1. INTRODUCTION

Owing to the advantages of the superior power density, the high performance in speed control – dynamic response and better accuracy, permanent magnet synchronous motors (PMSMs) have been used in many automation control fields as actuators [1]. Conventional motor control needs a speed sensor or an optical encoder to measure the rotor speed and feedback it to the controller for ensuring the precision speed control. However, such sensor presents some disadvantages such as drive cost, machine size, reliability and noise immunity [2]. In recent year, a sensorless control for PMSM drive become an important issue and variable sensorless control strategies have been investigated [2-5], such as, the sliding mode observer, the back EMF, Kalman filter, the neural network, etc. However, the back EMF and the sliding mode observer are suitable to be implemented by the fix-pointed processor. In industrial applications, there are many uncertainties, such as system parameter uncertainties, external load disturbances, unmodeled uncertainties, etc. which always diminish the performance quality of the pre-design specification in the motor driving system. To solve this problem, intelligent control techniques [6-8] such as the fuzzy control, the neural networks control, the AFC, have been developed and applied to the speed control of servo motor drives to yield high operating performance.

With rapid developments of the system-on-chip, a high-performance digital signal processor (DSP) has become a popular area of research in the field of the digital control [9] for ac drives because they exhibit high-speed performance, combine peripheral circuits, memory and have an optimized CPU structure on a single chip. In particular, the DSP controller TMS320F2812 [10] produced by Texas Instrument, which provides a high speed (150MIPS) computational power, up to 128Kx16 flash, two sets (total 12 lines) of PWM outputs, two sets (total 4 lines) of QEP inputs, a 16-channels 12-bit A/D converter (200ns conversion time) and a 56-bits GPIO. Therefore, a complex control algorithm, such as the neural network control or the AFC, applied to servomotor to improve the dynamic performance become possible. Accordingly, in this work, a TMS320F2812 DSP controller used to realize a current vector control, a SVPWM (Space Vector Pulse Width Modulation) scheme, a flux angle and a rotor speed estimation scheme, an AFC, etc, has been developed for a sensorless PMSM drive. All software implemented in DSP is coded in C language. The excellent characteristic of the proposed DSP-based controller system makes the sensorless PMSM drive robust, well performance and easy to program.

2. CONTROLLER DESIGN OF SENSORLESS PMSM DRIVE

The architecture of the proposed speed controller applied in the sensorless PMSM drive is shown in Fig.1. In this figure, the functionalities of Park, Park, Clark, Clark, SVPWM, AD converter, PI current control, flux angle and rotor speed estimation, AFC etc. are all implemented in a DSP chip. Detailed design technologies are described as follows.

2.1 Mathematical model of a PMSM drive

PMSM is a coupled and nonlinear system. Under vector control in current loop, it will become a decoupled and linear system which likes DC motor. Generally, the mathematical model of the PMSM in d-q axis can be expressed by,
The block diagram of the proposed speed controller for a sensorless PMSM drive is shown in Fig. 1. The controller consists of a reference model, PI controller, Clarke and Park transformation, SVPWM, and current detection pulse signal detection of the encoder, etc. After using vector control (control $i_d$ to 0 in Fig.1), it will make the nonlinear and coupling characteristics of PMSM become decouple. Thus, the torque magnitude control of a PMSM is only need to control the q-axis current. The transformations between stationary a-b-c frame and synchronously rotating d-q frame are shown in Fig. 2, where $\hat{f}_e$ is a space vector refer to current, voltage or flux. The rotor speed of PMSM is synchronous with the stator electrical speed, thus we have $\omega_e = (P/2)\omega_r$ and $\theta_e = (P/2)\theta_r$, where $\omega_r$ is synchronous with the rotor speed, $\theta_r$ is the electrical angle and $P$ is the number of the magnetic pole. In Fig. 2, the coordinate transformations between stationary a-b-c frame, stationary $\alpha$-$\beta$ frame and synchronously rotating d-q frame are described as follows:

- **Clarke transformation:** stationary a-b-c frame to stationary $\alpha$-$\beta$ frame.
  \[
  f_r = \begin{bmatrix}
  2 & -1 & -1 \\
  3 & 3 & 3 \\
  1 & 1 & 1 \\
  \end{bmatrix} f_e
  \]

- **Modified Clarke** transformation: stationary $\alpha$-$\beta$ frame to stationary a-b-c frame.
  \[
  f_a = \begin{bmatrix}
  1 & 0 \\
  \frac{1}{\sqrt{3}} & \frac{1}{\sqrt{3}} \\
  \frac{-1}{\sqrt{3}} & -\frac{1}{\sqrt{3}} \\
  \end{bmatrix} f_{\alpha}
  \]

2.2 Vector control in current loop of a PMSM drive

The configuration of the current loop using vector control for a PMSM in Fig.1, includes PI controller, Clarke, modified Clarke$^{-1}$, Park, Park$^{-1}$ coordinate transformation, SVPWM, current detection pulse signal detection of the encoder, etc. After using vector control (control $i_d$ to 0 in Fig.1), it will make the nonlinear and coupling characteristics of PMSM become decouple. Thus, the torque magnitude control of a PMSM is only need to control the q-axis current. The transformations between stationary a-b-c frame, stationary $\alpha$-$\beta$ frame and synchronously rotating d-q frame are shown in Fig. 2, where $\hat{f}_e$ is a space vector refer to current, voltage or flux. The rotor speed of PMSM is synchronous with the stator electrical speed, thus we have $\omega_e = (P/2)\omega_r$ and $\theta_e = (P/2)\theta_r$, where $\omega_r$ is synchronous with the rotor speed, $\theta_r$ is the electrical angle and $P$ is the number of the magnetic pole. In Fig. 2, the coordinate transformations between stationary a-b-c frame, stationary $\alpha$-$\beta$ frame and synchronously rotating d-q frame are described as follows:

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  \end{bmatrix} f_e
  \]

- **Modified Clarke** transformation: stationary $\alpha$-$\beta$ frame to stationary a-b-c frame.
  \[
  f_a = \begin{bmatrix}
  1 & 0 \\
  \frac{1}{\sqrt{3}} & \frac{1}{\sqrt{3}} \\
  \frac{-1}{\sqrt{3}} & -\frac{1}{\sqrt{3}} \\
  \end{bmatrix} f_{\alpha}
  \]
2.3 The estimation of a rotor flux angle

A flux angle estimator constructed by a current observer, a bang-bang controller, a back EMF estimator and a flux angle estimator is shown in Fig. 3. The feedback currents $i_{a}$, $i_{b}$ and the voltage commands $v_{a}$, $v_{b}$ are the inputs and the estimated flux angle $\hat{\theta}$ is the output. When the estimated currents have a difference with the actual measured currents, the error will be used to drive the bang-bang controller. Then, the output of a bang-bang controller will regulate the back EMF value and force the estimated currents to follow the actual measured currents. From (4), the mathematic model of current observer is rearranged by the following equation

$$\frac{d}{dt}i = Ai_a + B(v_a - e_a)$$

with

$$A = \frac{R}{L}I_s, \quad B = \frac{1}{L}I_s, \quad L_s = 2L_m, \quad i_a = [i_a, i_b]^T,$$

$$v_a = [v_a, v_b]^T$$

and $e_a = [e_a, e_b]^T$. The $R_s$ and $L_m$ are winding resistance and inductance, respectively. The $I_2$ is a 2x2 unity matrix. Next, the sliding mode observer is used to estimate the current observer in Fig.3, and the formulation is expressed as

$$\frac{d}{dt}\hat{i} = A\hat{i}_a + B(v_a - \hat{e}_a + z)$$

(12)

Where $\hat{i}_a = [\hat{i}_a, \hat{i}_b]^T$, $\hat{e}_a = [\hat{e}_a, \hat{e}_b]^T$, $z = k \cdot \text{sgn}(\hat{i}_a - i_a)$

The sgn is defined as follows

$$\text{sgn} = \begin{cases} 1 & (\hat{i}_a - i_a) > 0 \\ 0 & (\hat{i}_a - i_a) = 0 \\ -1 & (\hat{i}_a - i_a) < 0 \end{cases}$$

(14)

and $k \in R^{2x2}$ is a gain matrix. In (13), the $z$ is the output gain of the bang-bang controller for use to compensate a value to reduce the error between estimated current and measured current. In Fig. 3, the back EMF $\hat{e}_a$ is used to compute the flux angle $\hat{\theta}$. To consider the noise effect, a low-pass filter is applied to estimate back EMF, as follows,

$$\frac{d}{dt}\hat{e}_a = -\omega_e\hat{e}_a + \omega_e(z - \hat{e}_a)$$

(15)

where $\hat{e}_a = [\hat{e}_a, \hat{e}_b]^T$, $\omega_e = 2\pi f_0$ and $f_0$ is the cut-off frequency of the filter.

In implementation, the above formulations in the continuous system have to transfer to the discrete system; therefore, the sliding mode current observer in (12) can be formulated to the difference equation by

$$\hat{i}_a(n+1) = F\hat{i}_a(n) + G(v_a(n) - \hat{e}_a(n) + z(n))$$

(16)

where $F = e^{\frac{\omega_e}{2}T_s}I_s$, $G = \frac{1}{R_s}(1 - e^{-\frac{\omega_e}{2}T_s})$ and $T_s$ is the sampling time. In addition, the difference equation of the back EMF estimator can also be expressed by

$$\hat{e}_a(n+1) = \hat{e}_a(n) + 2\pi f_0(z(n) - \hat{e}_a(n))$$

(17)

Once the back EMF is estimated, the flux angle can be computed as,

$$\hat{e}_a = -\lambda_s\omega_s\sin\hat{\theta}$$

(18)

$$\hat{e}_b = \lambda_s\omega_s\cos\hat{\theta}$$

(19)

and the magnet angle $\hat{\theta}$ can be directly computed by

$$\hat{\theta} = \tan^{-1}\left(\frac{-\hat{e}_b}{\hat{e}_a}\right)$$

(20)
Finally, a summary for estimating the magnet position is shown by the following design procedures:
Step1: Estimate of the current by an observer in (12).
Step2: Calculate the current error.
Step3: Use (13) to obtain the Z gain of the current observer.
Step4: Estimate the back EMF in (17).
Step5: Obtain in (20).

2.4 Adaptive fuzzy controller (AFC) in speed loop of PMSM drive

The structure of an AFC for PMSM drive is depicted in the dotted line of Fig. 1, it includes a fuzzy controller, a reference model and an adjusting mechanism. In Fig.1, the tracking error \( e \) and the error change \( \Delta e \) are defined as
\[
e(n) = \omega_{ref}(n) - \dot{\omega}_r(n), \quad (21)
\]
\[
\Delta e(n) = e(n) - e(n-1), \quad (22)
\]
and \( K_{se} e, K_{ae} \Delta e \) and \( u_0 \) are the input and output variable of fuzzy controller, respectively. The design procedure of the fuzzy controller is shown in the followings:

- Define the linguistic value as \{\( A_1, A_2, E \)\}, they are the symmetrical triangular membership function:

\[
\xi_m(x_1, x_2, x_3) = \begin{cases}
0 & x_1 \leq x_1^m - \frac{w_1^m}{2} \\
\frac{x_1 - x_1^m + w_1^m}{2} & x_1^m - \frac{w_1^m}{2} < x_1 < x_1^m \\
\frac{w_1^m}{2} & x_1^m < x_1 < x_1^m + \frac{w_1^m}{2} \\
0 & x_1 \geq x_1^m + \frac{w_1^m}{2}
\end{cases} \quad (23)
\]

where \( x_1 \) is input value, \( \xi_m(*) \) is output value, \( x_1^m, w_1^m \) are mean value and width of the triangular function, respectively.

- Derive \( M \) fuzzy control rules from the dynamic response characteristics as initial condition, such as, IF \( e^m \) is \( A_1^m \) and \( \Delta e^m \) is \( A_1^m \) THEN \( u^m \) is \( E^m \), \( m=1,2,..,M(24) \)

- Construct a fuzzy system with \( u_f(x|\emptyset) \) from those \( M \) rules using the singleton fuzzifier, product-inference rule, and central average defuzzifier method. Therefore, (24) is replaced by the following expression:

\[
u_f(x|\emptyset) = \frac{\sum_{m=1}^{M} c_m \prod_{i=1}^{3} \xi_m(x_i, x_2^m, x_3^m)}{\sum_{m=1}^{M} \prod_{i=1}^{3} \xi_m(x_i, x_2^m, x_3^m)} = \frac{\sum_{m=1}^{M} c_m \mu_m}{\sum_{m=1}^{M} \mu_m} \quad (25)
\]

where those \( c_1, c_2, ..., c_M \) are adjustable parameters.

The least means squares method is adopted to derive the fuzzy control law in Fig. 1. The main purpose of adjusting the parameters of the fuzzy controller is to minimum the sum of square errors (cost function) between the rotor speed and the output of the reference model. The instantaneous cost function is defined as:

\[
J(k+1) = \frac{1}{2} e_n(k+1)^2 = \frac{1}{2} [\omega_{ref}(k+1) - \dot{\omega}_r(k+1)]^2 \quad (26)
\]

and the parameters of \( c_j \) is adjusted with,

\[
\Delta c_j(k) \propto -\frac{\partial J(k+1)}{\partial c_j(k)} \quad (27)
\]

To derive the formulation of the adjusting parameters, \( c_j \), the mathematical model of the ac motor is represented by the difference equation. Firstly, assume \( T_L \) to be zero, and take Laplace transformation in (9)-(10) and then

\[
\frac{o_x(s)}{i_q'(s)} = \frac{K_i}{B_m} \frac{B_m/J_m}{s+B_m'/J_m} \quad (28)
\]

the bilinear transformation is applied and the difference equation of a PMSM drive system can be derived as

\[
\frac{o_x(k)}{i_q'(k)} = \frac{K_i}{B_m} \frac{1-e^{-\frac{B_m'T}{T}}}{1-e^{-\frac{B_m'T}{T}}} z^{-1} \quad (29)
\]

where \( z^{-1} \) is a back-shift operator, \( T \) is the sampling period.

The current command, \( i_q^* \) in (29), is the output of the fuzzy controller through an integrator, so that it can be represented with

\[
i_q^* = i_q(k-I) + K_u u_f(k) \quad (30)
\]

From (29) and (30), we have

\[
o_x(k+I) = A o_x(k) + B i_q^*(k-I) + BK_u u_f(k) \quad (31)
\]

with \( A = \exp(-B_m'T/T), B = K_l(1-A)/B_m' \)

Furthermore, according to the chain rule, the partial differential equation of \( J(k+1) \) in (26) can be rewritten as

\[
\frac{\partial J(k+1)}{\partial c_j} = -\alpha \frac{\partial e_n(k+1)}{\partial c_j} \frac{\partial e_n(k+1)}{\partial u_f(k)} \quad (32)
\]

where \( \alpha \) is learning rate. Furthermore, using (31) and (25), we have

\[
\frac{\partial e_n(k+1)}{\partial c_j} \propto \frac{\partial e_n(k+1)}{\partial u_f(k)} = BK_u \quad (33)
\]

and

\[
\frac{\partial e_n(k)}{\partial c_j} \equiv \frac{\partial e_n(k)}{\partial u_f(k)} \quad (34)
\]

then, substituting (33) and (34) into (32) and (27), the parameters \( c_j \) of fuzzy controller in (25) can be adjusted at each control sampling interval by the following expression.

\[
\Delta c_j(k) = -\alpha BK_u e_n(k) \sum_{m=1}^{M} c_m \mu_m \quad (35)
\]

3. EXPERIMENTAL SYSTEM AND RESULTS

3.1 Experimental system

The overall block diagram of sensorless PMSM control system is depicted in Fig. 1, and its experimental system is shown in Figs. 4–5. The experimental system includes a TMS320F2812 DSP controller, a voltage source IGBT inverter and a PMSM. The detailed description of system component is as follows.

- PMSM: The power of the PMSM is 2.2KW, 8 poles, the rating speed is 2000rpm, and the maximum torque is 1.1Kg-m. The parameters of the motor are: \( R_s = 0.63 \Omega, L_d = L_q = 2.77 \mu H \). The torque magnitude of the brake is adjustable in the range with 0.2~5N*m. An incremental
optical encoder (1000 ppr) attached to PMSM is used to recheck the accuracy and correctness of the estimated rotor speed.

- **Inverter**: The inverter has 6 sets of IGBT type power transistors. The collector-emitter voltage of the IGBT is rating 600V, the gate-emitter voltage is rating ±20V, and the collector current in DC is rating 25A and in short time (1ms) is 50A. The photo-IC, Toshiba TLP250, is used for gate driving circuit of IGBT. Input signals of the inverter are PWM signals from DSP chip.

- **DSP Controller**: The DSP controller TMS320F2812 produced by Texas Instrument, which provides the high speed (150MIPS) computational power, up to 128Kx16 flash, two sets (total 12 lines) of PWM outputs, two sets (total 4 lines) of QEP inputs, a 16-channels 12-bit A/D converter (200ns conversion time) and a 56-bits GPIO. In Fig.1, a current vector control scheme, a SVPWM generation, two A/Ds conversion, the coordinate transformation, a rotor speed estimation and an intelligent fuzzy control strategy are all realized by the software in DSP chip. The PWM switching frequency of inverter is designed with 16kHz, dead-band is 1μs, and the control sampling frequency of current and speed loop are 16kHz and 1kHz, respectively. The flow chart of main program and the interrupt service routine for digital motor control algorithm are designed and shown in Fig. 6. Those programs are coded with C language, the computation time in DSP for executing current loop and AFC algorithm of speed loop is 25μs.

3.2 Experimental results

In the implementation, the PWM frequency, dead-band, current loop sampling frequency, speed loop sampling frequency in PMSM drive are designed with 16kHz, 1μs, 16kHz and 500Hz, respectively. The cut-off frequency of low-pass filter for back EMF estimator in Fig. 3 is set to be 50Hz. In the experiment of the flux angle estimation, the running PMSM are tested in the condition of 60rpm (4Hz flux rotation frequency), 300rpm (20Hz) and 1200rpm (80Hz). To evaluate the correctness of the measured flux angle, an encoder attached to the PMSM is adopted to measure the rotor flux position for comparing the accuracy of the estimated value; and the experiment results are shown in Figs. 7–9. In Figs. 7–8, the PMSM runs at 60rpm and 300rpm and the estimated flux angle has 30° and 7° phase error compared with the measured value by encoder sensor, respectively. However, the PMSM runs at 1200rpm in Fig.9, the estimated flux angle almost match the measured value by encoder sensor. The experiment results show that the method of sliding mode current observer applied to estimate the flux angle is suitable for a moderate or a high speed motor running. At a low speed of the motor rotor, the weak back EMF signal will affect the accuracy of the flux position. After confirming the correctness of the sensorless flux estimation, and further, the vector control and AFC are applied to the current loop and speed loop, respectively, for improving the speed control performance. Then the 0.5Hz square wave with amplitude varying between 300-600, 600-900 and 900-1200rpm are used as the tested command signal. The experimental rotor speed responses corresponding with aforementioned command under without encoder sensor are shown in Fig. 10 to Fig.12. The results show that all the rotor speed track the output of reference model well after 1-2 period square wave command in different running speed condition of the PMSM. Thus, the experiments in Figs. 7–12 demonstrate that our proposed DSP-based sensorless PMSM drive control system using AFC is effectiveness and correctness.
Fig. 7. PMSM running at 60rpm (rotating frequency of flux position: 4Hz) (a)Estimated flux angle and measured flux angle from encoder (b)Error of flux angle

Fig. 8. PMSM running at 300rpm (rotating frequency of flux position: 20Hz) (a)Estimated flux angle and measured flux angle from encoder (b)Error of flux angle

Fig. 9. PMSM running at 1200rpm (rotating frequency of flux position: 80Hz) (a)Estimated flux angle and measured flux angle from encoder (b)Error of flux angle

Fig. 10. Speed step response by using the proposed sensorless and AFC control under the speed variation between 300-600rpm

Fig. 11. Speed step response by using the proposed sensorless and AFC control under the speed variation between 600-900rpm

Fig. 12. Speed step response by using the proposed sensorless and AFC control under the speed variation between 900-1200rpm

4. CONCLUSION

A DSP-based PMSM drive without the encoder sensor is successfully demonstrated in this work. The work herein is summarized as follows. Firstly, the functionalities required to build a sensorless PMSM drive, such as the A/D converter, the flux angle estimator, the sliding mode observer, the current vector control, the AFC, the SVPWM generation etc., have been integrated and realized in one TMS320F2812 DSP chip. Secondly, the flux angle estimator constructed by a current observer, a bang-bang controller, a back EMF estimator and a flux angle calculation, has been proven an effectiveness in the estimation of the rotor speed. Thirdly, the use of AFC in the speed control of PMSM drive has been demonstrated good performances under system uncertainty condition by some experimental results.

REFERENCES