechanical speed command increasing from 2000 rpm to 3500 rpm under 5.43 Nm load torque, 2000 rpm to 3750 rpm under 5.07 Nm load torque, and 2000 rpm to 4000 rpm under 4.75 Nm load torque.
respectively. Figs. 9(a), 10(a), 11(a) present the mechanical speed command $\omega_{\text{rm}}^*$, the mechanical speed $\omega_{\text{rm}}$, the mechanical speed error $e_1$ and the $q$-axis current command $i_{q}^e$. Figs. 9(b), 10(b), 11(b) depict the stator voltage command $v'_{\text{s}}$. Figs. 9(c), 10(c), 11(c) show the voltage angle command, the feedforward voltage angle command, the compensating voltage angle command. Figs. 9(d), 10(d), 11(d) illustrate the estimated distorted voltage, the $q$-axis current, the estimated $q$-axis current, the estimated $q$-axis inductance.
v_q^*, the q-axis voltage command v_q^* and the d-axis voltage command v_d^*. Figs. 9(c), 10(c), 11(c) show the voltage angle command \( \theta_v^* \), the feedforward voltage angle command \( \theta_{vf}^* \) and the compensating voltage angle command \( \Delta \theta_v^* \). The estimated distortion voltage \( \hat{V}_{\text{dead}} \), the q-axis current \( \hat{i}_q \), the estimated q-axis current \( \hat{i}_q \), and the estimated q-axis inductance \( \hat{L}_q \) are shown in Figs. 9(d), 10(d), 11(d) by using the scheme shown in Fig. 5. From the experimental results shown in Figs. 9-11, once the stator voltage command \( v_s^* \) reaches the maximum phase voltage \( V_{s_{\text{max}}} \) of the inverter, the switch signal of the mode selector will transit to the voltage control mode from the current control mode. During the transition, due to the robust control characteristic of the ABNC, the speed error converges immediately. Moreover, when the motor operates at rated speed with different load conditions, the operating points are located at the intersections of torque trajectories and MTPA curve. Above the rated speed, the voltage limit ellipse will shrink with the increasing speed. Then, the operating points will move to the intersections of torque trajectories and the voltage limit ellipses. According to Fig. 1, the higher the operating speed, the more negative value of d-axis current \( i_d \) is at the same load condition. The above situation will result in higher voltage angle command \( \theta_v^* \) and more negative value of d-axis voltage command \( v_d^* \) as shown in Figs. 9(b), 10(b), 11(b) and 9(c), 10(c), 11(c). Furthermore, the estimated distortion voltage \( \hat{V}_{\text{dead}} \) can be obtained by using Eq. (43), which is larger at higher speed with lower torque as shown in Figs. 9(d), 10(d), 11(d). The same phenomenon can also be found in [18]. In addition, the estimated \( \hat{L}_q \) are 8.23mH, 7.33mH and 6.43mH, respectively, by using the proposed intelligent q-axis inductance estimation scheme at three FW operating conditions as shown in Figs. 9(d), 10(d), 11(d). The estimated value of the q-axis inductance decreases with higher voltage angle at higher speed. The above result has also been obtained in both [6] and [27]. Additionally, the successful model following dynamics of the estimated voltage model of the IPMSM servo drive and the voltage model of the IPMSM servo drive by the tracking of the estimated q-axis current \( \hat{i}_q \) to the q-axis current \( i_q \) also can be found in Figs. 9(d), 10(d), 11(d). Therefore, the voltage control with rated output power could be achieved at high speed by using the proposed voltage control scheme.

For the comparison of control performance of the q-axis inductance \( \hat{L}_q \) estimation scheme, the WFNN shown in Fig. 5 is replaced by a traditional PI estimation method. Figs. 12(a) and 12(b), Figs. 12(c) and 12(d), and
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Figs. 12(e) and 12(f) depict the experimental results of the current and voltage control for the mechanical speed command increasing from 2000 rpm to 3500 rpm under 5.43 Nm load torque, 2000 rpm to 3750 rpm under 5.07 Nm load torque, and 2000 rpm to 4000 rpm under 4.75 Nm load torque, respectively, by using the PI parameter estimation scheme. Figs. 12(a), 12(c) and 12(e) show the voltage angle command $\theta_v^\ast$, the feedforward voltage angle command $\theta_v^{\ast f}$ and the compensating voltage angle command $\Delta \theta_v^\ast$. Moreover, the estimated distortion voltage $\hat{V}_{\text{dead}}$, the $q$-axis current $i_q$, the estimated $q$-axis current $\hat{i}_q$, and the estimated $q$-axis inductance $\hat{L}_q$ are also 8.23 mH, 7.33 mH and 6.43 mH as shown in Figs. 12(b), 12(d) and 12(f). Though the estimated steady-state values of $\hat{L}_q$ by using PI estimation scheme at
three FW operating conditions are the same as the WFNN estimation scheme shown in Figs. 9(d), 10(d) and 11(d), the sluggish response of the PI parameter estimation scheme obstructs its usefulness in practical applications.

Comparing with the real-time torque control and online stator flux estimation method developed in [12], an ABNC speed controller considering nonzero d-axis current was proposed in this study to improve the robustness of the speed control. Moreover, a feedforward voltage angle controller, in which an intelligent parameter estimation method using WFNN is adopted to estimate the q-axis inductance online, was successfully developed for the speed control in the constant power region. On the other hand, the estimation model shown in (33) and (34) ignores the influence of stator resistance. If it is considered, the estimation accuracy can be improved. Besides, the operation cycles and execution time of the WFNN estimator is 12261 cycles/102.175 μs. Though the execution time is still within 1 ms sampling time of the speed control loop, the proposed WFNN estimator is rather complicated. A simplified FNN model will be considered in the future to reduce the calculation burden of the DSP.

VI. CONCLUSION
A novel voltage control scheme using a feedforward voltage angle controller with online intelligent parameter estimation method for the q-axis inductance was successfully developed for an IPMSM servo drive in this study. First, an ABNC speed controller was developed to improve the transient response of the speed control. Then, the MTPA method is used in the current control mode for the constant torque region. Moreover, the feedforward voltage angle controller with intelligent parameter estimation using WFNN was designed to improve the dynamic response in the voltage control mode for the constant power region. Furthermore, the q-axis current controller was retained to ensure the steady-state response. In addition, the experimental results of three operating conditions transforming from the current control mode to the voltage control mode to achieve the constant power control at rated output power were provided to demonstrate the validity of the proposed control methods of the IPMSM servo drive system.

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