Real-Time Flux-weakening Control for an IPMSM Drive System Using a Predictive Controller

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1. Introduction

Use of the interior permanent magnet synchronous motor (IPMSM) has greatly increased due to its advantages over other motors, which include high efficiency, wide operating speed range, high power density, and robustness. The IPMSM has two different torque components: electromagnetic torque and reluctance torque. By strategically controlling the d-axis and q-axis currents, the maximum torque per ampere (MTPA) control or flux weakening control can be achieved at different operating speeds.

This paper is an extension of work originally presented in 2017 IEEE 26th International Symposium on Industrial Electronics (ISIE) Conference [1]. Several researchers have proposed different flux-weakening control methods to increase the high-speed operating range of an IPMSM. For example, Bolognani et al. proposed an adaptive flux-weakening controller for an IPMSM drive system. The adaptation algorithm, however, is very complicated to implement [2]. Zhu et al. implemented a flux-weakening control for a PMSM drive system [3]. The proposed method accounted for resistance voltage drop and inverter nonlinearity. This method, however, is only suitable for a PMSM drive system, not an IPMSM drive system. Kwon et al. proposed a flux-weakening control for an IPMSM with quasi-six-step operation [4]. A feed-forward path, which consists of a one-dimensional look-up table, is required. Pan et al. proposed a robust flux-weakening control strategy for a surface-mounted permanent-magnet motor drive [5]. A closed-form solution of the maximum available torque-producing current is generated to achieve both fast responses and real-time tuning flux-weakening control. This method, however, is suitable for a surface mounted PMSM but not an IPMSM. Uddin et al. proposed an on-line parameter-estimation-based high-speed control of an IPMSM drive. The method of estimating the d-axis inductance, q-axis inductance, and external load, however, is very complicated [6]. Pan et al. proposed a voltage-constraint-based flux-weakening control of an IPMSM drive system [7]. The method requires a d-axis current command and q-axis current command generator, which is rather complicated. Jung et al. proposed a hexagonal voltage limit to increase the voltage amplitude and to improve the DC-link voltage utilization in an IPMSM drive system. A torque control method based on the voltage angle was investigated. The idea is very good; however, the voltage vector selection algorithm is very complicated [8]. Chaithongsuk et al. proposed an optimal design of permanent magnet motors to improve flux-weakening performance in variable speed drives. It was shown that a salient-pole motor performed better at high speeds than a non-salient pole motor [9]. Kim proposed a novel magnetic flux-weakening method for permanent magnet synchronous motors in electric vehicles [10]. By adjusting the air gap between the stator and rotor, an increased maximum speed and maximum output power could be reached. By adjusting the air gap between the stator and rotor, an increased maximum speed and maximum output power could be reached.

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A B S T R A C T

This paper proposes extended-range high-speed control for an IPMSM drive system. A simple real-time tuning flux-weakening control algorithm is proposed and implemented to control an IPMSM drive system in a wide variable speed range, from 3 r/min up to 2700 r/min. This flux-weakening control algorithm does not require any motor parameters and only needs simple mathematical computations. The proposed drive system adjusts the angle between the d-axis and q-axis current to reach flux-weakening control. In addition, a multiple sampling predictive controller is implemented to enhance the dynamic responses of the proposed drive system, which yields improved overall transient responses, superior load responses, and good tracking responses. A detailed analysis of the proposed drive system’s stability is discussed as well. A 32-bit digital signal processor, TMS-320F28335, is used to execute the predictive controller and the flux-weakening control algorithm for the IPMSM drive system. Experimental results can validate the theoretical analysis.
In this paper, a systematic predictive controller design is proposed for an IPMSM drive system. The proposed predictive controllers, including a predictive current controller and a predictive speed controller, provide a multiple sampling rate control system and a closed-loop speed-control block diagram that are different from the predictive controllers proposed in previously published papers [15]-[16], [18]-[20].

The design of a predictive controller requires the following. First, a model of the uncontrolled drive system is developed. Next, a cost function that represents the desired behavior of the system is defined. Finally, a predictive controller for the motor drive system is derived.

2.1. Predictive Current Controller

The state-variable equation of an IPMSM can be expressed as

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} -\frac{r_d}{L_d} & 0 \\ 0 & -\frac{r_q}{L_q} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \end{bmatrix} v_d + \begin{bmatrix} 1 \\ 0 \end{bmatrix} v_q - \omega_e L_{re} i_d - \omega_e (L_d j \omega_e + \lambda_s)$$  (1)

Where \(d/dt\) is the differential operator, \(i_d\) and \(i_q\) are the d-axis and q-axis stator currents, \(r_s\) is the stator resistance, \(v_d\) and \(v_q\) are the d-axis and q-axis stator voltages, \(L_q\) and \(L_q\) are the d-axis and q-axis self-inductances, \(\lambda_s\) is the permanent magnetic flux linkage, and \(\omega_e\) is the electric motor speed.

After transferring the continuous-time domain into the discrete-time domain, (1) can be expressed as

$$i_d(z) = \frac{\beta_d}{1 - \alpha_d z^{-1}}(v_d - \omega_e L_{re} i_d)$$  (2)

and

$$i_q(z) = \frac{\beta_q}{1 - \alpha_q z^{-1}}(v_q - \omega_e (L_d j \omega_e + \lambda_s))$$  (3)

The related parameters in (2) and (3) are shown as follows

$$\alpha_d = e^{-\frac{\omega_e T_s}{r_s}}$$  (4a)

$$\beta_d = \frac{1}{r_s}(1 - \alpha_d)$$  (4b)

$$\alpha_q = e^{-\frac{\omega_e T_s}{r_q}}$$  (4c)

$$\beta_q = \frac{1}{r_q}(1 - \alpha_q)$$  (4d)
where \( T_{cur} \) is the sampling interval of current-loop control. Because the sampling time interval is shorter than the time constant of the stator current-loop, Euler approximation can be used here. As a result, the parameters shown in (4a)-(4d) can be described as follows

\[
\alpha_{d} \approx 1 - \frac{r T_{cur}}{L_d} \quad (5a)
\]
\[
\beta_{d} \approx \frac{r T_{cur}}{L_d} \quad (5b)
\]
\[
\alpha_{q} \approx 1 - \frac{r T_{cur}}{L_q} \quad (5c)
\]
\[
\beta_{q} \approx \frac{r T_{cur}}{L_q} \quad (5d)
\]

Then (2) and (3) can be rewritten as follows

\[
X_{cur}(k + 1) = A_{cur} X_{cur}(k) + B_{cur} U_{cur}(k) + B_{cur} F_{cur}(k) \quad (6)
\]

and

\[
X_{cur}(k) = \begin{bmatrix} i_d(k) \\ i_q(k) \end{bmatrix}^T \quad (7a)
\]
\[
F_{cur}(k) = \begin{bmatrix} \omega_r(k)L_q i_q(k) - \omega_r(k)(L_d i_d(k) + \lambda_m) \end{bmatrix}^T \quad (7b)
\]
\[
U_{cur}(k) = \begin{bmatrix} v_d(k) \\ v_q(k) \end{bmatrix}^T \quad (7c)
\]
\[
A_{cur} = \begin{bmatrix} 1 - \frac{r T_{cur}}{L_d} & 0 \\ 0 & 1 - \frac{r T_{cur}}{L_q} \end{bmatrix} \quad (7d)
\]
\[
B_{cur} = \begin{bmatrix} \frac{T_{cur}}{L_d} & 0 \\ 0 & \frac{T_{cur}}{L_q} \end{bmatrix} \quad (7e)
\]

According to (6), the (k+1)-th predictive current can be expressed as

\[
X_{cur}(k + 1) = A_{cur} X_{cur}(k) + B_{cur} (U_{cur}(k) + F_{cur}(k)) \quad (8)
\]

The performance index of the d-axis current control is defined as

\[
J_{cur} = \left[ i_d(k + 1) - i_d(k + 1) \right]^2 \quad (9)
\]

The following equation is the partial differential of (9) manipulated to equal zero.

\[
\frac{\partial J_{cur,d}}{\partial U_d} = \frac{\partial \left( i_d(k + 1) - i_d(k + 1) \right)^2}{\partial v_d(k)} = 0 \quad (10)
\]

Substituting (8) into (10), we can obtain

\[
\frac{\partial J_{cur,d}}{\partial v_d(k)} = \frac{\partial \left(1 - \frac{T_{cur}}{L_d} \right) i_d(k) \left( + \frac{T_{cur}}{L_d} \right)}{\partial v_d(k)} \left( v_d(k) + \omega_r(k)L_q i_q(k) - i_d(k) \right) - i_d(k + 1) \right)^2 = 0
\]

After arranging the above, it is not difficult to obtain

\[
\left(1 - \frac{T_{cur}}{L_d} \right) i_d(k) + \left( \frac{T_{cur}}{L_d} \right) \left( v_d(k) + \omega_r(k)L_q i_q(k) - i_d(k) \right) - i_d(k + 1) = 0 \quad (12)
\]

Finally, from (12), we can derive the d-axis voltage command as

\[
v_d(k) = r_i(k) \left( i_d(k + 1) - i_d(k) \right) - \omega_r(k)L_q i_q(k) \quad (13)
\]

By using the same principle, we can derive the q-axis voltage command as follows

\[
v_q(k) = r_i(k) \left( i_q(k + 1) - i_q(k) \right) + \omega_r(k)(L_d i_d(k) + \lambda_m) \quad (14)
\]

2.2. Predictive Speed Controller

The differential equation of an IPMSM is

\[
\frac{d}{dt} \omega_m = \frac{1}{J_m} (T_e - T_L - B_m \omega_m) \quad (15)
\]

Where \( J_m \) is the inertia, \( \omega_m \) is the mechanical speed, \( \frac{d}{dt} \) is the differential operator, \( T_e \) is the total output torque, \( T_L \) is the external load, and \( B_m \) is the viscous frictional coefficient.
According to (15) and assuming $T_L = 0$, the speed can be expressed as

$$
\omega_m(s) = \frac{T_r}{J_m s + B_m} \quad (16)
$$

By converting (16) into discrete form, one can obtain

$$
\omega_m(k+1) = e^{-T \nu} \omega_m(k) + \frac{1}{B_m} \left( 1 - e^{-T \nu} \right) T_r(k)
$$

and

$$
a_i = e^{-T \nu}
$$

$$
b_i = \frac{1}{B_m} \left( 1 - e^{-T \nu} \right)
$$

$$
\omega(k) = T_r(k)
$$

where $T_{\nu}$ is the sampling interval of the speed-loop control. Next, by setting the predictive window as 1 and the predictive control as 1, the performance index $J_{sp}$ is expressed as

$$
J_{sp} = \sum_{i=1}^{1} [P(z)\omega_m^{\nu}(k+i) - P(1)\omega_m^{\nu}(k+i)] - P(1)\omega_m^{\nu}(k+1) + \rho \sum_{i=0}^{1} [Q(z)u(k+i-1)]^2
$$

where $\omega_m^{\nu}$ is the predictive speed, $P(z)$ and $Q(z)$ are polynomials of $z$, and $\rho$ is the weighting factor. By computing $\frac{\partial J_{sp}}{\partial u(k)}$, one can obtain

$$
\frac{\partial J_{sp}}{\partial u(k)} = 2 P(z)\omega_m^{\nu}(k+1) - P(1)\omega_m^{\nu}(k+1) + \rho \sum_{i=0}^{1} [Q(z)u(k+i-1)]^2 + 2 \rho Q^2(z)u(k) = 0
$$

Substituting (17) into (22), one can derive

$$
P(z)h_{1} \left( P(z)\left( h_{1}u(k) + a_i\omega_m(k) \right) - P(1)\omega_m^{*}(k+1) \right) + \rho Q^2(z)u(k) = 0
$$

The speed command at (k+1)th sampling interval can be expressed as a fixed command $S_p$ and then can be defined as

$$
\omega_m^{*}(k+1) = S_p
$$

By substituting (24) into (23), we can obtain

$$
(P(z)h_{1} \hat{y}(k) + \rho Q^2(z)u(k)) = P(z)h_{1}P(1)S_p - P^2(z)h_{1}a_i\omega_m(k)
$$

The control input can be expressed as

$$
u(k) = R(\omega_m - N\omega_m(k))
$$

where

$$
R = \frac{1}{1 + \frac{1}{b_i} \frac{Q(z)}{P(z)}^2}
$$

The predictive controller is shown in Figure 1. The block diagram of the predictive control includes a command gradation $M(z)$, a controller $R(z)$, a plant $P(z)$, and a feedback gain $N$. The structure is very similar to a standard control system; however, the parameters can be systematically derived. Unlike [16], this paper outlines a systematic closed-loop block diagram of the predictive controller, which provides a clear feedback control structure for an IPMSM drive system. This is an unprecedented method of predictive control and also a major aspect of this paper.

![Figure 1 Block diagram of predictive speed-loop control.](image)

2.3. Stability Analysis of the Proposed Drive System

The desired torque command of an IPMSM can be generated by the following equation

$$
\frac{d}{dt} \omega_m^* = \frac{1}{J_m} \left( T_e^* - T_i - B_m \omega_m^* \right)
$$

In the real world, there are parameter variations of the IPMSM and the external load measuring error. To compensate for that, the total equivalent load with uncertainty can be expressed as $T_{eq}$, which includes the summation of the external load, the load influenced...
by the parameter variations, torque tracking errors of both MTPA and flux weakening, and external load measuring error. Then (30) can be rewritten as

\begin{equation}
\frac{d}{dt}\omega_m^* = -\frac{B_m}{J_m}\omega_m^* + \frac{1}{J_m}T_e - T_D
\end{equation}

and

\begin{equation}
T_D = \frac{1}{J_m}(\Delta J_m \frac{d}{dt}\omega_m^* + \Delta B_m\omega_m^* + T_L)
\end{equation}

where \(T_D\) is the total load disturbance, \(\Delta J_m\) is the variation of inertia, and \(\Delta B_m\) is the variation of the viscous coefficient. Next, the dynamic speed equation of the IPMSM is

\begin{equation}
\frac{d}{dt}\omega_m^* = -\frac{B_m}{J_m}\omega_m^* + \frac{1}{J_m}T_e - \tilde{T}_D
\end{equation}

Where \(\tilde{T}_D\) is the estimated total load disturbance, which includes the influence of the motor parameters’ uncertainty. The tuning step of the \(\theta_{ad}^*\), which is 0.18°, creates a varying torque that is less than 1% of the output torque \(T_e\). As a result, the influence of the uncertainty is bound. The speed error, which is the difference between the speed command and the real speed, can be expressed as

\begin{equation}
e = \omega_m^* - \omega_m
\end{equation}

Taking the differential of the speed error, one can obtain

\begin{equation}
\dot{e} = \frac{d}{dt}(\omega_m^* - \omega_m) = \frac{d}{dt}\omega_m^* - \frac{d}{dt}\omega_m
\end{equation}

Substituting (31) and (33) into (35), one can obtain

\begin{equation}
\dot{e} = \left(-\frac{B_m}{J_m}\omega_m^* + \frac{1}{J_m}T_e - T_D\right) - \left(-\frac{B_m}{J_m}\omega_m^* + \frac{1}{J_m}T_e - \tilde{T}_D\right)
\end{equation}

\begin{equation}
= -\frac{B_m}{J_m}(\omega_m^* - \omega_m) - \tilde{T}_D
\end{equation}

We can define the load estimation error as \(\tilde{T}_D = T_D - \tilde{T}_D\). Assuming the external load changes slowly when compared to the quick responses of the speed control loop, it is possible to let \(\tilde{T}_D = 0\). The differential of the estimated load, therefore, can be expressed as follows

\begin{equation}
\tilde{T}_D = -\tilde{T}_D
\end{equation}

Define a lyapunov function as

\begin{equation}
V = e^2 + \frac{1}{k_1}(\hat{T}_D)^2
\end{equation}

where \(k_1\) is a weighting factor. By taking the derivative of the Lyapunov function, one can derive

\begin{equation}
\dot{V} = 2e\frac{d}{dt}e + \frac{2}{k_1}\hat{T}_D \dot{\hat{T}}_D
\end{equation}

Substituting (36) into (39), one can obtain

\begin{equation}
\dot{V} = 2e\left(-\frac{B_m}{J_m}e - \tilde{T}_D\right) + \frac{2}{k_1}\hat{T}_D \dot{\hat{T}}_D = 2\frac{B_m}{J_m}e^2 + 2\hat{T}_D \left(\frac{1}{k_1}\hat{T}_D - e\right)
\end{equation}

According to (40), it is possible to choose an adaption law as follows

\begin{equation}
\dot{T}_D = k_1e
\end{equation}

Substituting (41) into (40), we can obtain the differential of Lyapunov function as

\begin{equation}
\dot{V} = 2\frac{B_m}{J_m}e^2 \leq 0
\end{equation}

From (42), the differential of Lyapunov function is negative semi-definite. Then, Barbalet lemma can be applied. First, a function \(y_1\) is set, which is equal to \(\dot{V}\). Next, by integrating the \(y_1\), we can obtain

\begin{equation}
\int_0^\infty y_1(\tau)d\tau = V(0) - V(\infty) < \infty
\end{equation}

From (43), we can conclude

\begin{equation}
\lim_{t \to \infty} y_1(t) = 0
\end{equation}

According to (44), the speed error converges to zero as time approaches infinity.

3. MTPA Control and Flux-weakening Control

3.1. Basic Principle

The dynamic d-q axis voltages of an IPMSM can be expressed as

\begin{equation}
v_d = r_l i_d + L_d \frac{di_d}{dt} - \omega_n L_s i_q
\end{equation}

and

\begin{equation}
v_q = \frac{1}{2}\omega_n L_s i_d + L_d \frac{di_q}{dt}
\end{equation}
The total output torque of an IPMSM is

\[ T_e = \frac{3 P}{2} \left[ \lambda_m - (L_d - L_q) i_d \right] i_q \]
\[ = \frac{3 P}{2} \left[ \lambda_m i_q - (L_d - L_q) \sqrt{i_q^2 - i_d^2} \right] i_q \]  

(47)

Where \( I_s \) is the current amplitude. Two major control algorithms can be derived from (47) and can be used to adjust the torque of an IPMSM: field-oriented control and MTPA control. The field-oriented control sets a fixed d-axis current and adjusts the q-axis current to obtain linear proportional torque. The MTPA control sets a variable d-axis current, which is related to the q-axis current, to obtain maximum torque/ampere. The method can generate more torque than field-oriented control. However, the method creates a nonlinear relationship between the torque and q-axis current [3], which is a disadvantage of MTPA control. While it is a fact that MTPA’s influence on torque linearity is a drawback, achieving maximum torque is a greater benefit that outweighs that disadvantage. The issue will be researched further by the authors of this paper.

Taking \( \frac{d}{di_q} T_e = 0 \) and substituting it into (47), we can obtain

\[ \frac{3 P}{2} \left[ \lambda_m + (L_d - L_q) i_d - (L_d - L_q) i_q^2 \right] = 0 \]

(48)

From (48), the d-axis current under MTPA control can be expressed as

\[ i_d MTPA = \frac{-\lambda_m}{2(L_d - L_q)} - \frac{\lambda_m^2}{4(L_d - L_q)^2} + i_q^2 \]

(49)

The advance angle under MTPA control is

\[ \theta_{MTPA} = \sin^{-1} \left( \frac{i_d MTPA}{I_s} \right) \]

(50)

and the current amplitude \( I_s \) is expressed as

\[ I_s = \sqrt{i_q^2 + i_d^2} \]

(51)

In steady-state and neglecting the voltage drops of the resistance and inductance, it is not difficult to obtain

\[ v_d = -\omega_re L_q i_q \]

(52)

The voltage amplitude is

\[ v_s = \sqrt{v_d^2 + v_q^2} \]

(54)

If \( V_{om} \) is the maximum voltage amplitude, the voltage constraint can be shown as

\[ (L_d i_q)^2 + (L_d i_d + \lambda_m)^2 = \left( \frac{V_{om}}{\omega_re} \right)^2 \]

(55)

The maximum current amplitude \( I_{sm} \) can be expressed as

\[ i_q^2 + i_d^2 = I_{sm}^2 \]

(56)

From (56), it is easy to obtain

\[ i_q = \sqrt{I_{sm}^2 - i_d^2} \]

(57)

Combining (49), (54), and (57), the constraints of the IPMSM drive system can be obtained. The drive system operates under MTPA control when the motor speed is below the rated base-speed. After the drive system reaches its base speed, there are two constraints: the current constraint of the maximum amplitude, \( I_{sm} \), and the voltage constraint of the maximum voltage amplitude, \( V_{om} \). However, the \( \frac{V_{om}}{\omega_re} \) is decreased as the motor speed increases, which is shown in Figure 2. As a result, the axes of the ellipse are reduced when the motor speed increases. When \( \omega_re < \omega_{base} \), the motor operates under MTPA control, which is indicated as OCBA in Figure 2. When \( \omega_{base} < \omega_re < \omega_{cr} \), the motor operates in region II, in which the motor can operate either under MTPA control or flux-weakening control. If \( i_d < i_d MTPA \), MTPA control is used; on the other hand, if \( i_d > i_d MTPA \), flux-weakening control is used.

When \( \omega_re < \omega_{cr} \), the motor operates in region III, in which only flux-weakening control can be used.

3.2. Proposed Real-Time Flux-weakening Control

In the real world, flux-weakening control is very complicated because it requires \( \lambda_m, L_d, \) and \( L_q \), which are very difficult to measure [15].

To solve this difficulty, in this paper, a real-time tuning flux-weakening control is proposed. First, the voltage amplitude of the IPMSM, \( v_q \), which is expressed in (54) is compared to the maximum voltage amplitude, \( V_{om} \). The \( \Theta_{FV} \), which is shown in Figure 3(a), is tuned as follows
4. Implementation

The implemented IPMSM drive system is shown in Figure 5(a) and (b). Figure 5(a) is a block diagram of the proposed IPMSM. The system consists of two main parts: the DSP system and the hardware circuit. The DSP system uses a TMS320F28335, which is a 32-bit, digital signal processor manufactured by Texas Instruments. The sampling interval of the current-loop control is 100 μs, while that of the speed-loop control is 1 ms.

Taking the MTPA operating region as an example, the controller works as follows. First, as shown in Figure 5(a) and based on the predictive speed controller, after the speed command is input and the motor speed is feedback, the q-axis current control $i_q^*$ is computed. Second, the load compensator compensates for the load disturbance $\tilde{T}_d$ and then generates the compensation current $i_{q_T}$. Then the total q-axis current command, which is the summation of the $i_q^*$ and $i_{q_T}$, is computed and expressed as $i_q^*$. Next, the d-axis current is computed through (49) when the motor is operating in the MTPA region. After that, the advance phase angle $\theta_{\text{MTPA}}$ and stator current amplitude $I_y^*$ are computed by using (50) and (51). Finally, the flux-weakening is obtained by adding $\theta_{\text{FW}}$.

By using the MTPA table and the real-time tuning flux-weakening control, the advance angle of the current can be obtained. After that, the d-axis current command $i_d^*$ and the q-axis current command $i_q^*$ are calculated. A current regulated control and a space vector PWM modulation are executed. Finally, the switching states of the inverter are determined and output. A closed-loop system can thus be obtained. A DC motor is coupled with the IPMSM. The motor parameters are: $L_s = 31 \text{ mH}$, $L_q = 15 \text{ mH}$, $\lambda_s = 0.227 \text{ V.s/deg}$, and $R_s = 1.9 \Omega$. The motor is 4-pole, 1-HP, and rated at 1500 r/min.

Figure 5(b) shows a photograph of the entire hardware circuit, which includes a DSP, gate drivers, peripheral devices, current sensing circuits, and an inverter. Figure 5(c) shows a photograph of the implemented IPMSM drive system, including an IPMSM and a dynamometer.

5. Experimental Results

The performance of the proposed flux-weakening control for an IPMSM drive system was evaluated according to experimental results. Figure 6(a) shows the measured steady-state torque-speed curves using the zero d-axis current control and the MTPA control. Figure 6(b) shows measured steady-state torque-speed curves.
using the MTPA control and the hybrid control, which is a combination of the MTPA control and the flux-weakening control.

![Block diagram of IPMSM drive system](image1.png)

As you can observe, the hybrid control has better performance than the MTPA control. Figure 6 (c) shows the q-axis current to the d-axis current trajectory when the IPMSM is operating from the MTPA control region to the flux-weakening control region. The d-axis current increases slightly in the MTPA region; however, the q-axis current increases significantly in the flux-weakening region.

Figure 7(a)(b) and (c) show the measured waveforms of the IPMSM operating at 1800 r/min under loads of 0.57 N-m and 1.89 N-m. Figure 7(a) is the amplitude of the stator current $I_s$. Figure 7(b) shows the relative measured d-axis and q-axis currents. Figure 7(c) shows the measured advance angle $\theta_{adv}$ that is the sum of the $\theta_{FW}$ and $\theta_{MTPA}$, which are the measured flux-weakening angle and the measured MTPA control angle. The delay of the measured responses is due to the lag between the load command and the external load that is generated by the dynamometer. Figure 8(a) shows the load disturbance responses when using the proposed predictive controller, the previous predictive controller that was proposed by [21], and the PI controller. The parameters of the PI controller are obtained by using the pole assignment technique. The locations of the desired poles for the closed-loop drive system are assigned as -4.06+j0.13 and -4.06-j0.13, respectively. The proposed predictive controller performs better than the previous predictive controller and the PI controller. At this 2000 r/min high speed, the field weakening algorithm is applied. The maximum allowed external load at this operating speed is 1.6 N.m. Experimental results show the proposed predictive controller performs the best. Figure 8(b) shows the repetitive 1.6 N.m load disturbance responses. The proposed predictive controller performs well again.

Figure 9(a) shows the measured step-input transient speed responses at the maximum speed, 2700 r/min. In this paper, the rated speed is 1500 r/min. The measured responses show the proposed predictive controller has a quicker rise time than the previous predictive controller proposed by [16] and the PI controller. Figure 9(b) shows the measured step-input transient speed responses at 3 r/min, which is the minimum operating speed. According to Figure 9(a) and Figure 9(b), we can conclude that the adjustable speed ratio of the maximum speed to minimum speed is 900:1. Figure 10 (a)(b)(c)(d) shows the current command and measured current errors using different controllers. Again, the proposed predictive controller has the smallest tracking errors. Figure 11(a)(b)(c) show the measured dynamic responses of the proposed flux-weakening control at 1800 r/min. A 5V step input at $V_{om}$ is injected for every 1 second. The real voltage amplitude of the motor, which is expressed as $V_r$, can track the voltage command $V_{om}$ well. The rise time of the voltage control is about 50 ms. Figure 11(a), Figure 11(b) and Figure 11(c) show the responses of speed, voltage amplitude, and controlling angle $\theta_{FW}$ respectively. Figure 12(a) shows the measured efficiency of the whole drive system. The proposed method has higher efficiency than the linear flux-weakening method that was proposed by [21]. The major reason is that the proposed method provides greater output torque at the same motor speed. Figure 12(b) and (c) show the measured speed responses as the inertia of the drive system is increased to 5 times. The proposed predictive controller can provide a lower overshoot and a shorter settling time as the inertia increases.

![Graph showing the measured step-input transient speed responses](image2.png)

![Graph showing the measured dynamic responses](image3.png)
Figure 6 Measured steady-state characteristic curves (a) MTPA and zero d-axis current (b) MTPA and hybrid control (c) q-axis current to d-axis current curve.

Figure 8 Measured field weakening operating speed with a large 1.6 N.m external load. (a) step-input load (b) repetitive loads.

Figure 7 Measured 1800 r/min with loads of 0.57 N-m and 1.89 N-m (a) $I_d$ (b) measured d-q axis currents (c) advance angle.

Figure 9 Measured step-input transient speed responses (a) highest speed (b) lowest speed.
Figure 10 Comparison of measured current responses using different controllers
(a) current command (b) proposed predictive (c) previous predictive (d) PI.

Figure 11 Measured dynamic responses of the flux-weakening control at 1800 r/min
(a) speed (b) voltage magnitude regulation (c) $\theta_{FW}$.

Figure 12 Efficiencies of the flux-weakening control
(a) Linear flux-weakening (b) Proposed

$\omega_{in}$ (r/min)
Figure 12 Measured efficiency and varied inertia speed responses. (a) efficiency (b) proposed predictive controller (c) PI controller.

6. Conclusions

In this paper, an MTPA control and a real-time flux-weakening control drive system using a predictive controller is proposed to improve the adjustable speed range and dynamic responses of an IPMSM. Experimental results show that the proposed drive system has good performance, including fast transient responses, good load disturbance rejection capability, and good tracking responses. Furthermore, the proposed method is easily implemented by using a DSP. Experimental results validate the theoretical analysis.

References


