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Modulated Predictive Control to Improve the Steady-State Performance of NSI-Based Electrification Systems

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Abstract: This paper presents a modulated model predictive control (M^2PC) strategy for a nine-switch inverter (NSI) based electrification system to improve the steady-state performance. The model predictive control method has gained significant interest due to its straightforward structure. However, the traditional finite control set model predictive control (FCS-MPC) imposes a high computational burden that is problematic in practical applications. This prevents reaching the high sampling frequencies due to an excessive increase in algorithm run-time. Selecting a low sampling frequency causes an unpleasant distortion in the control variable or poor power quality. An M^2PC method for the NSI is proposed in this work to remove this trade-off. One zero vector and two active vectors are selected by evaluating a cost function for each allowed switching state of the NSI. The duty cycles of these vectors are calculated by assessing the cost function employing current error terms. An optimized sequence of these vectors is applied to the system that operates with the fixed-modulation frequency. Thus, an improvement in power quality (reduced harmonics with a better spectral content) with a lower sampling frequency is achieved. The computational burden rate (CBR) on the processor is reduced. These enhancements were proved by simulation and experimental studies. The comparison work was conducted to highlight the advantages of the proposed method over the other techniques reported in the literature. The proposed M^2PC method was verified on a lab-scale NSI prototype driving two induction machines. The machine torques and speeds are well regulated, and the quality of the stator current is improved.

Keywords: nine-switch inverter; multi-drive systems; modulated model predictive control; indirect field-oriented control; field-programmable gate array



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1. Introduction

In power electronics applications, reduced count topologies have been receiving attention due to their compact structures [1]. Among these topologies, the nine-switch inverter (NSI) is a promising solution for various applications. The NSI has the capability of driving two three-phase ac loads using fewer switching devices. The key benefit of using reduced count topology such as NSI is a remarkable reduction in the power stage's size, volume, and weight [2,3].

The NSI topology and its application were first introduced in [3]. Following the initial emergence of the NSI, some studies focused on developing appropriate control and modulation techniques for NSI [4–11]. Several other research groups investigated the improved modulation strategies that reduce the thermal stress of the NSI by taking into account conduction and switching losses [12–15]. Another research consideration was to derive the NSI topologies combined with impedance source networks to increase the inverter's voltage utilization [16,17].

The NSI has been used as a core topology in various power energy conversion applications such as an integrated motor drive and battery charger [18], a DFIG based wind power system [19], and grid-connected hybrid ac/dc energy sources [20]. Apart from

these applications, various control strategies for the NSI have been reported in the literature to control the two independent three-phase loads/machines or a single six-phase load [2,3,7,17,21–26]. These strategies are mostly combined versions of vector control and suitable pulse-width modulation (PWM) techniques. Among the control strategies for NSI-driven two separate three-phase loads, the model predictive control (MPC) method is a desirable feedback strategy to regulate multi-phase/load dynamics [27,28]. The MPC uses the explicit system model (including the converter and load models) to predict the control variables [29]. In ac drive applications, the control variables are usually electromechanical torque, rotor speed, and stator phase currents [30,31]. The MPC predicts the control variables for each feasible control input and assesses the multi-objective cost function. The control input that offers minimum cost value is picked and applied to the system. In general, the MPC method is a very promising control strategy in ac drive applications due to the ease of including control objectives, control constraints, and system limitations [32–34]. However, there are several drawbacks with the traditional finite control set (FCS) MPC method—poor steady-state performance and high computational burden. In particular, the FCS-MPC drastically suffers from poor steady-state performance due to the lack of a modulator stage, especially for systems with small electrical time constants requiring that high sampling frequency is mandatory to achieve acceptable power quality [35,36]. Since the MPC requires a high number of calculations using a system model, it becomes challenging for existing DSP technologies to complete these calculations at high sampling frequencies [37]. Moreover, the MPC produces the gate signals directly, and the system has a non-fixed switching frequency. The variable operating frequency causes variable switching harmonics with a wide-range frequency spectrum. In this case, shaping the frequency contents is not possible, resulting in uncontrolled harmonics and complicated filter designs. Therefore, the control variables (output variables) suffer from unwanted distortions and harmonics. Thus the steady-state performance of the traditional MPC is degraded due to the uncontrolled switching frequency [38,39]. Several strategies have been proposed to improve the steady-state performance, including a switching frequency control term in the cost function [40]. However, penalizing the switching effort in the objective function does not noticeably reduce the total of harmonic distortions (THD). The most common technique to attain a good steady-state performance is to reduce the sampling period. To reduce the sampling period effectively reduces the THD in control variables; however, the practical systems usually have a limitation on computational power. The sampling period must be long enough to cover the control algorithm execution time. Because of this, the energy conversion operation is not always possible with a lower sampling period [41].

The exciting strategy to improve the steady-state performance of the traditional MPC is to include the modulator in the closed-loop design. This strategy is called modulated model predictive control (M^2PC), which considers the dedicated switching pattern to generate the gate pulses [42–47]. The M^2PC operates under a fixed switching frequency, and the frequency harmonics are distributed around the switching frequency and its harmonics. Hence the unwanted variable frequency contents imposed by the variable system frequency are eliminated. The controlled switching frequency remarkably reduces the THD in control variables and enhances the steady-state characteristics. The essential advantage of the M^2PC is that superior steady-state performance can be achieved even at a lower sampling frequency [39]. This is quite a vital point since a decent steady-state performance can be obtained without increasing the sampling frequency. In some practical applications, a sampling frequency constraint (limitation on the maximum value of the sampling frequency) has been considered. Hence, achieving steady-state goals at a lower sampling frequency is a good capability of the feedback strategies [48]. Various M^2PC techniques have been reported for the ac drive applications in the literature. In [43], an M^2PC technique that utilizes virtual voltage vectors to reduce the low-order harmonics in the phase currents has been proposed for a five-phase voltage source inverter (VSI). The interesting implementation of the M^2PC method with optimal switching time and reduced complexity [46] has been proposed for a neutral point clamped (NPC) three-phase

inverter. Another research paper [47] investigated a modulated-type predictive control that uses optimal two voltage vectors and associated duty cycles to reduce the phase current THD. In [49], a modulated predictive current control with properly adopted space-vector PWM technique was presented to control an asymmetrical dual three-phase induction motor driven by a six-phase VSI. Although the reported strategies are useful for traditional topologies such as VSI and NPC, no M²PC type strategy has been presented to control the NSI topology.

Motivated from the above discussion, a novel M²PC strategy is proposed in this paper to generate the required gate signals for the NSI in order to regulate the dynamics of two induction motors. Essentially, the proposed method can be used for any energy conversion system employing the NSI as far as the system model is available. The proposed method reduces the undesired harmonics and offers better steady-state performance even at a lower sampling frequency. The stator current THD is noticeably reduced to improve the power quality. The independent induction motors are controlled under different loading conditions, and the demanded torque and speed profiles are maintained. The primary benefit of the proposed method is that steady-state torque ripple and stator current THD are reduced. The torque ripples degrade the mechanical performance of the induction motors and negatively affect the rotor speed. Due to these unpleasant aspects, improving the reference trajectory tracking performance is crucial for a better energy conversion process. The main contributions of the paper are summarized as follows:

1. A novel M²PC method is proposed to control two induction motors fed by a nine-switch inverter. The proposed strategy effectively controls the independent motors and provides a robust operation without undesired interactions.
2. The steady-state performance is significantly improved by reducing the stator current THD and the torque ripple. The problem with the steady-state performance of the MPC strategy is solved by the proposed method without increasing the sampling frequency, and the stator current THD is reduced approximately by 50% at low current levels. The THD reduction in stator current also decreases the associated losses. The torque ripple is reduced by 25% under steady-state conditions.
3. The proposed method is then experimentally validated using a real hardware NSI prototype. The mathematical framework of the proposed method is proved under different real test scenarios. Different experimental cases are considered to demonstrate the effectiveness of the proposed method. Furthermore, comprehensive comparison works are performed between the proposed method and the traditional MPC strategy. Quantitative comparison results are provided to prove the advantages of the proposed method over the conventional MPC.

2. Nine-Switch Inverter Basics

The Nine Switch Inverter (NSI) was built by merging two VSIs to decrease the number of required switching devices by 25% compared to the conventional system (two parallel inverters). Using fewer switching devices in the power stage design increases the effective power density. The NSI can drive two independent three-phase loads or a single six-phase load (with galvanically isolated two three-phase windings), see Figure 1. The NSI consists of three switches in each leg, and the middle switches S_4 , S_5 , and S_6 are shared by the upper and lower load stages [3]. During operation, only two switches in each leg of the NSI are allowed to conduct simultaneously, and this constraint can be defined as:

$$S_i + S_{i+3} + S_{i+6} = 2 \text{ with } i \in \{1, 2, 3\} \quad (1)$$

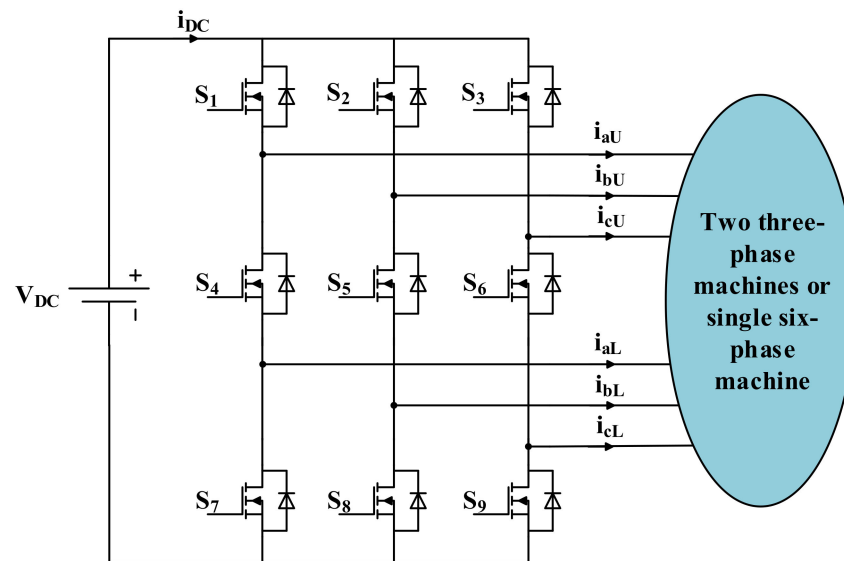


Figure 1. Nine-switch inverter-based electrification system.

Thus, short-circuiting of the dc bus is prevented while the current continuity of the inductive loads is guaranteed. So, there are three possible switching states for each leg and a total of 27 allowed switching states for NSI, as listed in Table 1. According to Table 1, the switching states of NSI can be divided into three groups. While Group I is the group with zero vectors for both load stages, Group II has zero vectors for one of the loads and active vectors for the other, while Group III consists of active vectors for both loads [9].

Table 1. NSI Switching Vectors.

Group	States	NSI Switch Position *	V_{aUn}	V_{bUn}	V_{cUn}	$V_{aLn'}$	$V_{bLn'}$	$V_{cLn'}$
I	sw ₁	[1 1 1; 1 1 1; 0 0 0]	0	0	0	0	0	0
	sw ₂	[0 0 0; 1 1 1; 1 1 1]	0	0	0	0	0	0
	sw ₃	[1 1 1; 0 0 0; 1 1 1]	0	0	0	0	0	0
II	sw ₄	[1 0 0; 0 1 1; 1 1 1]	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	0	0	0
	sw ₅	[1 1 0; 0 0 1; 1 1 1]	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	0	0	0
	sw ₆	[0 1 0; 1 0 1; 1 1 1]	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	0	0	0
	sw ₇	[0 1 1; 1 0 0; 1 1 1]	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	0	0	0
	sw ₈	[0 0 1; 1 1 0; 1 1 1]	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	0	0	0
	sw ₉	[1 0 1; 0 1 0; 1 1 1]	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	0	0	0
	sw ₁₀	[1 1 1; 1 0 0; 0 1 1]	0	0	0	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₁₁	[1 1 1; 1 1 0; 0 0 1]	0	0	0	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$
	sw ₁₂	[1 1 1; 0 1 0; 1 0 1]	0	0	0	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₁₃	[1 1 1; 0 1 1; 1 0 0]	0	0	0	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$
	sw ₁₄	[1 1 1; 0 0 1; 1 1 0]	0	0	0	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$
	sw ₁₅	[1 1 1; 1 0 1; 0 1 0]	0	0	0	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$
III	sw ₁₆	[1 0 0; 1 1 1; 0 1 1]	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₁₇	[1 1 0; 1 1 1; 0 0 1]	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$
	sw ₁₈	[0 1 0; 1 1 1; 1 0 1]	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₁₉	[0 1 1; 1 1 1; 1 0 0]	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$
	sw ₂₀	[0 0 1; 1 1 1; 1 1 0]	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$
	sw ₂₁	[1 0 1; 1 1 1; 0 1 0]	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$
	sw ₂₂	[1 1 0; 1 0 1; 0 1 1]	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₂₃	[1 1 0; 0 1 1; 1 0 1]	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₂₄	[0 1 1; 1 1 0; 1 0 1]	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$
	sw ₂₅	[0 1 1; 1 0 1; 1 1 0]	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(1/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$
	sw ₂₆	[1 0 1; 0 1 1; 1 1 0]	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$	$(2/3)V_{DC}$
	sw ₂₇	[1 0 1; 1 1 0; 0 1 1]	$(1/3)V_{DC}$	$(-2/3)V_{DC}$	$(1/3)V_{DC}$	$(2/3)V_{DC}$	$(-1/3)V_{DC}$	$(-1/3)V_{DC}$

* NSI switch position = [S₁ S₂ S₃; S₄ S₅ S₆; S₇ S₈ S₉].

The transition between the input and output quantities of the NSI is performed via the matrices given in (2) and (3):

$$\mathbf{T}_U = \begin{bmatrix} S_1 & S_2 & S_3 \end{bmatrix} \quad (2)$$

$$\mathbf{T}_L = \begin{bmatrix} (1 - S_7) & (1 - S_8) & (1 - S_9) \end{bmatrix} \quad (3)$$

The NSI leg voltages for the upper and lower load stages can be defined as:

$$\mathbf{V}_U = V_{DC} \mathbf{T}_U^T \quad (4)$$

$$\mathbf{V}_L = V_{DC} \mathbf{T}_L^T \quad (5)$$

where $\mathbf{V}_U = \begin{bmatrix} V_{aU} & V_{bU} & V_{cU} \end{bmatrix}^T$, $\mathbf{V}_L = \begin{bmatrix} V_{aL} & V_{bL} & V_{cL} \end{bmatrix}^T$. The three-phase load voltages for a balanced system can be calculated by using (6) and (7):

$$\begin{bmatrix} V_{aUn} \\ V_{bUn} \\ V_{cUn} \end{bmatrix} = \frac{V_{DC}}{3} \mathbf{T}_U^T = \frac{V_{DC}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_1 \\ S_2 \\ S_3 \end{bmatrix} \quad (6)$$

$$\begin{bmatrix} V_{aLn'} \\ V_{bLn'} \\ V_{cLn'} \end{bmatrix} = \frac{V_{DC}}{3} \mathbf{T}_L^T = \frac{V_{DC}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} (1 - S_7) \\ (1 - S_8) \\ (1 - S_9) \end{bmatrix} \quad (7)$$

3. Induction Machine Model

3.1. Dynamic Model of the Induction Machine

The dynamic model of an induction machine in the d-q reference frame, that rotates with a synchronous angular speed of ω_g and an instantaneous angle of θ with respect to the fixed-frame, can be defined with the set of Equations (8)–(13) [50]. Here, the electrical state variables are direct (d) and quadrature (q) axis stator current components and d - q axis rotor flux components. The dynamic model of the current components can be defined as:

$$\frac{di_{sd}}{dt} = -\frac{L_r R_s + \frac{L_m^2}{\tau_r}}{\sigma L_s L_r} i_{sd} + \omega_g i_{sq} + \frac{L_m}{\tau_r \sigma L_s L_r} \Psi_{rd} + \frac{v_{sd}}{\sigma L_s} + \frac{L_m}{\sigma L_s L_r} \omega_r \Psi_{rq} \quad (8)$$

$$\frac{di_{sq}}{dt} = -\frac{L_r R_s + \frac{L_m^2}{\tau_r}}{\sigma L_s L_r} i_{sq} - \omega_g i_{sd} + \frac{L_m}{\tau_r \sigma L_s L_r} \Psi_{rq} + \frac{v_{sq}}{\sigma L_s} - \frac{L_m}{\sigma L_s L_r} \omega_r \Psi_{rd} \quad (9)$$

where stator, rotor, and mutual inductances are denoted with L_s , L_r , and L_m , respectively, the stator and rotor winding resistances are designated with R_s and R_r , the term $\tau_r = L_r / R_r$ is the rotor time constant, and the leakage constant is calculated by $\sigma = (L_r L_s - L_m^2) / L_r L_s$. The electrical angular speed, ω_r , is calculated by $\omega_r = p \omega_m$ in rad/s where p is the number of pole pairs, ω_m is the mechanical angular speed of the rotor.

The dynamic model of the rotor flux is defined as:

$$\frac{d\Psi_{rd}}{dt} = \frac{L_m}{\tau_r} i_{sd} - \frac{\Psi_{rd}}{\tau_r} + (\omega_g - \omega_r) \Psi_{rq} \quad (10)$$

$$\frac{d\Psi_{rq}}{dt} = \frac{L_m}{\tau_r} i_{sq} - \frac{\Psi_{rq}}{\tau_r} - (\omega_g - \omega_r) \Psi_{rd} \quad (11)$$

The rotor electrical angular speed modelling the mechanical dynamic is given by:

$$\frac{d\omega_r}{dt} = \frac{3}{2} \frac{p L_m}{J L_r} (\Psi_{rd} i_{sq} - \Psi_{rq} i_{sd}) - \frac{B}{J} \omega_r - \frac{T_L}{J} \quad (12)$$

where T_L and B designate the load torque and the friction factor, respectively, and the inertia is symbolized with J . The angle of the synchronously rotating frame, θ , is defined as:

$$\frac{d\theta}{dt} = \omega_g \quad (13)$$

3.2. Discrete-Time Model of the Induction Machine

The model predictive control requires the system model in the discrete-time domain. The Forward Euler discretization technique defined by (14) can be applied to obtain discrete-time expressions of continuous differential equations [51,52].

$$\frac{d\beta(t)}{dt} = \frac{\beta(k+1) - \beta(k)}{T_s} \quad (14)$$

where T_s implies the sampling period and the time duration between the two consecutive samplings. The discrete-time domain model of the stator current components is obtained as in (15) and (16):

$$i_{sd}(k+1) = \left(1 - \frac{L_r R_s + \frac{L_m^2}{\tau_r}}{\sigma L_s L_r} T_s\right) i_{sd}(k) + T_s \left(\omega_g(k) i_{sq}(k) + \frac{L_m}{\tau_r \sigma L_s L_r} \Psi_{rd}(k)\right) + T_s \left(\frac{L_m}{\sigma L_s L_r} \omega_r(k) \Psi_{rq}(k) + \frac{v_{sd}(k)}{\sigma L_s}\right) \quad (15)$$

$$i_{sq}(k+1) = \left(1 - \frac{L_r R_s + \frac{L_m^2}{\tau_r}}{\sigma L_s L_r} T_s\right) i_{sq}(k) + T_s \left(-\omega_g(k) i_{sd}(k) + \frac{L_m}{\tau_r \sigma L_s L_r} \Psi_{rq}(k)\right) - T_s \left(\frac{L_m}{\sigma L_s L_r} \omega_r(k) \Psi_{rd}(k) - \frac{v_{sq}(k)}{\sigma L_s}\right) \quad (16)$$

The discrete-time model to estimate rotor flux components is obtained as in (17) and (18):

$$\Psi_{rd}(k) = \Psi_{rd}(k-1) \left(1 - \frac{T_s}{\tau_r}\right) + T_s \left(\frac{L_m}{\tau_r} i_{sd}(k-1) + (\omega_g(k-1) - \omega_r(k-1)) \Psi_{rq}(k-1)\right) \quad (17)$$

$$\Psi_{rq}(k) = \Psi_{rq}(k-1) \left(1 - \frac{T_s}{\tau_r}\right) + T_s \left(\frac{L_m}{\tau_r} i_{sq}(k-1) - (\omega_g(k-1) - \omega_r(k-1)) \Psi_{rd}(k-1)\right) \quad (18)$$

The discrete-time models for the angle of the synchronously rotating frame $\theta(k)$ with respect to fixed-frame and the synchronous angular speed $\omega_g(k-1)$ are obtained as in (19) and (20) respectively.

$$\theta(k) = \theta(k-1) + T_s \omega_g(k-1) \quad (19)$$

$$\omega_g(k-1) = \frac{L_m}{\tau_r \Psi_{rd}} i_{sq}(k-1) + \omega_r(k-1) = \frac{L_m}{\tau_r \Psi_{rd}^*} i_{sq}(k-1) + \omega_r(k-1) \quad (20)$$

4. Control Strategy

The schematic representation of the overall control strategy is shown in Figure 2. The IFOC is adopted to generate the reference signals for the IM's speed and torque regulation, and it operates as an outer loop. The proposed M²PC method is cascaded to IFOC to generate the required gate signals and serves as an inner loop.

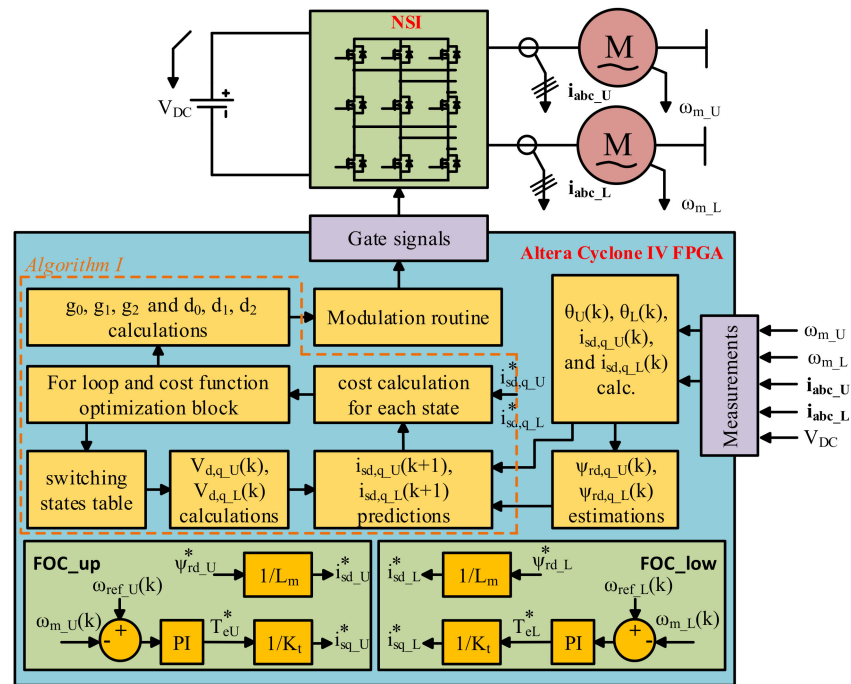


Figure 2. The scheme for proposed IFOC based M²PC of the NSI drive system. An FPGA (Altera Cyclone IV) is used to accomplish the proposed control.

4.1. Reference Generation in Indirect Field-Oriented Control of IM

One of the vector control-based methods used in machine drivers is field orientation. In this technique, the electromagnetic torque of a three-phase ac motor can be regulated similar to the control of a separately excited dc motor. However, for an induction machine, it is necessary to take into account the slip between the magnetic fields of the stator and rotor, which varies with the load torque. The stator current is split into two current components, i.e., torque- and flux-producing components, to achieve a similar torque control strategy to that of a dc motor. If the rotor flux vector is aligned with the d-axis of the reference frame, then $\Psi_{rq} = 0$ is achieved for the steady-state. Thus, the torque expression (21) becomes more straightforward, and the electrical variables slowly change [53].

$$T_e = \frac{3}{2} \frac{pL_m}{L_r} (\Psi_{rd}i_{sq} - \Psi_{rq}i_{sd}) \quad (21)$$

In order to convert the stator currents measured in the fixed reference frame into the variables in the synchronously rotating reference frame, the angle between these two frames must be found. Aligning the rotor flux with d-axis yields $\Psi_{rq} = d\Psi_{rq}/dt = 0$ for the steady-state. By replacing these terms in (11) with zero, the slip angular frequency ω_{sl} is obtained as in (22). This slip is considered as in (20) to estimate the synchronous angular speed $\omega_g(k-1)$. Then, the Ψ_{rq} asymptotically reaches zero.

$$\omega_{sl} = \omega_g - \omega_r = \frac{L_m}{\tau_r \Psi_{rd}^*} i_{sq} \quad (22)$$

Replacing the Ψ_{rq} with zero in (10) yields (23) for the steady-state, and the field orientation is attained.

$$\Psi_{rd} = L_m i_{sd} \quad (23)$$

Therefore, the magnetizing current component i_{sd} regulates the rotor flux. Even though the reference value of Ψ_{rd}^* , which is set as a constant, is substituted in Equations (20) and (22)

instead of the instantaneous value of Ψ_{rd} , the rotor flux can be aligned with the d-axis. Then, the reference for the magnetizing current component is calculated by:

$$i_{sd}^* = \frac{\Psi_{rd}^*}{L_m} \quad (24)$$

In this case, like that of a separately excited dc motor, the torque term becomes in proportion to the torque current component, i_{sq} , as in (25).

$$T_e = K_t i_{sq} \quad (25)$$

where K_t is the torque constant and $K_t = \frac{3}{2} \frac{p L_m \Psi_{rd}^*}{L_r}$. The reference torque value T_e^* in speed control applications can be generated from the speed error using a proportional-integral (PI) compensator. Then, the torque current reference, i_{sq}^* is calculated by (26).

$$i_{sq}^* = T_e^* / K_t \quad (26)$$

4.2. Proposed Modulated Model Predictive Control of NSI

The proposed modulated model predictive control operates as follows: the controller samples the stator currents, dc bus voltage, and mechanical speeds. First, the stator currents measured in the abc frame are converted into the d - q frame variables. Then, the future values of the d - q frame currents are predicted for each candidate switching vector of the NSI by using (15) and (16). The errors, which are the difference between the reference values and the predicted currents, are evaluated in the cost function (27). Since the cost function employs the variables in the same order of magnitude, there is no need for a weighting factor to control the different objectives properly.

$$g = \sum_{j=d,q} \left| i_{sj_up}^* - i_{sj_up} \right|^2 + \sum_{j=d,q} \left| i_{sj_low}^* - i_{sj_low} \right|^2 \quad (27)$$

In modulation techniques like the space-vector strategy developed for the voltage-source inverter, specific switching vectors (two active vectors + one zero vector) are applied to synthesize the sinusoidal voltage/current during the switching period. In this study, a modulated model predictive control strategy is proposed for the NSI. In Algorithm 1, the procedure to select the optimum switching vectors used during the one switching cycle is summarized. Accordingly, one of the zero vectors from Group I of Table 1 is selected as the zero vector, S_0 . Since the switching states of Group I produce the same voltages, thus the same cost, for the load stages, the index of 'for loop' is started from 3 to eliminate the repeatedly calculated same cost. Hence, the cost calculated according to (27) is stored as g_0 . The vectors that provide the lowest two costs from the remaining vectors sw_4 – sw_{27} are considered g_1 and g_2 , and the corresponding switching vectors are kept as S_1 and S_2 . As can be observed from Algorithm 1, there is no need to determine the sector of the space vector by optimizing an additional cost function. The proposed technique differs from traditional M²PC methods with this feature. The next step is to calculate the duty cycles d_0 , d_1 , and d_2 for each of these switching states S_0 , S_1 , and S_2 . The sum of the duty cycles over a single switching cycle must be equal to 1 as stated by (28).

$$d_0 + d_1 + d_2 = 1 \quad (28)$$

Algorithm 1 The routine for selecting the best two switching states and calculating the associated duty cycles

```

1: for  $i = 3:27$ 
2:   calculate cost: Equation (27)
3:   if ( $i = 3$ ) then
4:      $g_0 = \text{cost}$ 
5:   else
6:     if ( $\text{cost} < g_1$ ) then
7:        $g_2 = g_1$ 
8:        $g_1 = \text{cost}$ 
9:        $\text{index\_of\_}g_2 = \text{index\_of\_}g_1$ 
10:       $\text{index\_of\_}g_1 = i$ 
11:     elseif ( $\text{cost} < g_2$ ) then
12:        $g_2 = \text{cost}$ 
13:        $\text{index\_of\_}g_2 = i$ 
14:     end if
15:   end if
16: end for
17: calculate duty cycles;  $d_0, d_1, d_2$ : Equation (30)
18:  $S_0 = [1 \ 1 \ 1; 0 \ 0 \ 0; 1 \ 1 \ 1]$ 
19:  $S_1 = \text{sw}(:, :, \text{index\_of\_}g_1)$ 
20:  $S_2 = \text{sw}(:, :, \text{index\_of\_}g_2)$ 
21: if ( $\text{sum}(\text{abs}(S_0 - S_1)) > \text{sum}(\text{abs}(S_0 - S_2))$ ) then
22:    $\text{temp\_1} = S_1$ 
23:    $S_1 = S_2$ 
24:    $S_2 = \text{temp\_1}$ 
25:    $\text{temp\_2} = d_1$ 
26:    $d_1 = d_2$ 
27:    $d_2 = \text{temp\_2}$ 
28: end if

```

Their allocation over the switching cycle is determined as the reciprocal of the corresponding cost by (29):

$$d_0 = \frac{k}{g_0}, d_1 = \frac{k}{g_1}, d_2 = \frac{k}{g_2} \quad (29)$$

Simultaneously solving (28) and (29) yields (30), which allows calculation of the associated duty cycle for each vector.

$$\begin{aligned} d_0 &= \frac{T_s g_1 g_2}{g_0 g_1 + g_0 g_2 + g_1 g_2} \\ d_1 &= \frac{T_s g_0 g_2}{g_0 g_1 + g_0 g_2 + g_1 g_2} \\ d_2 &= \frac{T_s g_0 g_1}{g_0 g_1 + g_0 g_2 + g_1 g_2} \end{aligned} \quad (30)$$

To complete the proposed modulated model predictive approach, there is a need for an optimized switching vector sequence. With a proper algorithm, the number of switches that change their states simultaneously can be reduced. Thus, this decreases the EMI radiation due to high dv/dt and di/dt at the switched nodes. The routine to determine the optimum switching sequence is shown between rows 18–28 of Algorithm 1. For this aim, the switching state sw_3 is stored as S_0 , and it is the only zero vector allowed in Group I. Then, S_1 and S_2 are compared with S_0 , and the vector with the minimum number of state changes for the switches is kept as S_1 . So, the other one is held as S_2 . The allocation of the switching states over the switching cycle and the optimum switching sequence is shown in Figure 3. When Table 1 is closely inspected, the states between sw_{16} and sw_{21} , which have the on-state condition for the whole middle switches and require more switching to change their state, can be removed from the switching state table to optimize the switching sequence further. Accordingly, the proposed modulation strategy can also be operated with a reduced number of switching states, and the computational burden on the processor can be reduced.

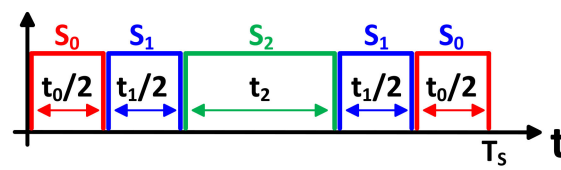


Figure 3. The allocation of the switching period between the associated switching vectors.

5. Comparison with Other Control Strategies

The proposed control strategy was compared with hysteresis-based [21], predictive torque control (PTC) [27], and FCS-MPC [31] techniques regarding the total harmonic distortion in stator currents, torque variance, and execution time. All methods were simulated in Simulink with the parameters listed in Table 2.

Table 2. System parameters.

Description	Parameter	Value
stator inductance	L_s	452.3 mH
rotor inductance	L_r	452.3 mH
mutual inductance	L_m	442.2 mH
pole pair	p	2
stator resistance	R_s	3.9190 Ω
rotor resistance	R_r	4.9618 Ω
friction factor	B	0.002985 N·m·s
inertia	J	0.0131 kg·m ²
stator flux reference	Ψ_s^*	0.61 Wb.
dc-link voltage	V_{DC}	250 V
sampling period of the inner loop	T_s	100 μ s
PI parameter	K_p	1
PI parameter	K_i	16
PI controller discretization period	T_{s_PI}	5 ms

The sampling time T_s is selected as 100 μ s for all methods to make a fair comparison. Accordingly, the modulation frequency for the proposed method is 10 kHz. Table 3 tabulates the THD values for the upper and lower load stages for two test scenarios. The upper motor speed reference is 40 rad/s in these scenarios, while the lower motor speed reference is 25 rad/s. Each motor operates with a no-loading condition (0 N·m) for scenario I. The magnitude of the fundamental current component for the upper motor is 1.37 A @ 12.866 Hz. The magnitude of the fundamental current component for the lower motor is 1.37 A @ 8.055 Hz. For scenario II, the upper motor operates with a load torque of 3 N·m and the lower motor operates with a load torque of 4 N·m, and the magnitude of the fundamental current component for the upper motor is 2.22 A @ 14.928 Hz, the magnitude of the fundamental current component for the lower motor is 2.66 A @ 10.880 Hz. According to Table 3, the proposed M²PC method significantly reduces THD values compared to other methods.

Table 3. Comparison results between different control strategies.

Compared Quantity → Control Strategy↓	Scenario I		Scenario II		Execution Time
	THD% for Upper Load	THD% for Lower Load	THD% for Upper Load	THD% for Lower Load	
Proposed Method	6.79	6.79	5.99	5.25	202.4 μ s
FCS-MPC	18.44	19.64	11.66	9.63	165.7 μ s
PTC	18.98	20.10	11.39	9.79	170.1 μ s
Hysteresis	22.76	23.15	14.48	12.70	67.66 μ s

Each method's controller block execution time is measured with the “Run and Time” menu in Matlab. The increment in the execution time for the proposed method is acceptable

to decrease the THD values. The execution times reported in the table are the run time measured in Matlab for a single loop of each algorithm. These are used to compare fairly each algorithm's computational burden and are not actual execution times for the code blocks that implement the algorithms on a real DSP or FPGA.

To better show their performance in terms of computational burden, the figure of merit “Computational Burden Rate (CBR)” can be defined as:

$$\text{CBR} = \frac{\text{execution time}}{\text{sampling period}} \quad (31)$$

The proposed method can be compared with the FCS-MPC technique in CBR. For this aim, the sampling frequency of the FCS-MPC can be increased until it provides similar performance in THD compared to the proposed method. Accordingly, the CBR for the proposed method is less than that of the FCS-MPC, as listed in Table 4.

Table 4. Comparison results in terms of the computational burden.

Compared Quantity → Control Strategy ↓	Scenario I		Scenario II		Sampling Period	Execution Time	CBR
	THD% for Upper Load	THD% for Lower Load	THD% for Upper Load	THD% for Lower Load			
Proposed Method	6.79	6.79	5.99	5.25	100 μ s	202.4 μ s	2.2024
FCS-MPC	7.50	8.03	4.59	4.37	40 μ s	165.7 μ s	4.1430

To prove the improved steady-state performance of the proposed M²PC method, it was compared with the FCS-MPC in torque ripples. The upper motor's electromagnetic torque is plotted in Figure 4 for both control methods, and it is observed that the ripple in the torque is less for the M²PC. It can also be concluded from Figure 4 that the transient responses for both techniques are similar as they follow the same trajectories during the start-up and load step transitions.

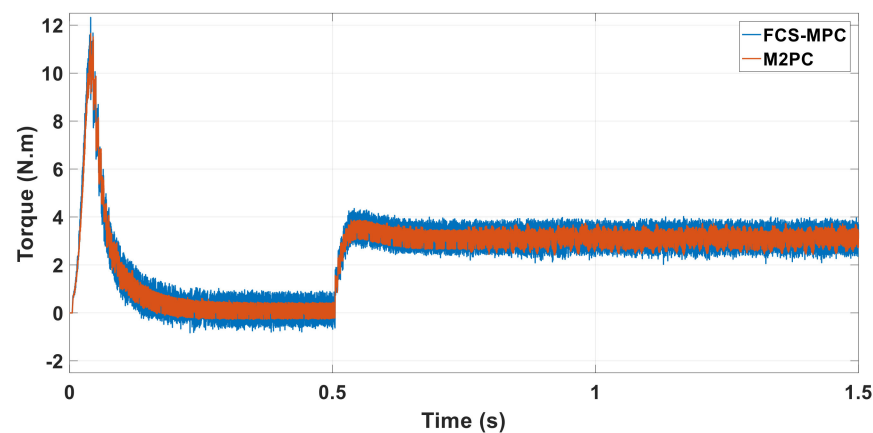


Figure 4. The comparison of oscillations in the electromagnetic torque for both control methods.

Figure 5 compares the spectral content of the stator currents of the lower motor for scenario II when the NSI is controlled with both M²PC and FCS-MPC methods. The M²PC has a spectral content with reduced order of magnitudes since it is a strategy with a fixed modulation frequency. In comparison, FCS-MPC has a significant spectral content covering broadband since it is a strategy without a modulator stage and operates with a variable frequency operation. As a result, the EMI problems are reduced, and the required filter designs are much simpler for the proposed modulated technique.

The traditional FCS-MPC employs only one voltage vector during the sampling period and results in large current ripple and variable switching frequency since the same voltage vector can be applied during the consecutive sampling periods. Conversely, the proposed

method considers the switching sequence of three voltage vectors during each sampling period. Therefore, an average voltage vector is achieved for the proposed strategy. The use of an average voltage vector offers a reduction in current ripple, also resulting in torque ripple. The other problem of the variable switching frequency operation of the FCS-MPC is that the spectral harmonic is randomly distributed. Hence, estimating the harmonic contents in the frequency spectrum is quite challenging, and selecting the filter parameter is not a trivial task. On the other hand, the proposed method uses the modulator to generate the gate signals. Therefore, the major parts of the harmonic content are located on the order of the switching frequency and its harmonics. Because of all these reasons, the estimation of the filter parameter is much easier for the proposed control technique, and the benefit of using a modulator is fully exploited [39].

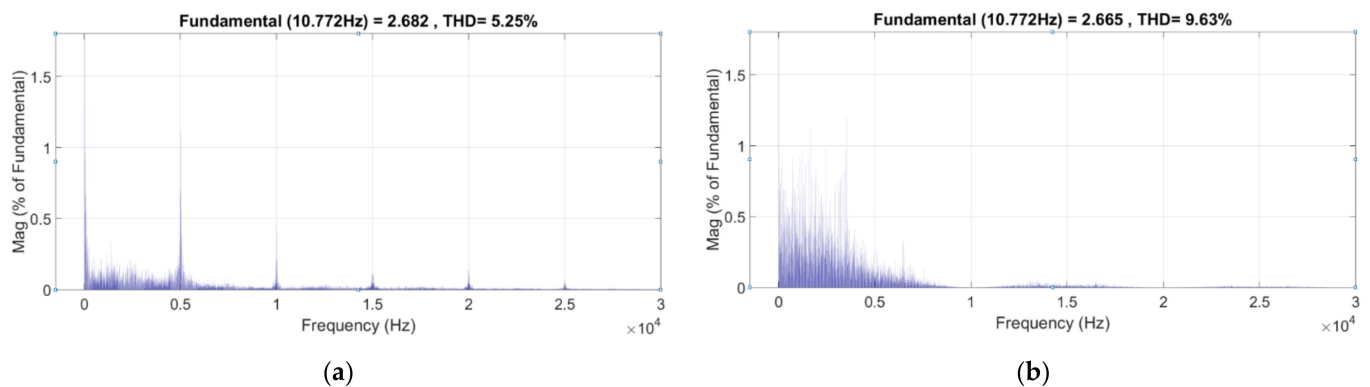


Figure 5. Comparing the spectral content of stator currents: (a) M²PC; (b) FCS-MPC.

6. Experimental Results

In this section, experimental results are presented to show that the proposed control strategy successfully achieves the independent operation of two induction machines driven by NSI. The test bench built in the laboratory is shown in Figure 6, and the system parameters are tabulated in Table 2. In the test bench, the separately excited compound dc machines, whose shafts are mechanically coupled to IMs, are used to test the proposed control strategy under different loading conditions. The incremental encoders with 5000 ppr are also coupled to the shafts of the IMs to measure the rotor mechanical speeds. The proposed M²PC method, covering the whole process from reading the speeds of IMs and ADCs (for measuring the currents and the dc bus voltage) to generating the required gate signals, is embedded into the Altera Cyclone IV FPGA (Terasic DE0-Nano Board).

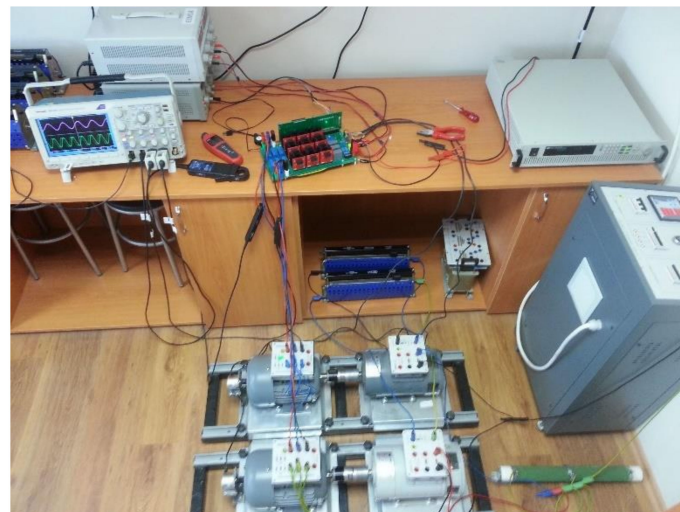


Figure 6. Overview of the experimental test bench.

The proposed M²PC method is first tested to observe its potency to synthesize the sinusoidal currents. Figure 7a shows the phase currents (i_{aU} and i_{aL}) with different frequencies for both output stages, whereas Figure 7b shows the phase currents (i_{aU} and i_{aL}) when they are forced to have the same frequency with a phase displacement of 60°. The leg-to-leg voltages (v_{abU} and v_{abL}) are also shown in Figure 7 for consistency. It can be concluded from the results in Figure 7 that the proposed M²PC can be used to drive both the two three-phase loads independently and a single symmetric/asymmetric six-phase load.

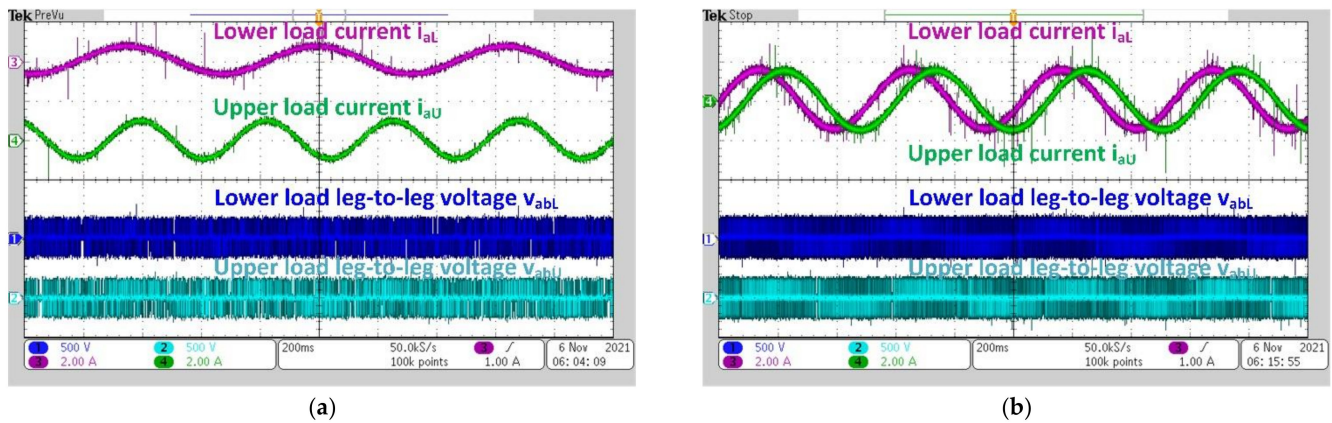


Figure 7. Stator currents (i_{aU} and i_{aL}) and leg-to-leg voltages (v_{abU} and v_{abL}) for both output stages. Scope channel 1: Leg-to-leg voltage v_{abL} of the lower stage. Scope channel 2: Leg-to-leg voltage v_{abU} of the upper stage. Scope channel 3: Phase current, i_{aL} , of the lower stage. Scope channel 4: Phase current, i_{aU} , of the upper stage. The current scale is 2 A per division. The voltage scale is 500 V per division. (a) stator current waveforms at different frequencies for both outputs; (b) stator current waveforms for the symmetrical six-phase reference with a phase displacement of 60°.

Figure 8 presents the oscilloscope screenshots showing the experimentally measured one-phase current waveforms of the lower and upper load stages for the cases where the NSI is controlled with the proposed M²PC method and conventional FCS-MPC. Both techniques are operated with a sampling period of 100 μ s. THD values and spectral contents of the currents obtained for both methods are also compared in Figure 8. The proposed modulated-based technique reduces the THD values and spectral content compared to FCS-MPC.

The proposed method was also experimentally verified in terms of speed reference tracking performance regarding the start-up and applied speed reference ramp transitions. Both machines' speeds (ω_{mU} and ω_{mL}) and stator currents (i_{aU} and i_{aL}) are measured and recorded through a digital oscilloscope. Figure 9a presents the measured waveforms for start-up transition. In this test, the speed references of both machines are initially set to 0 rad/s. Afterwards, a user-defined button is used to initiate the operation. Then the reference speeds, which are driven to the controller's comparator blocks, reach their final values with a ramp function having a slope of 10 rad/s per 1 s. The upper machine speed reference is set to 25 rad/s while the lower machine is set to 40 rad/s. According to the waveforms in Figure 9a, each machine follows its speed reference, and the stator currents for both machines are synthesized to correspond to their speed profiles. Figure 9b presents the experimental waveforms when the reference speeds are varied with a ramp function having a slope of 10 rad/s per 0.2 s. In this test scenario, a user-defined button triggers the speed reference transition from 25 rad/s to 35 rad/s for the upper machine and from 40 rad/s to 20 rad/s for the lower machine. As observed from Figure 9, each machine tracks its speed profile without any interaction between them.

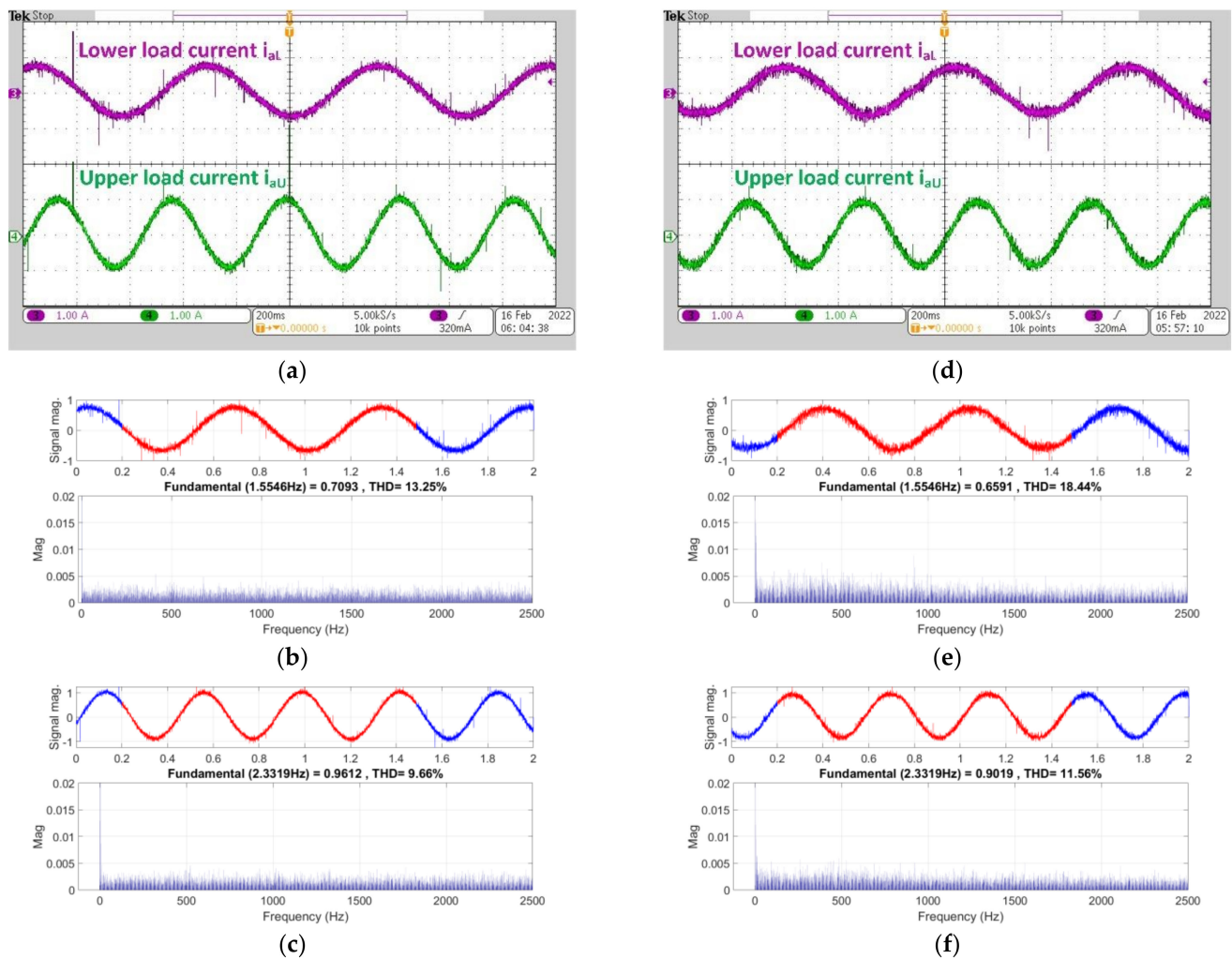


Figure 8. Comparison of the experimental current waveforms, THD results, and current spectrums for two control strategies: the left column is for the M²PC, the right column is for the FCS-MPC. (a) Experimental current waveforms for M²PC. (b) THD value and spectrum of the i_{aL} for M²PC. (c) THD value and spectrum of the i_{aU} for M²PC. (d) Experimental current waveforms for FCS-MPC. (e) THD value and spectrum of the i_{aL} for FCS-MPC. (f) THD value and spectrum of the i_{aU} for FCS-MPC.

The proposed controller is also tested under incremental load torque steps. Figure 10a presents the experimental waveforms when both IMs are loaded at the same instant. In this test, the upper machine speed reference is set to 25 rad/s, with 40 rad/s for the lower machine. As observed from Figure 10a, the speed waveforms have deviations from their reference trajectories at the time instant of the load torque step. Subsequently, the outer loop compensates the speed error and increases the reference for the torque-current component i_{sq}^* to meet the new torque demand. To investigate the response of the proposed control strategy to the asymmetric loading, one of the IM is exposed to the torque load step while the other one is not. According to the results presented in Figure 10b,c, the proposed controller can independently regulate both the machines' speeds and torques without cross-coupling and unwanted interaction.

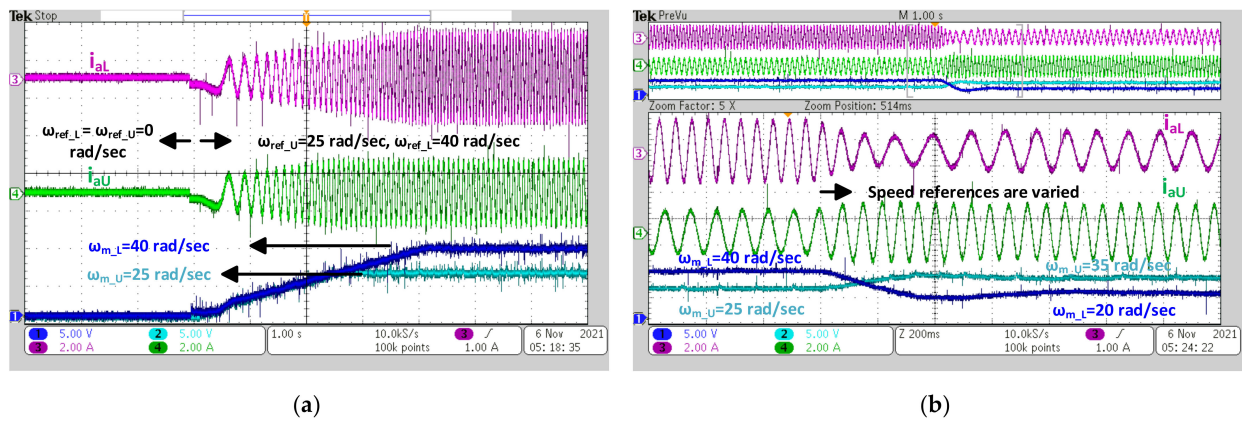


Figure 9. Experimental waveforms for speeds (ω_{m_U} and ω_{m_L}) and one-phase stator current (i_{aU} and i_{aL}) for each machine. Scope channel 1: Upper machine speed, ω_{m_L} . Scope channel 2: Lower machine speed, ω_{m_U} . Scope channel 3: Lower machine stator current, i_{aL} . Scope channel 4: Upper machine stator current, i_{aU} . The speed scale is 20 rad/s per division. The current scale is 2 A per division. (a) during start-up transient. (b) during the speed references the machines are varied with a ramp function having a slope of 10 rad/s per 0.2 s.

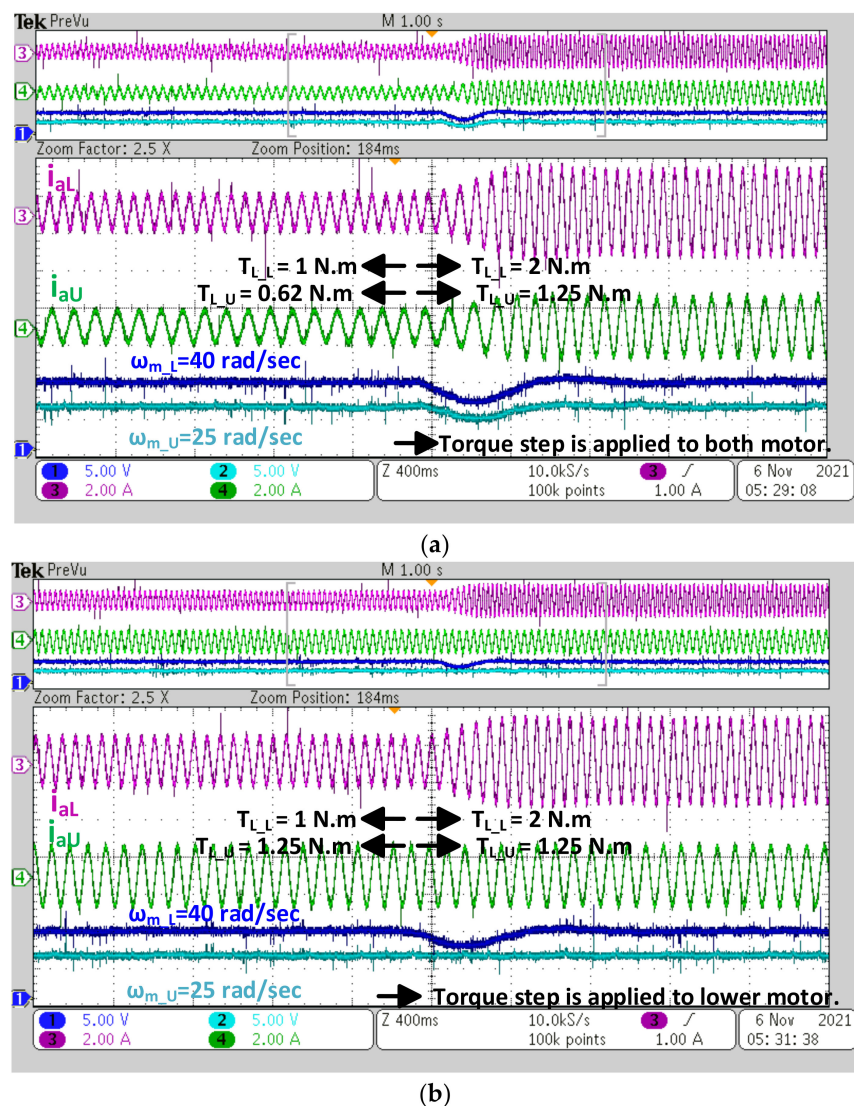
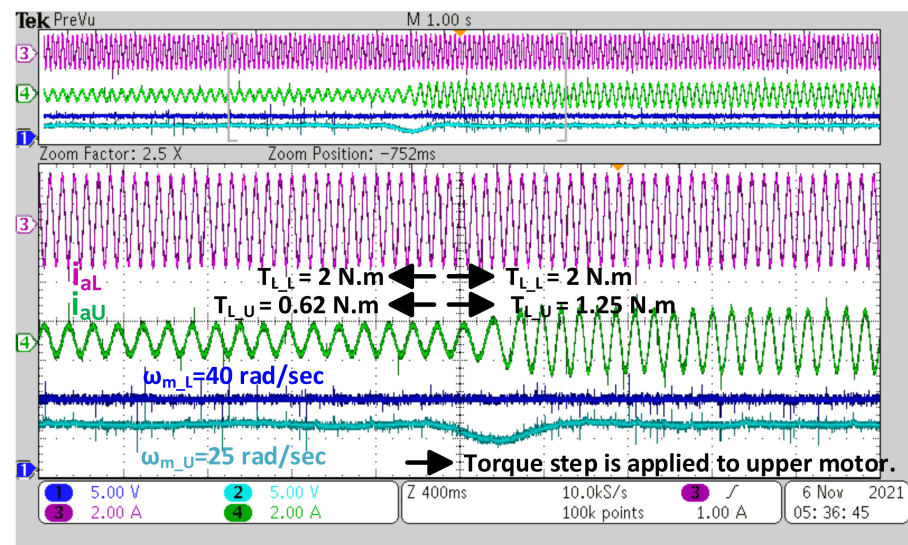


Figure 10. Cont.



(c)

Figure 10. Experimental waveforms for speeds (ω_{m_U} and ω_{m_L}) and one-phase stator currents (i_{aU} and i_{aL}) for each machine under the applied torque steps. Scope channel 1: Lower machine speed, ω_{m_L} , is 40 rad/s. Scope channel 2: Upper machine speed, ω_{m_U} , is 25 rad/s. Scope channel 3: Lower machine stator current, i_{aL} . Scope channel 4: Upper machine stator current, i_{aU} . The speed scale is 20 rad/s per division. The current scale is 2 A per division. (a) Torque steps of 2 N·m and 1.25 N·m are applied to the lower and upper machines, respectively. (b) A torque step of 2 N·m is applied to the lower machine while the upper machine is operated under a fixed torque of 1.25 N·m. (c) A torque step of 1.25 N·m is applied to the upper machine while the lower machine is operated under a fixed torque of 2 N·m.

7. Conclusions

In this study, a modulated model predictive control is proposed for the first-time for the NSI to improve the steady-state performance. Driving the two separate IMs with NSI was chosen as a case study to verify the proposed approach. Indeed, the proposed M²PC approach can be applied to any energy conversion system employing NSI as far as the discrete-time model of the system can be derived. The comparative simulation results for NSI driven two IMs prove that the proposed M²PC reduces the THD in stator currents and ripples in the electromagnetic torque of the machine compared to other methods. Moreover, the modulated approach allows the controller to operate with lower sampling frequencies. Thus, the computational burden rate (CBR) on the processor, which must be completed within a particular sampling period, is reduced. Especially for energy conversion systems with lower electrical time constants (systems with lower inductive characteristics), the M²PC is not an option but a necessity since the synthesized sinusoidal currents require very challenging sampling periods (around 10–20 μ s) to obtain acceptable THD levels. The M²PC also provides a spectral content with reduced order of magnitudes for stator current compared to the FCS-MPC technique, and the required filter design becomes simpler. The proposed M²PC was experimentally verified on a lab-scale NSI prototype, and the experimental results prove that the independent operation of two separate IMs can be achieved with the proposed M²PC.

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Conflicts of Interest: The author declares no conflict of interest.

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