Modulated Model Predictive Control of Permanent Magnet Synchronous Motors with Improved Steady-State Performance

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Abstract*—Finite Control Set Model Predictive Control (FCS-MPC) is an optimal control strategy that predicts the future trends of the control goals by assessing the discrete-time model of the system. FCS-MPC has many advantages, such as it has a fast dynamic response, and nonlinearities can be controlled by the customized cost function. Besides the featured benefits of the FCS-MPC strategy, the ripple in the output variable (in most cases, control variable) may be problematic due to the uncontrolled switching frequency. For that reason, the MPC-based closed-loop strategy offers a better regulation performance at high-sampling frequency. However, the selection of a low sampling rate causes an unpleasant distortion or poor power quality. A modulated model predictive control method is proposed in this work to suppress the unwanted distortion in the control variable. In the proposed method, a space vector modulator is integrated into the FCS-MPC-based control method to attain a fixed-switching frequency. By doing so, the distortions and unwanted harmonics are significantly decreased. In this paper, a modulated model predictive control (M^2PC) method is proposed for controlling the permanent magnet synchronous motor. The proposed method calculates the dwell-time of the modulator stage by assessing the multi-objective cost function. The noticeable lower distortions in the stator currents are obtained by the proposed routine. All theoretical concepts are verified by extensive simulations. Based on the simulation results, the proposed method provides a better control performance for permanent magnet synchronous motors (PMSM). Furthermore, the proposed modulated MPC strategy offers superior steady-state performance compared to the conventional MPC method in all regards.

Keywords—Model predictive control, Modulated model predictive control, PMSM, Inverter

I. Introduction

The use of electric vehicles in daily life is increasing due to their low greenhouse gas emissions and sustainability [1]. Due to their high efficiency, lower rotor losses compared to AC induction motors, and high torque to power ratio, the popularity of permanent magnet synchronous motors is increasing [2]. In surface-mounted PMSMs, the magnet is mounted on the surface of the rotor. For this reason, the torque is generated by the magnetizing flux. The most common control goals in the control of PMSM are the mechanical speed, torque, and machine phase currents. In particular, the field-oriented control (FOC) based closed-loop routines are quite popular in regulating the PMSM systems. In the FOC technique, the machine is controlled by regulating the d-q axis current components [3]. The other well-established control method in ac drive applications is direct torque control (DTC). In the DTC method, the mechanical torque and motor flux are estimated for feedback control, and stator currents are generally measured to improve the estimations and the low-speed operation performance. [4].

In DTC, the hysteresis controllers have been commonly used to apply the optimum voltage vector to the inverter. Because of the absence of the modulation stage, the system has a variable switching frequency when the hysteresis controller is applied to the system. Due to the modulator-free structure, the stator current (phase current of the PMSM) tends to have significant distortions, and the motor suffers from a high torque ripple in hysteresis-based DTC drives. In most cases, cascaded-controller are used to control the motor speed and stator current. To obtain acceptable closed-loop performance, the control parameters of the cascaded structure should be finely tuned.

Besides these conventional control methods, the FCS-MPC method is a promising control strategy for ac drive applications. The major drawback of the FCS-MPC method is its high computational effort. Since the optimization process must be performed online, a high computational burden is incurred by the FCS-MPC strategy. The required high computation power can be attained by the use of high-speed digital control platforms. With the development of powerful digital controllers (DSP, FPGA, etc.), the real-time implementation of the FCS-MPC is now feasible. In the FCS-MPC method, the control goals are predicted within a pre-defined prediction window. The prediction process is repeated for each allowable switching state. These switching states are the candidate solutions to the optimization problem. The error between the prediction and the instantaneous reference is introduced in the objective function. The switching state that offers the minimum cost value is picked [5]. The distinguishing features of the FCS-MPC are the ease of inclusion system constraints, fast transient performance, high closed-loop bandwidth, and multi-objective control [6]. Nevertheless, the FCS-MPC method does not regulate the operating frequency. Therefore, the converter operates under variable switching frequency conditions. The variable switching frequency negatively affects the steady-state performance regarding torque ripple and stator current total harmonic distortion (THD). The non-fixed switching frequency noticeably degrades the torque control performance

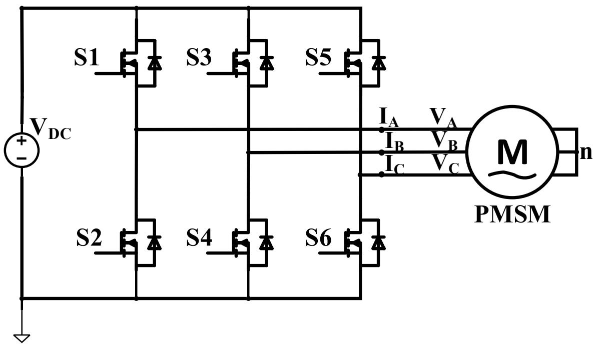


Fig. 1. Voltage-source inverter

and power quality (high stator current THD) [7], [8]. To eliminate this undesired effect, the sampling time can be reduced, but the lower sampling time increases the mathematical burden on the processor [9]. In this case, the control inputs are updated more often; thus, the computational burden increases exponentially. The other favorable solution without lowering the sampling period is using a modulator to control the switching frequency. By combining the modulator and FCS-MPC method, the desired steady-state performance can be attained at a lower sampling period. Hence, no strict sampling period constraint is required. In addition, the modulated model predictive method (M^2PC) provides a better energy conversion operation compared to the conventional MPC technique [10].

The different types of converters are controlled by the FCS-MPC method and M^2PC methods [11], [12]. The voltage source inverter (VSI) with RL load controlled by the MPC method is presented in [13] with experimental results. The modulated model predictive control strategy has been explained in [10], and the comprehensive comparison results are reported. Based on the reported results in [10], the M^2PC provides better closed-loop performance compared to the conventional MPC. The varieties of the predictive control methods have been successfully applied to the ac drive applications [14]–[16]. MPC method is used to control flux and torque of AC induction motors [17]. The control of PMSM is performed using three different MPC methods, and the experimental results are compared to each other [18]. The control of a brushless doubly-fed induction machine and polyphase induction motor has been presented using the M^2PC method [19], [20]. Furthermore, the predictive speed control strategies have been investigated in [21].

The limitations of the conventional MPC on the steady-state performance (high torque ripple and poor power quality) are the primary motivation of this paper. In this paper, a modulated model predictive control strategy is proposed for controlling the PMSM. The proposed method provides a noticeable improvement in the steady-state performance of the system. The proposed M^2PC strategy controls the mechanical speed, torque, and stator currents, and power quality is improved compared to the one offered by the conventional MPC method. Furthermore, the proposed control method remarkably reduces the torque ripple, and it suppresses the stator current THD at a low sampling rate. The mathematical concepts are proved by the simulation work, and the simulation results demonstrate the potency of the proposed modulated predictive control routine.

II. System Model

A. Voltage-source inverter model

The circuit diagram of a voltage-source inverter fed PMSM is shown in Fig. 1. The power converter has six switching devices, and three independent output terminals are available. The potentials of windings for a balanced star-connected three-phase system are expressed as

$$\begin{bmatrix} V_{An} \\ V_{Bn} \\ V_{Cn} \end{bmatrix} = \frac{V_{DC}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_1 \\ S_3 \\ S_5 \end{bmatrix} \tag{1}$$

where switch position $S_j \in \{0,1\}$.

B. PMSM Model

In this study, the PMSM model is expressed in the synchronously rotating reference frame where d-q axes rotate at the synchronous speed [2]. The dynamic model of the PMSM is given by

$$V_d = RI_d + L_d \frac{dI_d}{dt} - \omega_e L_q I_q \tag{2}$$

$$V_q = RI_q + L_q \frac{dI_q}{dt} + \omega_e (L_d I_d + \varphi_m) \tag{3}$$

$$T_e = \frac{3p}{2} (\varphi_m I_q - I_d I_q (L_q - L_d)) \tag{4}$$

$$\theta_e = \int \omega_e \, dt \tag{5}$$

where $V_{d\text{-}q}$ is d-q axis voltages, $I_{d\text{-}q}$ is d-q axis currents, $L_{d\text{-}q}$ is d-q axis inductances, R stator resistance, ω_e electrical synchronous angular frequency, φ_m permanent magnet flux linkage, T_e electromagnetic torque, θ_e rotor electrical angle, and p number of pole pairs. Since L_d=L_q in SPMSM, the electromagnetic torque can be controlled via the q-axis current component.

$$T_e = \frac{3p}{2} (\varphi_m I_q) \tag{6}$$

In M^2PC and MPC methods, system control can be done by expressing the mathematical model of the system in discrete time. In equations (2)-(3), the derivative terms can be converted to the discrete-time form by applying the Forward Euler method:

$$\frac{df}{dt} = \frac{f(k+1) - f(k)}{T_s} \tag{7}$$

where T_s is sampling time. The discrete-time models of the stator current (d-q frame) are defined in (8)-(9).

$$I_d(k+1) = \frac{V_d(k) T_s}{L_d} + \frac{\omega_e L_q I_q(k) T_s}{L_d} + I_d(k) \left(1 - \left(\frac{RT_s}{L_d}\right)\right) \tag{8}$$

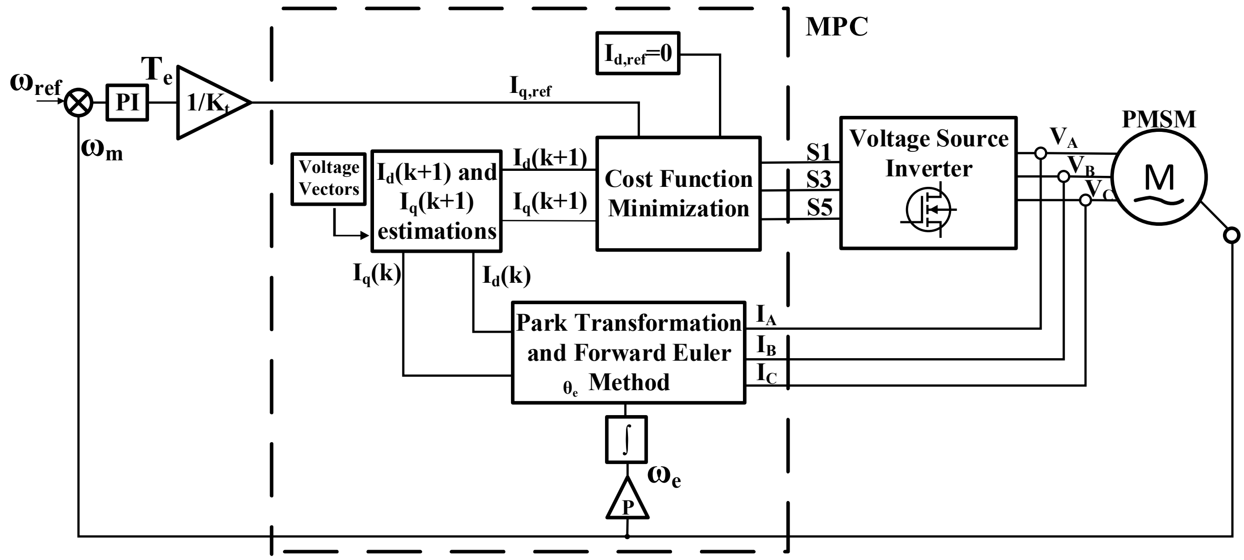


Fig. 2. Block diagram of conventional FCS-MPC

$$I_q(k+1) = \frac{V_q(k)T_s}{L_q} - \frac{\omega_e L_d I_d(k) T_s}{L_q} - \frac{w_e \varphi_m T_s}{L_q} + I_q(k)\left(1 - \left(\frac{RT_s}{L_q}\right)\right) \quad (9)$$

The discrete-time model of the rotor electrical angle is given by

$$\theta_e(k) = \theta_e(k-1) + \omega_e(k-1)T_s \quad (10)$$

III. Proposed Method

The block diagram of the traditional MPC method is shown in Fig. 2. In Fig. 2, ω_m, the mechanical speed of PMSM is measured via a speed sensor (such as an incremental encoder). The measured stator currents are transformed to the d-q frame. The control goals are predicted for each feasible control input. The cost function is assessed to determine the best switching state. The switching combination that minimizes the discrepancy from the instantaneous reference is selected. The determined switching combination is directly applied to the VSI. In this process, the feasible inputs are generated by the controller. This approach is a traditional MPC strategy, and as can be seen from Fig. 2, no modulator is used, and the switching frequency is variable. The objective function of the conventional MPC is expressed as

$$g = (I_{d,ref} - I_d)^2 + (I_{q,ref} - I_q)^2 \quad (11)$$

To improve the steady-state performance, the M^{2}PC method, which combines SVPWM and MPC, is proposed in this study. The block diagram for the M^{2}PC method is shown in Fig. 3. As typical in the SVPWM technique, two active vectors and a zero vector are used to obtain optimum voltage actuation. By utilizing the corresponding adjacent vectors and null vector, the average current error becomes zero. The dwell time of each active vector and null vector must be appropriately determined and applied according to the pre-defined switching pattern [21]. The use of a modulator enables the generation of any desired voltage vector. Thus, any desired voltage vector can be employed by the designed control system. Only discrete voltage vectors generated by the allowable switching combinations can be utilized in the conventional MPC method. On the other hand, the voltage vector that offers the zero tracking error can be produced in the proposed method. The duty cycle values of the active vectors and zero vector can be determined as

$$d_0 = \frac{T_s g_1 g_2}{g_0 g_1 + g_1 g_2 + g_0 g_2} \quad (12)$$

$$d_1 = \frac{T_s g_0 g_2}{g_0 g_1 + g_1 g_2 + g_0 g_2} \quad (13)$$

$$d_2 = \frac{T_s g_0 g_1}{g_0 g_1 + g_1 g_2 + g_0 g_2} \quad (14)$$

where $d_{0\text{-}1\text{-}2}$ are duty cycles of zero and two active voltage vectors, respectively, the $g_{0\text{-}1\text{-}2}$ are cost function values of zero and two active voltage vectors, respectively. Finally, a new cost function is used according to the cost function results that are kept with equation (11) and duty cycle values that are found with equations (12)-(14).

$$G = d_0 g_0 + d_1 g_1 + d_2 g_2 \quad (15)$$

The active voltage vectors that offer minimum cost value (the minimization of (15) and a null vector are selected for the next time interval. The selected active voltage vectors are the solutions to the optimization problem. The modulator uses the determined vectors and generates the gate pulses accordingly. Thus, the fixed-switching frequency is achieved. The prediction procedure of the proposed method is similar to the conventional method. However, the control input generation procedure and the optimization process are noticeably different from the conventional MPC strategy.

IV. SIMULATION RESULTS

To demonstrate the benefits of the proposed control strategy, the system is simulated using Matlab/Simulink©. The simulation parameters are listed in Table I. For the simulation results provided in Fig. 4-6, the mechanical speed reference is ω_{ref}=120 rad/s, and the applied torque is T_L=3Nm for conventional MPC and M^{2}PC. The phase-A current waveforms with the conventional MPC and proposed M^{2}PC

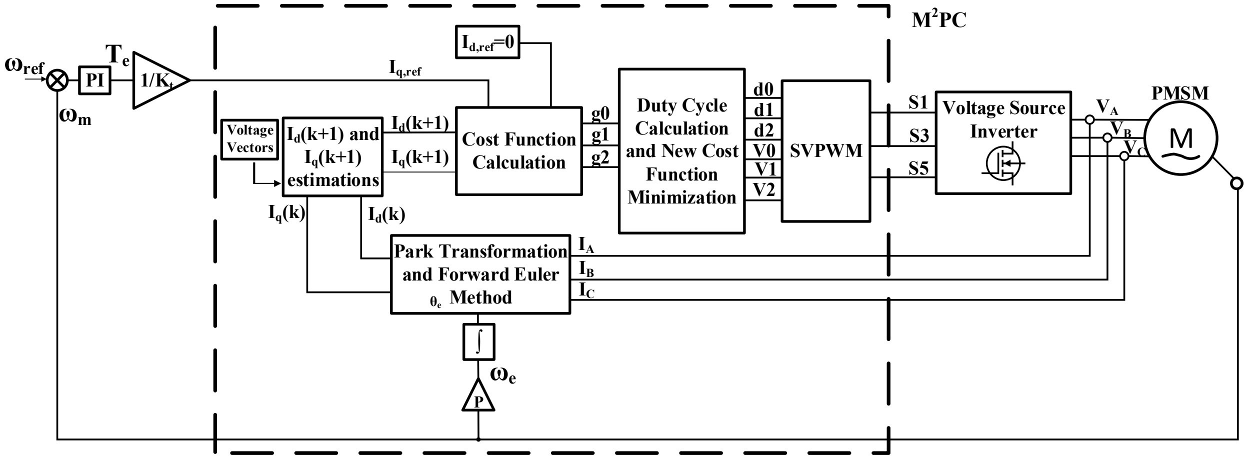
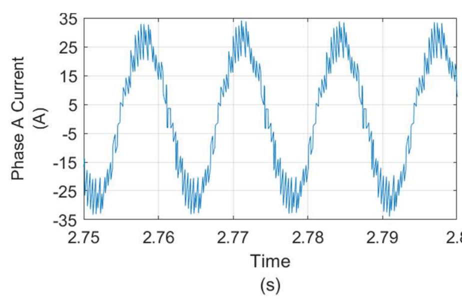
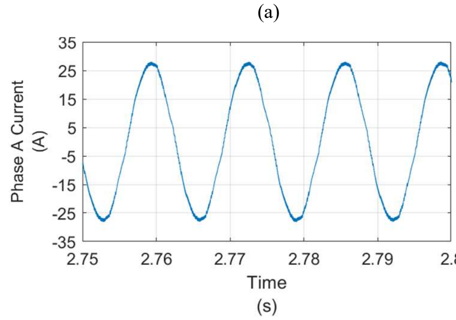


Fig. 3. The control schematic of the proposed M^{2}PC to control PMSM

methods are presented in Fig 4. According to Fig. 4, the stator current is less distorted when the proposed method is applied to the system, see Fig. 4(b). The electromechanical torque of the motor is presented in Fig. 5. The torque ripple is reduced when the proposed method is chosen as the control strategy. The phase-to-phase voltage waveform is shown in Fig. 6. The comparison results between the proposed method and the conventional MPC method are summarized in Table II in terms of stator current THD. The average switching frequency is also calculated for the conventional MPC method and reported in Table II. It has been validated that the sampling time must be reduced to obtain the optimum average switching frequency range. The stator current THD is 18.72% for a sampling period of 50 µs when the conventional MPC is applied to the system. On the other hand, the stator current THD is 2.75% at 50 µs, which corresponds to a switching frequency of 20 kHz, for the proposed control method. The simulation results prove that the proposed M^{2}PC method offers a significant reduction in stator current THD.



(a)



(b)

Fig. 4. Stator phase A current waveforms; (a) Conventional MPC method, (b) Proposed M^{2}PC method

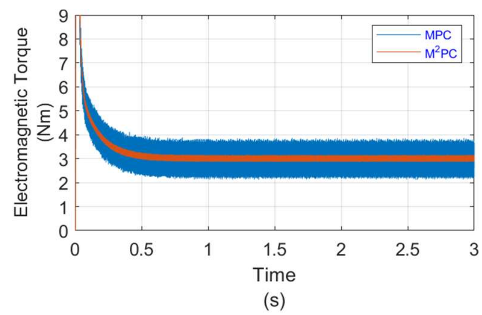
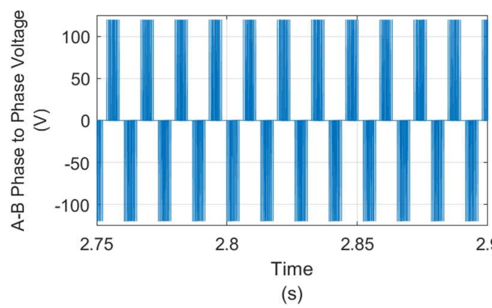
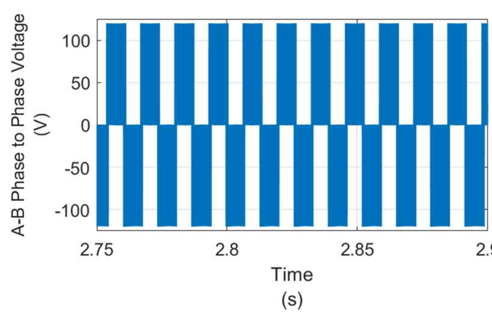


Fig. 5. Electromagnetic torque waveforms



(a)



(b)

Fig. 6. Inverter A-B phase to phase voltage waveforms: (a) Conventional MPC method, (b) The proposed M^{2}PC method

TABLE I
SIMULATION PARAMETERS

Parameter	Description	Values
T_S	Sampling Period	50 µs
p	Number of pole pairs	4
Φ	Number of phases	3
P	Continuous power	4.1 kW
ω_m	Nominal Speed	2500 rpm
L_d	Nominal d-axis Inductance	0.282 mH
L_q	Nominal q-axis Inductance	0.282 mH
φ_m	Nominal PM Flux Linkage	0.0182 Wb
R	Nominal Phase Resistance	0.0463 Ω
J	Inertia	0.0072 kgm^2

TABLE II
THD VALUES OF STATOR CURRENT FOR MPC AND M^{2}PC

Control Type (T_L=3Nm)	T_S	f_{sw}	THD (%)
M^{2}PC	200 µs	5 kHz	7.76%
M^{2}PC	100 µs	10 kHz	5.01%
MPC	100 µs	1.3 kHz (avg.)	39.26%
M^{2}PC	50 µs	20 kHz	2.75%
MPC	50 µs	2.7 kHz (avg.)	18.72%
MPC	20 µs	6.5 kHz (avg.)	7.97%
MPC	10 µs	13.5 kHz (avg.)	3.85%

To examine the transient performance of the proposed method, several system steps including the speed step and torque step are applied to the closed-loop system. Fig. 7 presents the dynamic response of the proposed control method. In this test scenario, the mechanical speed profile is ω_{ref}=120 rad/s between t=0-2.5 s and ω_{ref}=140 rad/s between t=2.5-3.5 s. The torque profile T_L=2Nm between t=0-1.5 s and T_L=6Nm between t=1.5-3.5 s. The sampling time T_s=50 µs. During the start-up transient, the motor speed achieves the reference speed (120 rad/sec), and no unpleasant overshoot is spotted. To examine the speed control performance of the proposed method under the torque variations, the torque step is applied at the time instant t=1.5 s. During the torque transition, the magnitude of the stator current is adjusted by the proposed method to satisfy the new loading condition. Furthermore, the mechanical speed control is quite stable when the torque step is applied. The motor speed is slightly decreased, and the outer-loop updates the torque reference. Once the new torque reference is updated, the proposed method compensates for the speed error. Then, the motor speed keeps tracking its reference.

The simulation studies have been further carried out to compare the computational burden with the MPC and M^{2}PC technics. The simulation has been operated for 3 seconds in real time by setting the reference speed to ω_{ref}= 1145 rpm and T_L = 3 Nm for the two methods. The sampling time is set to 50 µs for both methods. The simulation performed with the conventional MPC method lasted ~79 seconds while the simulation performed with the M^{2}PC method lasted ~127 seconds. The measured stator current THD value is %18.37 for the conventional MPC method. On the other hand, the stator current THD value for the M^{2}PC method is 2.38%.

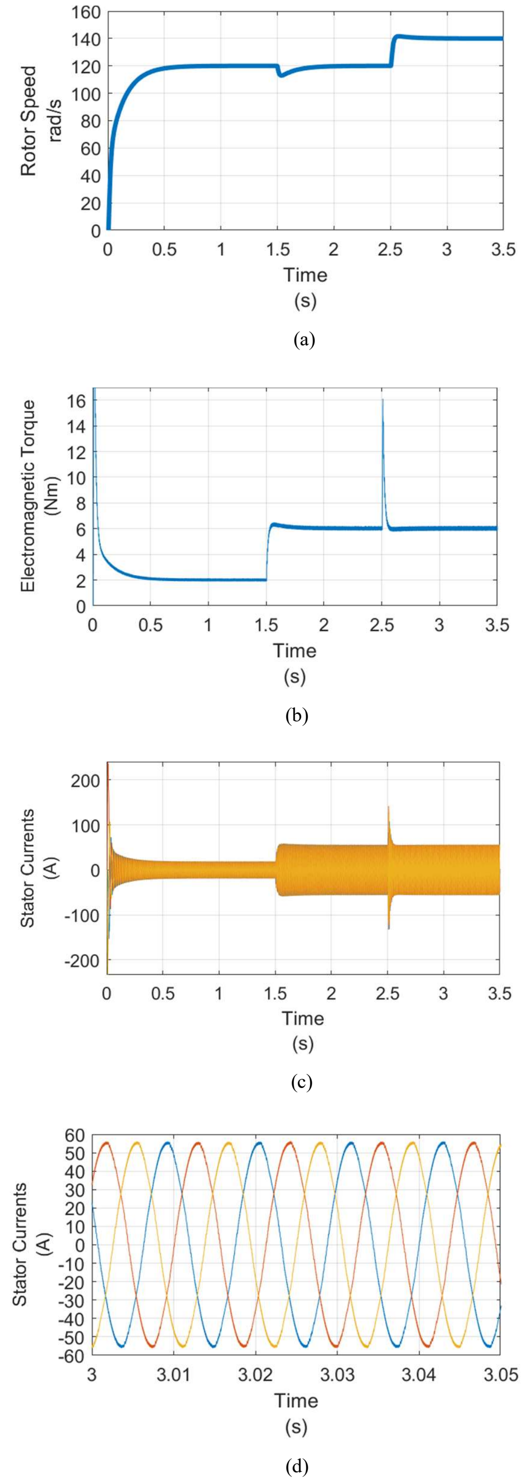


Fig. 7. Waveforms of load side; (a) rotor speed, (b) electromagnetic torque, (c) stator currents, (d) zoom-in stator currents

Even if the modulation block increases the execution time due to the extra calculations that come with the M^{2}PC method, the improvement in the stator current quality is quite important. Thus, the additional control calculations can be tolerated to improve the power quality. The simulation has been repeated for the conventional MPC method when the sampling time is

20 µs to reduce the THD. The execution time is measured as ~201 s for the same operating profile, and the stator current THD is 7.091%. Regarding the execution time, the proposed method still offers a reduced execution time even if the modulation block is included in the feedback design. As a result, the proposed method provides better steady-state performance compared to the conventional MPC method. The sampling frequency criterion is relaxed by adding the modulator to the system.

V. Conclusion

This paper proposes a modulated model predictive approach to control VSI driving a permanent magnet synchronous machine. SVPWM technique and MPC technique are combined to improve the steady-state performance of the system. The proposed method regulates the operating frequency. Hence, the proposed method offers a better THD performance, and the torque ripple of the motor is significantly reduced. The other important feature of the proposed method is the ease of selecting passive components for filter design. Since operating frequency is fixed, the selection of the filter parameters becomes much easier. The detailed comparison results between the proposed method and the conventional MPC strategy are provided. As a result, the power quality of the energy conversion process is improved by applying the proposed M^2PC method. Its superiority and robustness to step variations have been validated through realistic simulations.

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