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Control for induction motor drives using predictive model stator currents and speeds control

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ABSTRACT

This paper is presented for designing a new controller using the predictive model current and speed control method for the asynchronous motor. This control method is based on traditional predictive controller development to have a cascade structure similar to the rotor flux control (field-oriented control) and direct torque control (DTC). Therefore, this control method will have two control loops. Both inner and outer loop controllers use predictive power. The outer ring is speed control, while the internal circle is stator current control. The inner loop is based on the finite control set – model predictive control (FCS-MPC), while the outer ring to take full advantage of the high dynamic response of the inner circle uses the deadbeat MPC. MATLAB simulation results show that this control method has performance equivalent to traditional controllers while minimizing overshoot and having fast, on-demand response times.

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1. INTRODUCTION

In industrial and transportation, asynchronous motors are becoming more used. The fundamental reason for this is because this motor has appropriate mechanical qualities for loads like electric automobiles, cranes, pumps, and blowers [1]–[3]. Furthermore, this motor is improved and enhanced when controlled by a nonlinear and intelligent control approach using a converter to regulate stator currents and a speed controller [4], [5]. Two methods have dominated the field of high-performance ac drives so far. The first is field-oriented control (FOC), which generates gate signals for power transistors using linear controllers and pulse-width modulators [6]-[8]. The second extensively used technology is direct torque control (DTC), which recognizes the converter's discontinuous character and generates pulses using comparators with hysteresis [9], [10].

Predictive control has emerged as one of the important achievements in control engineering in recent years [11], and [12]. This approach is beneficial for electrical engineering because it combines solid prediction models with powerful microprocessors that can execute many computations at a low cost. Predictive control is gaining popularity in power electronics and drives, with encouraging results [13], 14]. In some converter topologies operating with FOC, finite-state predictive control has been employed to regulate the current [15], [16]. In the so-called predictive torque control (PTC) [17], also known as finite set model predictive control, predictive power has also been utilized to regulate the torque of AC machines (FS-MPC). The speed loop is likewise accomplished using predictive model approaches in [18] and [19]. These methods use the traditional cascaded structure of an outer speed controller that feeds the torque (current) reference to an inner current control loop. However, by employing a single controller for all variables: speed, flux, and current, it is feasible to avoid the cascaded structure [20], [21] describe preliminary research in this approach for the permanent-magnet synchronous and induction motor machine (PMSM). As a result, the research will focus on the design and control of induction motors utilizing predictive model current control and a speed controller.

To begin with, with the advancement of fast and powerful microprocessors, predictive model control (MPC) in power electronics has gotten a lot of attention. The core notion is that calculate the system's future behavior to acquire optimum values for the actuating variables. Predictive control may be applied to various approaches using this fundamental notion, in which restrictions and nonlinearities can be readily added, multivariable cases can be examined, and the resultant controller is simple to implement. These characteristics make the technique appealing and successful for power electronics system management, including drive control, particularly predictive torque control (PTC) [22], [23]. Predictive torque control is the topic of our study. The finite control set – model predictive control is used in the inner loop (FCS-MPC). Predictive torque control works by determining the optimum switching state ahead of time to maintain state variables near the intended values. To eliminate the requirement for a speed sensor, the stator flux that determines the stator voltage is pre-calculated utilizing predictive torque control.

The synchronous speed of an induction motor is determined by the frequency and poles of the machine. Because the revolving magnetic field formed in the stator generates flux, which helps the rotor revolve, the engine in an installation always operates at a slower speed than synchronous. Despite this, the rotor will never attain the synchronous speed of the spinning magnetic field owing to the lag of flux current in the rotor with flux current in the stator. Many industrial applications need induction motor control approaches because they operate at a rated speed [24]–[27]. The outer ring is the focus of our project, which uses the deadbeat MPC to take full use of the inner circle's dynamic solid responsiveness.

To design a controller that predicts stator current and speed in part 3. First, we need a model of the discontinuous state of current, magnetic flux, and speed is required, as shown in part 2. In addition, to evaluate the effectiveness of the solution to export, the simulation results are evaluated and compared in part 4. Finally, the analysis, comments, and directions for the development of the research direction are shown in the conclusion section.

2. CURRENT AND MAGNETIC FLUX PREDICTION MODEL ON DISCONTINUITY DOMAIN

Based on [7], equation of stator current on stator coordinates mode in (1).

$$\frac{d\boldsymbol{i}_s}{dt} = -\frac{1}{\tau_\sigma}\boldsymbol{i}_s + \frac{k_r}{r_\sigma \tau_\sigma}\left(\frac{1}{\tau_r} + j\omega\right)\psi_r - \frac{1}{r_\sigma \tau_\sigma}u_s \tag{1}$$

Where

$\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is total leakage factor

$\tau_r = \frac{L_r}{R_r}$, $\tau_s = \frac{L_s}{R_s}$ is rotor, stator time constant

$r_\sigma = R_s + R_r k_r^2$, $\tau_\sigma = \frac{\sigma L_s}{r_\sigma}$, $k_r = \frac{L_m}{L_r}$ are calculation coefficients

u_s, ψ_r are voltage and flux

Equation of magnetic flux and torque mode are presented in (2).

$$\begin{aligned} &\psi_{rd} + \tau_r \frac{d\psi_{rd}}{dt} = L_m i_{sd} \\ &(\omega_s - \omega)\psi_{rd}\tau_r = L_m i_{sq} \\ &T_e = \frac{3}{2} p \frac{L_m}{L_r}\psi_{rd} i_{sq} \end{aligned} \tag{2}$$

Where L_m is Mutual inductance; $i_{sd} i_{sq}$ are dq components of the rotor current; T_e is torque.

The instantaneous position of the rotor flux vector is shown in (3).

$$\vartheta_r = \int \left(\frac{L_m}{\tau_r}\frac{i_{sq}}{\psi_{rd}} + \omega\right)dt \tag{3}$$

Where: ω is mechanical rotor velocity. The system of equations for estimating the discontinuous state flux on the axis of rotation *dq*.

$$\begin{aligned} \psi_{rd}(k+1) &= \frac{T}{\tau_r} L_m i_{sd}(k) + \left(1 - \frac{1}{\tau_r}\right)\psi_{rd}(k) \\ \psi_{rd}(k) &= \frac{L_m}{\tau_r(\omega_s - \omega)} i_{sq}(k) \end{aligned} \tag{4}$$

3. DESIGN OF THE STATOR CURRENT AND THE SPEED CONTROLLER

3.1. Design steps of the predictive controller for an IM

To design a controller that predicts the speed and current to control the IM motor, we follow these four steps sequentially:

- Step 1: Modelling the system, including a 3-phase 2-level inverter.
- Step 2: Calculate the predictive model for the loop current, flux, and speed at times [k+1] and [k+2] if necessary.
- Step 3: From the forecasting model obtained in step 2, we calculate the cost function so that the system works according to the design desired.
- Step 4: Find the vector as the minor cost function, apply that vector to switch IGBT valves to control the motor.

In the study of this article, the two-level inverter is powered for the IM motor with eight switching states to have a voltage vectors.

3.2. The current and flux prediction model

Using a Euler approximation with a sampling time T_s in (1) and (2), thus current and flux prediction model is shown in (5), and in (6):

$$i_s(k+1) = \left(1 - \frac{T_s}{\tau_\sigma}\right) i_s(k) + \frac{T_s}{\tau_\sigma}\frac{k_r}{r_\sigma}\left(\frac{1}{\tau_r} - j\omega(k)\right)\psi_r(k) + v_s(k) \tag{5}$$

$$\psi_r(k+1) = \left(1 - \frac{T_s}{\tau_r}\right)\psi_r(k) + \frac{T_s L_m}{\tau_r} i_s(k) + T_s j\omega\psi_r(k) \tag{6}$$

3.3. The speed predictive controller-MPC

The speed prediction control loop is an alternative to the classical PI speed controller. The objective of the speed controller is to obtain the appropriate perpendicular stator current based on the set speed for the minimum time. The outer control loop is calculated based on the kinematics equations given in (7). The speed control loop can be down sampled because the mechanical dynamics are limited by the inertia of the system and the ability to trigger maximum torque. The sampling time used for the outer loop is ten times that of the inner control loop (T_{ds} = 10*T_s). This allows the effect of the internal current control loop to be bypassed with the speed control loop.

$$\frac{d\omega}{dt} = \frac{1}{J}(T_e - T_L) \tag{7}$$

Where: T_e is the electric torque of the motor, TL is the load torque.

The electric torque T_e is determined as:

$$\frac{d\omega}{dt} = \frac{1}{J}\left(\frac{3}{2} p \frac{L_m}{L_r}\psi_{rd} i_{sq} - T_L\right) \tag{8}$$

applying Euler approximation with sampling time T_{ds}, we have.

$$\frac{\omega^{k+1} - \omega^k}{T_{ds}} = \frac{1}{J}\left(\frac{3}{2} p \frac{L_m}{L_r}\psi_{rd} i_{sq} - T_L\right) \tag{9}$$

The Euler approximation produces significant model errors when operating at high frequencies, thereby causing control problems. To obtain a more precise approximation of the reference current i_{sq}^*, we perform and order expansion of the rotor speed:

$$\omega^{k+1} = \omega^k + T_{ds}\frac{d\omega}{dt} + \frac{T^2{}_{ds}}{2}\frac{d^2\omega}{dt} \tag{10}$$

with:

$$\frac{d^2\omega}{dt^2} = \frac{1}{J}\left(\frac{3}{2}p\frac{L_m}{L_r}\frac{d\psi_{rd}}{dt}i_{sq} + \frac{3}{2}p\frac{L_m}{L_r}\psi_{rd}\frac{di_{sq}}{dt} - \frac{dT_L}{dt}\right) \tag{11}$$

where: $a = \frac{1}{J}p\frac{L_m}{L_r}$; T_L is assumed to be invariant over the sampling period. Substituting in (10) and (11) into in (9) and backward Euler approximation for the derivatives at in (12), we get an expression to compute the value of i_{sq}. The value of i_{sq}^k is used as the reference current in (13), $i_{sq}^* = i_{sq}^k$, which is delayed by one sampling cycle T_s. However, this delay has no effect on the output rate of the control loop when it is sampled with a period T_{ds}=10*T_s. It is noteworthy that the outer loop produces no significant delay because the reference current is calculated and used by the inner control loop at the time of first sampling T_s in T_{ds}.

$$i^s{}_{sq} = \frac{\omega^* - \omega^k + \frac{3a}{4}T_{ds}\psi_{rd}^k i_{sq}^{k-1} + \frac{T_{ds}}{J}T_L^k}{3a\left(\psi_{rd}^k - \frac{1}{4}\psi_{rd}^{k-1}\right)} \tag{12}$$

To implement a speed predictive control loop, it is necessary to know the load torque value T_L, which requires calculating the amount of torque generated from the applied current. If considered in the loop alone, the load torque is the same as an input disturbance. Using reverse Euler with sampling time T_{ds} = 10. T_s, load torque T_L is calculated in (13).

$$T_L = \frac{-J}{T_{ds}}(\omega^k - \omega^{k-1}) + \frac{3}{2}p\frac{L_m}{L_r}\psi_{rd}^{k-1}i_{sq}^{k-1} \tag{13}$$

3.4. Calculate the cost function

Using the FCS-MPC method, the objective function has the form:

$$g = |(i_{s\alpha}^*(k+1) - i_{s\alpha}(k+1))| + \left|\left(i_{s\beta}^*(k+1) - i_{s\beta}(k+1)\right)\right| \tag{14}$$

where: $i_{s\alpha}^*(k), i_{s\beta}^*(k)$: real and imaginary values of the reference line vector at; $i_{s\alpha}(k+1), i_{s\beta}(k+1)$: the real-to-imaginary value of the load-side forecast flow vector.

The 3-phase 2-level inverter has 8 switching states corresponding to 8 output voltage vectors. So, we will obtain 8 different objective function values, compare, which vector makes the objective function have the smallest value, then we give the control of opening and closing the IGBT valves. Specifically, the appropriate voltage vector selection algorithm is presented as shown in Figure 1.

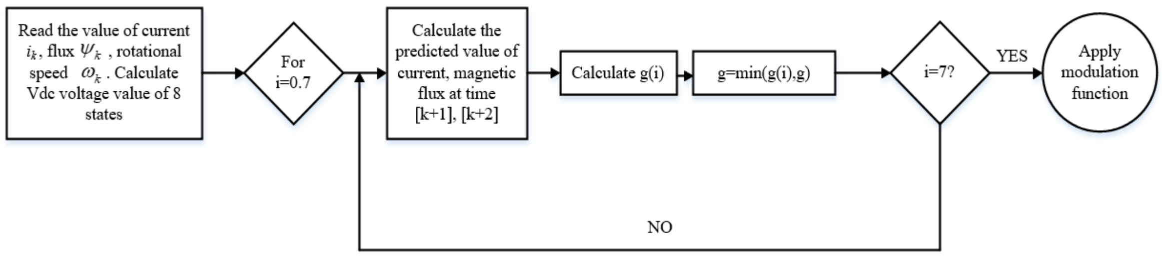

Figure 1. Algorithm to find the optimal voltage vector and modulation function

4. RESULTS AND DISCUSSION

To perform simulation on MATLAB-Simulink software with a motor parameter as Table 1. Simulation according to the following simulation scenario:

- Simulation time: 5s
- Sampling period: Ts = 200 s (f_C =5 kHz)
- Current I_{sd}*: I_{sd}* = 2.5 at initial time

– Speed ω*: ω*= 0 at the initial time, to the time from 0.5s-0.6s the speed gradually increases to 150 rad/s, stays the same until the time from 1.2s -1.6s the rate gradually decreases to -150 rad/s and stay the same until the end of the simulation.
– Resistance moment M_c: M_c = 0 at the beginning, then at t = [0.5, 1, 1.5, 1.75], Mc = [0.2TeN, -0.2TeN, 0.4TeN, -0.4TeN].

The simulation structure of predictive control is built into the simulation, as shown in Figure 2.

Table 1. The performance of an induction motor

Parameters	Symbol	Value	Unit
Rated power	P_N	2.2	kW
Frequency	f	50	Hz
Rated stator current	I_{sd}^*	2.5	A
Moment of inertia	J	0.0018	kg.m^2
Stator inductance	L_s	0.3072	H
Rotor inductance	L_r	0.4072	H
Stator resistance	R_s	1.89	Ω
Rotor resistance	R_r	1.99	Ω
Number of pole pairs	z_p	1	
Power factor	cosφ	0.87	

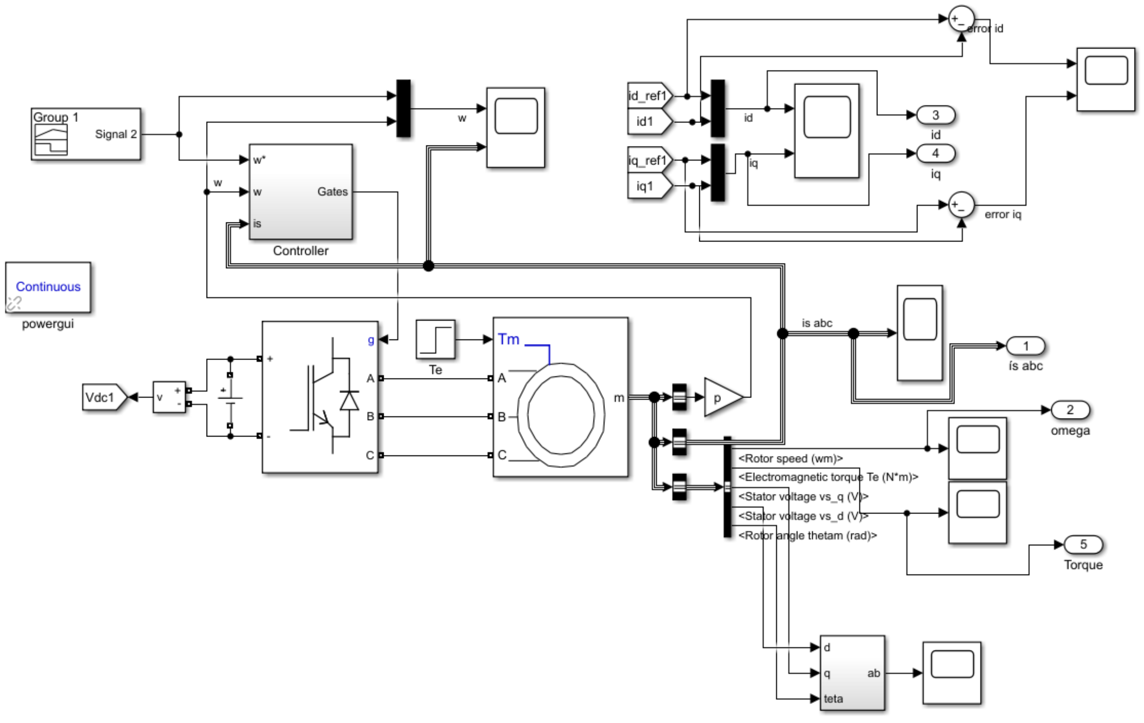

Figure 2. Simulation structure of predictive control

The speed response, 3-phase current of the converter, current error on the *dq* coordinate system, and torque response are shown in Figures 3, 4, 5, and 6 as follows. In Figure 3, the output speed from the tracking controller is breakneck after placing the signal, and the output speed from the controller is stable after about t=0.005 s. While in Figures 4 and 5, the current response time is about 0.02s. The error keeping the converter's set current and output current is relatively small, about 0.1 A for id and 0.4 A for i_q.

In Figure 6, it can be seen that the torque response meets the requirements. However, it is still over-regulated at the moment of transition and has a high overshoot. Conduct simulation with IM motor control system using the only predictive controller for current, speed loop circuit using traditional PI, using simulation parameters as above. The following results in Figures 7, 8 and 9.

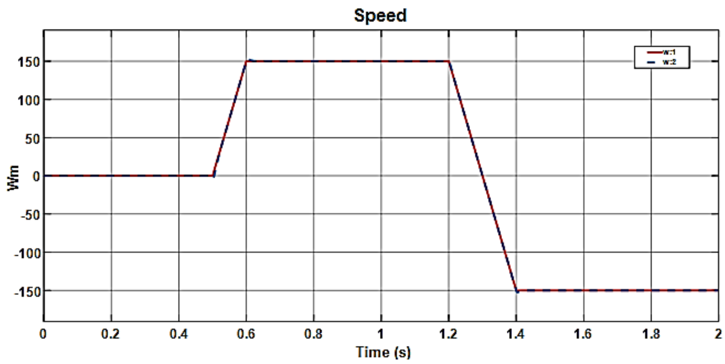

Figure 3. Reference speed and actual speed responses of the IM motor

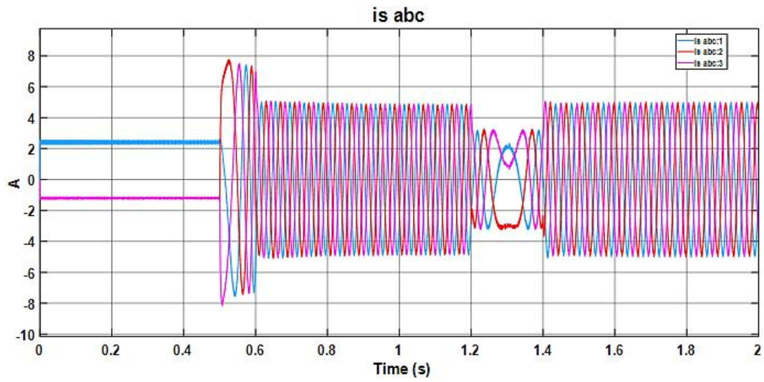

Figure 4. Stator current responses for inverter

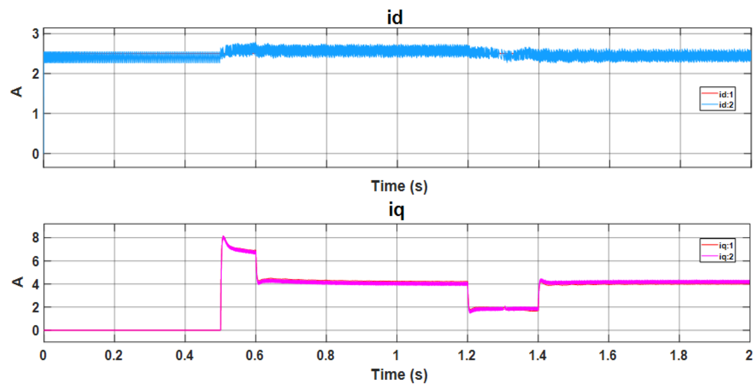

Figure 5. Error of stator current responses for inverter

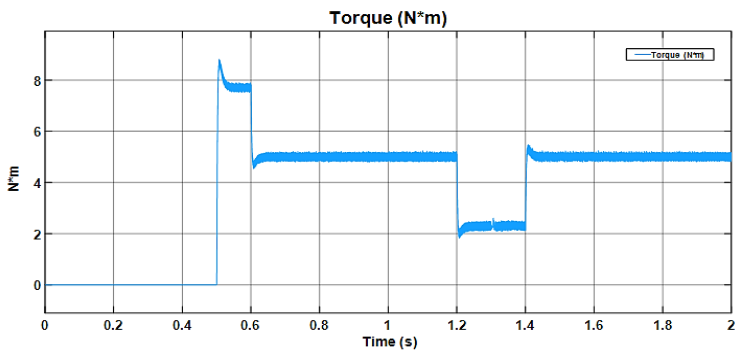

Figure 6. Torque responses

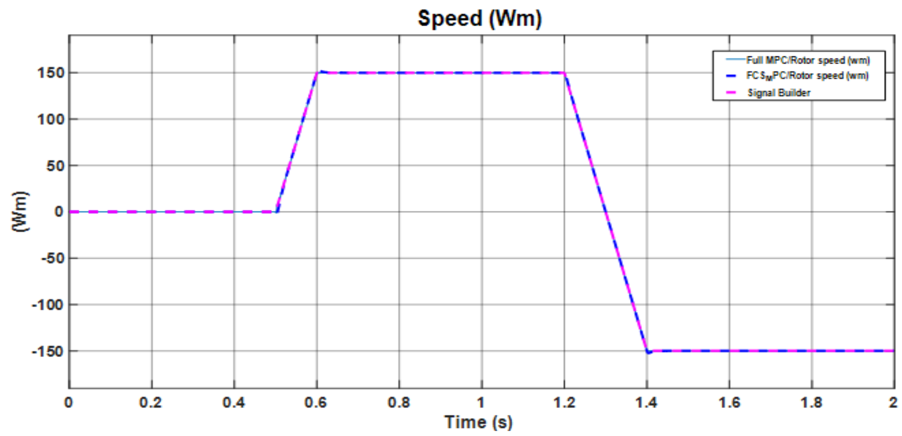

Figure 7. The speed responses for the PI and MPC controller

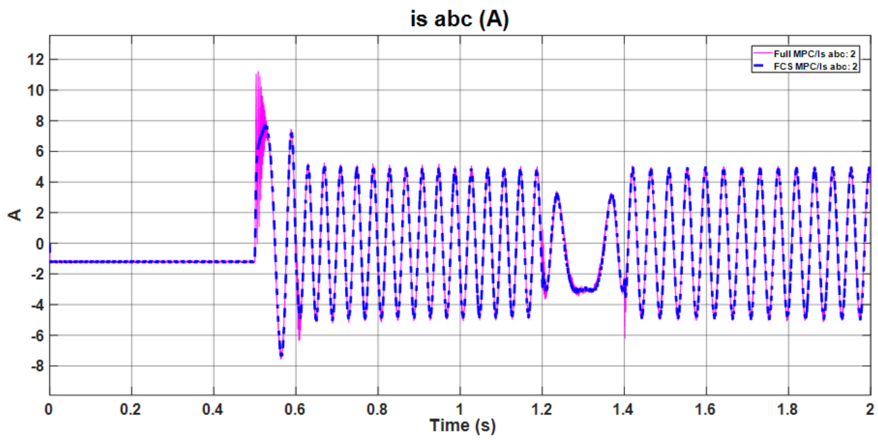

Figure 8. The *b* phase current response for the PI and MPC controller

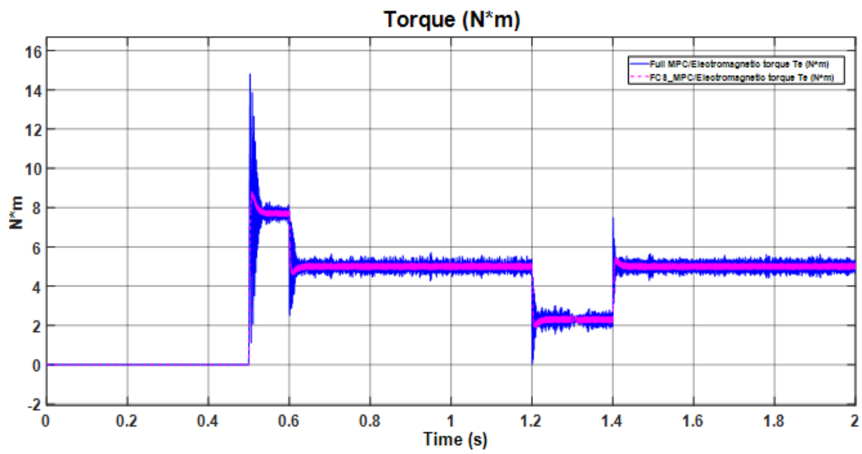

Figure 9. The torque response for the PI and MPC controller

The above results show that the tracking speed is faster when using both controllers as predictive controllers, the over-correction is also less than the torque control structure MPC, and the speed loop is a PI controller. In addition, the current and torque response of the traditional predictive controller is more stable, and the pulse rate is less than that of the predictive controller for both current and speed loops. In addition, both control methods give a sine phase current response.

5. CONCLUSION

The article has successfully designed torque and speed controllers for IM motors. The speed, current, and torque responses are correct as fast and accurate. However, these results only stop at the simulation, which has not yet evaluated the practicality of the problem. Therefore, to test the reliability of the control solution, we will conduct experiments on microcontrollers with fast computing power such as DSP, FPGA. to test the response of the traditional predictive controller for both current and speed loops.

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