

Article

Efficiency Analysis of Predictive Current Control of a Multiphase Drive

Esteban Marsal ^{1,*} , Francisco Colodro ¹  and Antonio Visioli ² ¹ Departamento de Ingeniería Electrónica, Universidad de Sevilla, 41092 Seville, Spain; pcolr@us.es² Dipartimento di Ingegneria Meccanica e Industriale, Università di Brescia, 25123 Brescia, Italy; antonio.visioli@unibs.it

* Correspondence: lmarsal@us.es

Abstract

In Finite Control Set Model Predictive Control for multiphase drives, weighting factors govern the relative priority among the multiple objectives included in the cost function. Although their influence on control performance, harmonic characteristics, and switching behavior has been extensively studied, their impact on energy efficiency remains largely unexplored. This paper addresses this gap through a comprehensive experimental assessment of the influence of cost function weighting factors on the energy efficiency of a five-phase induction motor drive. A dedicated laboratory test bench is used to evaluate drive efficiency across a broad range of operating conditions by systematically varying mechanical speed, load level and weighting factor combinations. The results reveal that cost function weighting factor tuning leads to efficiency variations of up to 7.6 percentage points depending on the operating regime. Furthermore, the study identifies an optimal parameter region that provides consistently high efficiency across the tested conditions.

Keywords: finite control set; model predictive control; electrical drives; multiphase motor; energy efficiency; weighting factors; cost function; induction motor

1. Introduction

Model Predictive Control (MPC) for variable speed drives have been the subject of research for several decades [1]. Current approaches can be classified into two categories according to the modulation strategy employed. The first uses a predictive controller whose output is a continuous voltage reference synthesized by the Voltage Source Inverter (VSI) through Pulse Width Modulation (PWM) [2]. Since the controller output is not quantized, this approach is referred to as Continuous Control Set Model Predictive Control (CCSMPC) [3].

The second category, known as Finite Control Set MPC (FCSMPC) or Finite State MPC (FSMPC), directly selects among the available switching states of the VSI without requiring a modulation stage [4]. This provides higher control bandwidth at the cost of a distributed harmonic spectrum [5]. The real-time deployment of FCSMPC faces two main challenges. The first is computational complexity, particularly in multiphase drives, where the large number of switching configurations increases the evaluation burden [6]. Although fast optimization methods have been proposed to alleviate this issue [7–9], they are applicable only to specific cost function formulations.

The second challenge is the selection of the weighting factors (WFs) that appear in the cost function. Some recent works propose their elimination altogether [10,11], while



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others exploit them to achieve flexible control objectives. Online tuning methods have also been developed [12–14], although these solutions are typically limited to cost functions that include a penalty for stator current content in the harmonic subspace. This particular choice is relevant because the harmonic plane currents do not produce torque but contribute to Joule losses in the drive [15]; reducing them should therefore improve efficiency. However, the voltage modulation performed by the VSI through switch commutations introduces an additional source of losses [16]. Including a penalty term for switching transitions addresses this issue but increases computational load [17,18].

An alternative approach adopted in several studies combines the dead-beat concept with a multi-vector strategy, in which more than one voltage vector is applied during each sampling period. This formulation avoids the need for WF by zeroing the harmonic content in open loop and dropping the commutation penalty [19]. The resulting method eliminates WF selection and allows shaping the loss distribution [20], but leads to an increase in switching frequency as a direct consequence of the multi-vector approach [21].

Importantly, the two loss mechanisms (harmonic currents and switching transitions) cannot be addressed independently due to the inherent coupling between control objectives in FCSMPC [3]. Determining the WFs that minimize total losses must therefore be carried out concurrently, which is not straightforward since analytical loss expressions are cumbersome to handle in practice [20,22].

This paper presents a comprehensive experimental assessment of the influence of the WF λ_{xy} and λ_{sc} , respectively associated with current regulation in the secondary subspaces and switching frequency reduction, on the energy efficiency of a five-phase induction motor drive [23]. A dedicated laboratory test bench is used to evaluate drive efficiency across a broad range of operating conditions by systematically varying mechanical speed, load level and WF combinations.

The novelty of this work lies in the consideration of overall energy efficiency as a criterion for WF tuning, an aspect that has not been addressed in the literature except in preliminary studies by the authors [24,25]. Compared to those earlier works, the present paper extends the analysis to the joint interaction of both WFs, examines their combined impact on overall efficiency and identifies optimal operating ranges.

2. Model Predictive Control for Multiphase Induction Motors

The FCSMPC strategy employed in this work regulates the stator currents of the induction motor (IM) by directly selecting the switching states of the VSI at each sampling instant. An outer feedback loop provides the speed regulation. A discrete-time model of the IM is used to predict the stator currents at future sampling instants, and these predictions are evaluated for all admissible voltage vectors. The switching state that minimizes a predefined cost function is then applied to the inverter.

Figure 1 shows the control scheme adopted in this work. The control loop begins by comparing the reference speed, ω^* , with the measured motor speed, ω . The speed error is processed by a PI controller in the outer loop, whose output generates the torque reference. The block T_p^{-1} , corresponding to the inverse Park transform, converts the rotational system references to the stationary reference frame $\alpha\beta xy$, thus obtaining the reference currents. Simultaneously, the measured motor currents are projected from the phase system to the stationary system using the Clarke transform T_C , obtaining the components $\alpha\beta xy$.

The FCSMPC controller uses a discrete motor model to predict the evolution of the currents associated with each switching state of the VSI. The predicted currents are evaluated for all possible states of the converter, and the switching state that minimizes this function is selected and applied directly to the VSI to power the five-phase induction motor, S^{opt} .

The five-phase, two-level voltage source inverter generates 32 possible voltage vectors, depending on the switching state of each leg. Based on their magnitude, these vectors can be classified into four groups: zero vectors, small vectors ($2V_{DC}(\sqrt{5}-1)/5$), medium vectors ($2V_{DC}/5$), and large vectors ($2V_{DC}(\sqrt{5}+1)/5$). These magnitudes are directly determined by V_{DC} and by the coordinate transformation used to project the switching states onto the α - β and x - y planes. Therefore, they are fixed and do not depend on the operating point or on the control strategy adopted. Figure 2 shows the voltage vectors generated in the α - β and x - y planes. Vectors in the α - β plane are directly related to the machine components that produce torque, while their projections onto the x - y plane are associated with spurious secondary current components that contribute to additional losses in the motor windings. As can be seen, switching states that produce large vectors in the α - β plane result in the smallest projections in the x - y plane, and vice versa. This inverse relationship is characteristic of multiphase drives and is particularly relevant to the control strategy: since no switching state can simultaneously provide a large vector in both planes, the controller must strike a balance between tracking the current references in the primary plane and limiting the secondary-plane currents responsible for these losses.

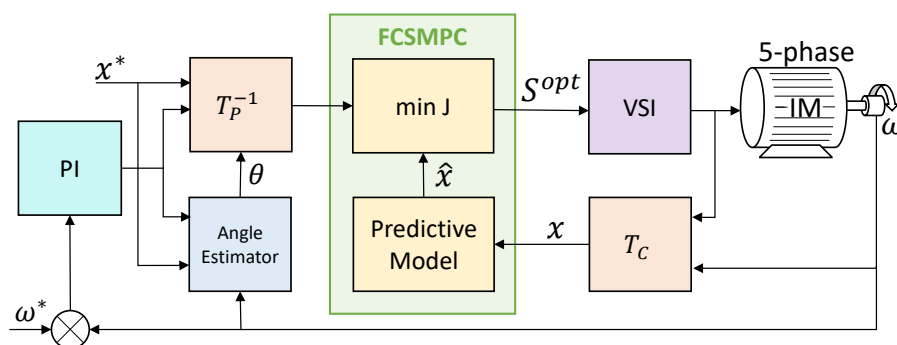


Figure 1. Scheme of the FCSMPC based control system used for the drive of a five-phase IM.

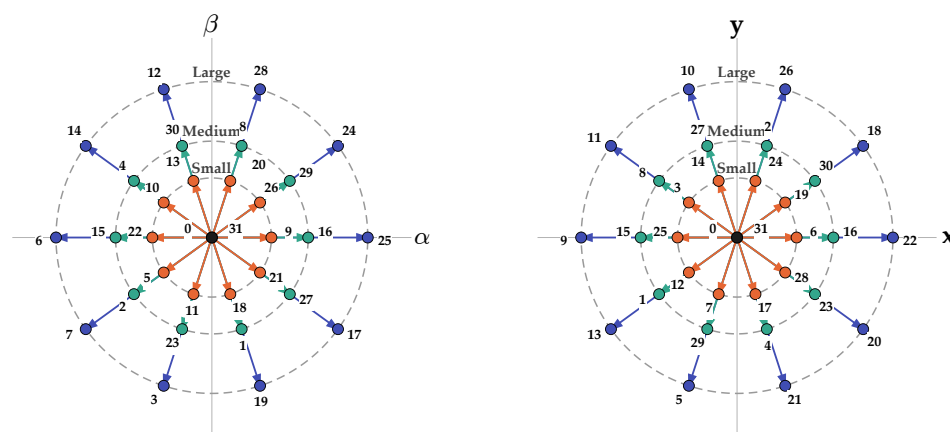


Figure 2. Zero, small, medium and large vectors in the α - β (left) and x - y (right) planes for a five-phase converter.

The predictions used by the FCSMPC algorithm are obtained from a discrete-time model derived by applying Euler discretisation to the continuous-time equations of the IM. The one-step-ahead prediction of the stator current vector $i = (i_{s\alpha}, i_{s\beta}, i_{sx}, i_{sy})$ is expressed as

$$\hat{i}(k+1|k) = A i(k) + B v(k|k-1) + \hat{G}(k|k), \quad (1)$$

where $\hat{i}(k+1|k)$ denotes the current prediction computed at the discrete-time instant k for the following instant $(k+1)$, and $v(k|k-1)$ is the stator voltage generated by the VSI according to the switching state selected at $(k-1)$.

The matrices A and B depend on the rotor electrical speed, the stator and rotor resistances, the stator and rotor inductances, the stator leakage inductance and the mutual inductance. Matrix $\hat{G}$ is an additional term that accounts for the rotor-current contribution, which is generally not directly measurable. Nevertheless, several methods have been proposed to estimate this contribution [26–28].

The matrices A and B are given by

$$A = \begin{pmatrix} -a_{s2} & a_{m4} & 0 & 0 & a_{r4} & a_{l4} \\ -a_{m4} & -a_{s2} & 0 & 0 & -a_{l4} & a_{r4} \\ 0 & 0 & -a_{s3} & 0 & 0 & 0 \\ 0 & 0 & 0 & -a_{s3} & 0 & 0 \\ a_{s4} & -a_{m5} & 0 & 0 & -a_{r5} & -a_{l5} \\ a_{m5} & a_{s4} & 0 & 0 & a_{l5} & -a_{r5} \end{pmatrix}, \quad (2)$$

$$B = \begin{pmatrix} c_2 & 0 & 0 & 0 \\ 0 & c_2 & 0 & 0 \\ 0 & 0 & c_3 & 0 \\ 0 & 0 & 0 & c_3 \\ -c_4 & 0 & 0 & 0 \\ 0 & -c_4 & 0 & 0 \end{pmatrix}, \quad (3)$$

where the coefficients of A are defined as

$$\begin{aligned} a_{s2} &= R_s c_2, & a_{s3} &= R_s c_3, & a_{s4} &= R_s c_4, \\ a_{r4} &= R_r c_4, & a_{m4} &= M c_4 \omega_r, & a_{l4} &= L_r c_4 \omega_r, \\ a_{r5} &= R_r c_5, & a_{m5} &= M c_5 \omega_r, & a_{l5} &= L_r c_5 \omega_r, \end{aligned} \quad (4)$$

and the constants c_x depend on the machine parameters as

$$c_1 = L_s L_r - M^2, \quad c_2 = \frac{L_r}{c_1}, \quad c_3 = \frac{1}{L_{ls}}, \quad c_4 = \frac{M}{c_1}, \quad c_5 = \frac{L_s}{c_1}. \quad (5)$$

Matrix A depends explicitly on the rotor electrical speed ω_r through the coefficients a_{l4} , a_{l5} , a_{m4} and a_{m5} ; since the mechanical dynamics are considerably slower than the electrical dynamics, ω_r can be assumed constant within a single sampling period, so that A is treated as time invariant for the purpose of the discrete time prediction.

The cost function considered in this work is defined as

$$J(k+2) = \|\hat{e}_{\alpha\beta}(k+2)\|^2 + \lambda_{xy} \|\hat{e}_{xy}(k+2)\|^2 + \lambda_{sc} S(k+1), \quad (6)$$

where $\hat{e}_{\alpha\beta}$ represents the predicted current-tracking error in the α - β plane, $\hat{e}_{xy}$ denotes the predicted error in the x - y plane and $S(k+1)$ is the number of switching transitions produced by the candidate control action. This last term is defined as

$$S(k+1) = \sum_{h=1}^n |u_h(k+1) - u_h(k)|, \quad (7)$$

where $u_h(k)$ and $u_h(k+1)$ represent the switching states of the h -th leg of the VSI at instants k and $k+1$, respectively. Therefore, the absolute value $|u_h(k+1) - u_h(k)|$ indicates whether

a switching transition occurs in the corresponding inverter leg when the candidate control action $u(k+1)$ is applied.

In this formulation, λ_{xy} weights the importance assigned to the suppression of the non-torque-producing current components in the x - y plane, whereas λ_{sc} penalizes switching transitions and is therefore directly related to the switching activity of the inverter. Variations in λ_{xy} and λ_{sc} may simultaneously affect current tracking, secondary-plane current components, switching frequency and the resulting loss distribution of the drive.

The stator current reference trajectories provided by the outer control loop are sinusoidal and are given by

$$i_s^*(t) = \begin{pmatrix} i_{s\alpha}^*(t) \\ i_{s\beta}^*(t) \\ i_{sx}^*(t) \\ i_{sy}^*(t) \end{pmatrix} = \begin{pmatrix} I_s^* \sin(\omega_e t) \\ I_s^* \cos(\omega_e t) \\ 0 \\ 0 \end{pmatrix}, \quad (8)$$

where I_s^* is the reference stator current amplitude and ω_e is the electrical angular frequency. The current references in the x - y plane are set to zero because these components do not contribute to electromagnetic torque production.

3. Experimental Test Bench

The experimental test bench consists of a five-phase IM, which is the main object of this study, and a direct-current (DC) machine, mechanically coupled through their shafts. This coupling makes it possible to impose different load conditions on the motor under test and to evaluate its dynamic and energetic behavior at different operating points. Figure 3 shows the general scheme of the experimental test bench and its main subsystems.

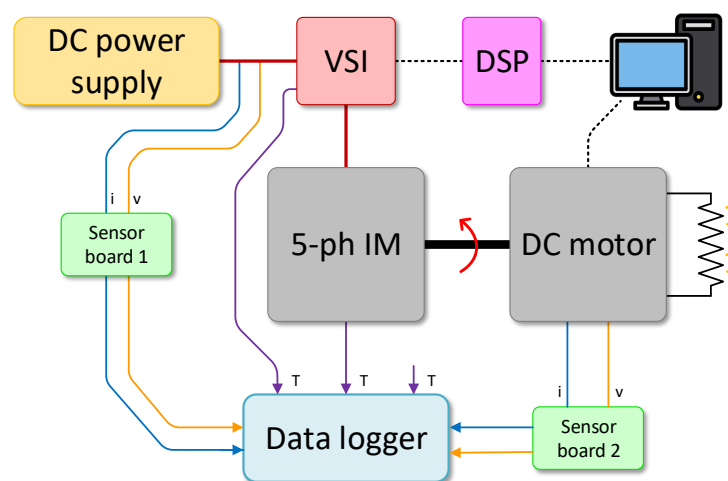


Figure 3. General scheme of the experimental test bench.

The experimental setup includes a DC power supply (DC-link) that supplies the five-phase VSI, implemented using an SKS22FB6U power electronic inverter module, manufactured by Semikron Danfoss Elektronik GmbH & Co. KG, Nuremberg, Germany. The control algorithm execution and the generation of the inverter switching signals are carried out by a control board based on the TMS320F28335, a 32-bit DSP manufactured by Texas Instruments Incorporated, Dallas, TX, USA.

The five-phase IM is mechanically coupled to the DC machine, which operates as a generator during testing. The armature circuit of the DC machine is connected to a resistive load, allowing the mechanical power delivered by the IM to be converted into electrical power and dissipated in the load. The loading conditions can be adjusted by varying the

resistance value. The parameters associated with the laboratory setup are summarized in Table 1.

Table 1. Parameters of the experimental test bench.

Parameter	Value	Units
Stator resistance, R_s	12.85	Ω
Rotor resistance, R_r	4.80	Ω
Stator leakage inductance, L_{ls}	79.93	mH
Rotor leakage inductance, L_{lr}	79.93	mH
Mutual inductance, L_M	681.7	mH
Number of pairs of poles, P	3	-
Direct current voltage, V_{DC}	300	V
Sample time, T_s	80	μs

The data acquisition system consists of a DAS220 datalogger, manufactured by SEFRAM Instruments SAS, Saint-Étienne, France, connected to conditioning and sensing boards specifically designed for the test bench. These boards incorporate AMC1311B voltage sensing amplifiers and AMC1300B current sensing amplifiers, both manufactured by Texas Instruments Incorporated, Dallas, TX, USA, as well as AD8495 temperature sensor amplifiers, manufactured by Analog Devices, Inc., Wilmington, MA, USA. These elements allow the main variables to be recorded during system operation.

The computer connected to the system allows the operation of the test bench to be monitored, the test conditions to be configured, and the data recorded by the acquisition system to be stored. In this way, the test bench enables an experimental evaluation of the energetic behavior of the drive under different speeds, load conditions and control parameters.

4. Experimental Design

The experimental campaign was conducted on the test bench depicted in Figure 4, covering a wide range of operating conditions defined by combinations of mechanical speed, resistive load and control strategy parameters.

The mechanical speed was set to $\omega \in \{100, 200, 500, 750\}$ rpm to cover a wide range of operating points. The resistive load was varied across $R \in \{73, 100, 110, 220\} \Omega$. Lower resistance values demand higher currents from the drive, representing heavier loading conditions, while higher values correspond to lighter loads.

Regarding the parameters of the control strategy, the cost function comprises three terms associated with different physical objectives. The tracking error in the α - β plane is included without a WF (i.e., with an implicit weight equal to 1). Therefore, it serves as the reference term against which the WFs of the remaining two components are adjusted.

The next one, $\lambda_{xy} \in \{0.01, 0.2, 0.5, 1, 5\}$, weights the current tracking error in x - y subspace. This wide range of values was selected to characterize the full spectrum of compensation. At the lowest value (0.01), x - y error is virtually neglected, allowing non-torque-producing currents to flow with little restriction; at the highest value (5), the tracking error in this subspace is weighted more heavily than that of the torque-producing plane, forcing the controller to actively suppress these currents.

The last factor, $\lambda_{sc} \in \{6, 8, 10\} \cdot 10^{-3}$, weights the switching effort associated with the number of converter commutations. Its values were kept comparatively small to avoid degrading tracking performance in the main subspace, favoring instead voltage vectors that provide similar control performance while reducing the number of switching transitions and, consequently, the switching frequency and the associated converter losses.

where N_s is the number of samples, T_s is the sampling period and $n_{sc}(k)$ is the number of switch-state transitions recorded at sample k .

5. Results

The experimental results are presented in two main stages. First, the direct influence of the WFs included in the cost function is analyzed. In particular, the effects of λ_{xy} on the current tracking error in both planes, the fundamental α - β plane and the secondary x - y plane, as well as the effect of λ_{sc} on the inverter switching frequency, are evaluated. This preliminary analysis makes it possible to verify the expected action of each WF before discussing their impact on the overall efficiency of the drive system.

As expected, increasing λ_{xy} reduces the tracking error in the x - y plane, as shown in Figure 5. For very low values of λ_{xy} , the controller assigns little relevance to the secondary subspace, resulting in comparatively high values of e_{xy} . When λ_{xy} is increased from 0.01 to 0.2, a sharp reduction in the error is observed. Beyond this value, however, the improvement becomes much less significant and a slight increase in e_{xy} can even be observed for higher values of λ_{xy} . This behavior highlights the nonlinear nature of the predictive controller and indicates that excessively large values of λ_{xy} do not necessarily provide a proportional improvement in the secondary-plane current tracking. Moreover, the error also exhibits a certain dependence on the operating speed, with higher speeds generally leading to lower e_{xy} values.

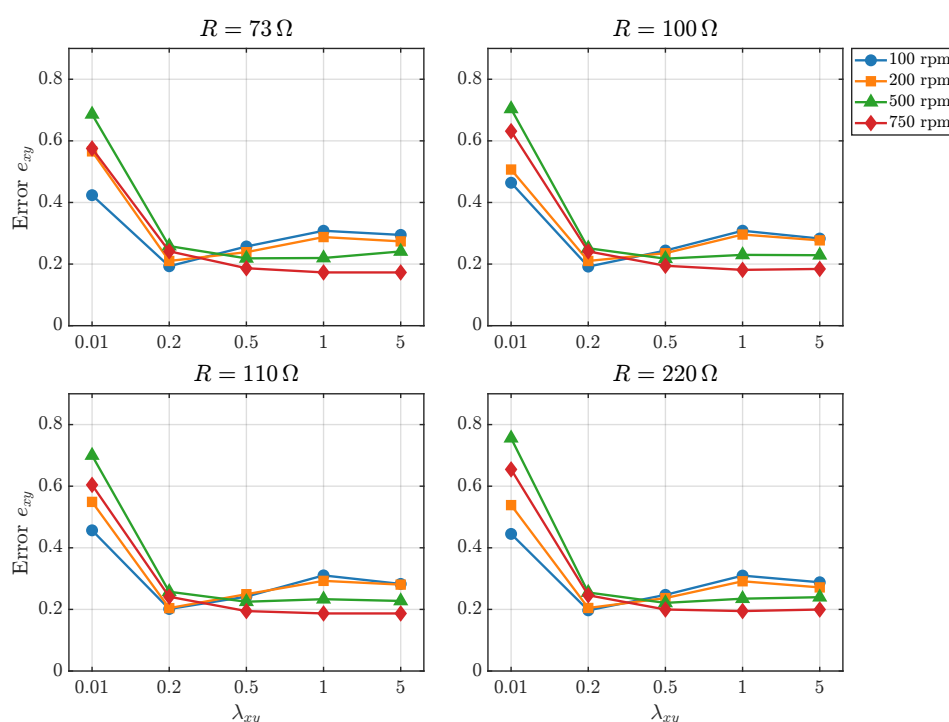


Figure 5. Influence of the control tuning parameter λ_{xy} on the tracking error in the x - y plane.

However, prioritizing the secondary subspace comes at the expense of the fundamental plane performance. As shown in Figure 6, the tracking error $e_{\alpha\beta}$ remains low and stable for λ_{xy} values between 0.01 and 1, but increases sharply when $\lambda_{xy} = 5$.

This degradation is consistent across all speeds and load conditions and is particularly pronounced at higher speeds where the current amplitudes are larger. This result reveals that excessively high values of λ_{xy} shift the control effort away from the fundamental plane, compromising the quality of the torque-producing current components. The combined analysis of e_{xy} and $e_{\alpha\beta}$ therefore suggests that intermediate values of λ_{xy} offer the best compromise between fundamental-plane tracking and secondary-current suppression.

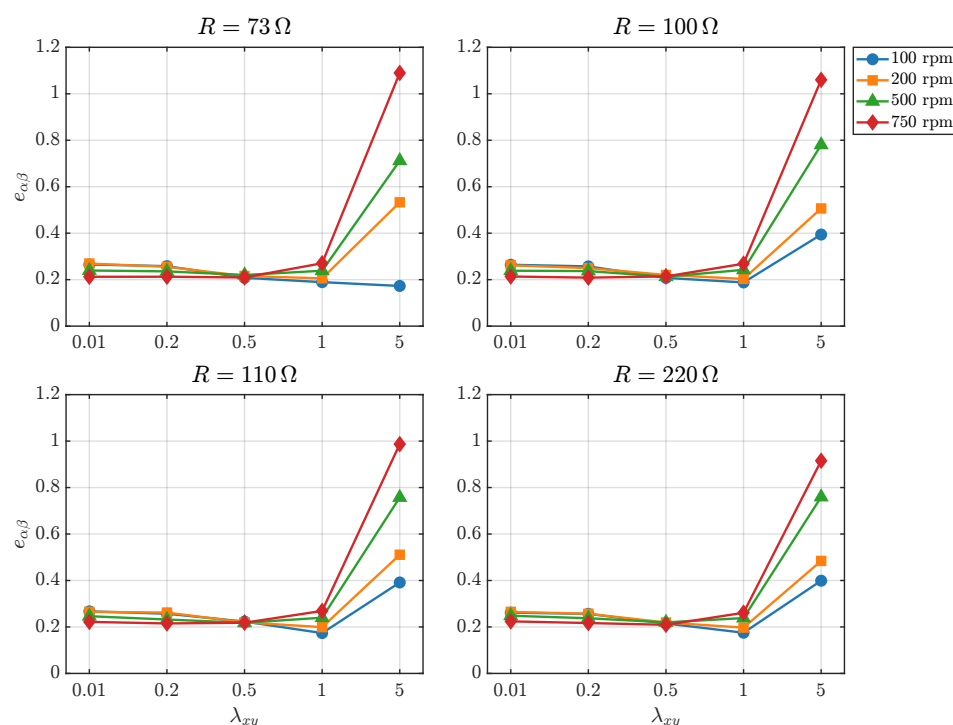


Figure 6. Influence of the control tuning parameter λ_{xy} on the tracking error in the α - β plane.

The influence of λ_{sc} on the switching behavior of the converter is illustrated in Figure 7. Since this WF penalizes switching transitions on the converter, increasing λ_{sc} is expected to reduce the average switching frequency. The experimental results confirm this trend, showing that f_{sw} decreases as λ_{sc} increases for all the operating conditions considered. This result demonstrates that the switching penalty effectively modifies the voltage vector selection process, favoring switching states that require fewer changes with respect to the previous converter state.

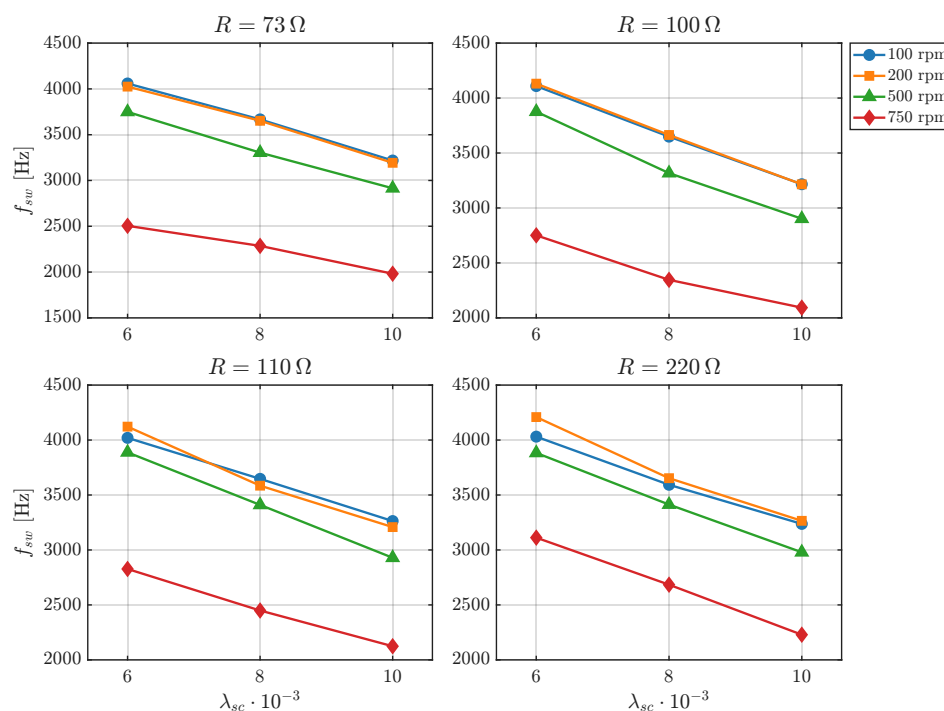


Figure 7. Influence of the control tuning parameter λ_{sc} on the switching frequency.

Having characterized the individual effect of each WF, the analysis now turns to the overall system efficiency. Figure 8 shows the average system efficiency as a function of the load resistance for different mechanical speeds. The results reveal a clear dependence of efficiency on the operating point. For all load conditions, efficiency increases with mechanical speed, mainly because the generator delivers higher output power at greater speeds. As useful output power rises, fixed losses (including mechanical friction, iron losses and converter dissipation) represent a smaller fraction of the total, leading to higher efficiency values. The highest efficiency recorded in the campaign reaches approximately 36% at 750 rpm under the lowest load resistance.

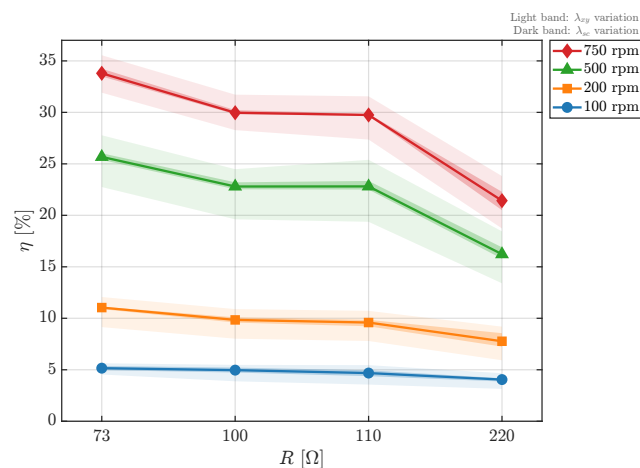


Figure 8. System efficiency as a function of load resistance for different mechanical speeds. Shaded bands indicate the efficiency range due to variations in λ_{xy} (light) and λ_{sc} (dark).

Moreover, for a given speed, the efficiency decreases as load resistance increases. This trend is especially noticeable at medium and high speeds. At low speed, however, efficiency remains below 6% and is nearly insensitive to load variations, confirming that the energy balance under these conditions is governed by speed dependent losses that the load level cannot offset. At higher speeds, the useful power transferred to the load becomes more significant, making efficiency increasingly sensitive to the load condition.

Beyond the operating point, the controller tuning also affects system efficiency. The shaded bands in Figure 8 represent the range of efficiency values obtained when λ_{xy} and λ_{sc} are varied for each operating condition; the bandwidth thus reflects the sensitivity of efficiency to controller tuning. The light band corresponds to variations in λ_{xy} and the dark band to variations in λ_{sc} . The wider light band indicates that efficiency is more sensitive to λ_{xy} than to λ_{sc} , reaching variations of up to 7.6 percentage points at 500 rpm.

At low speed, the bands remain narrow across all load values, confirming that the WFs have a limited effect when fixed losses dominate the energy balance. As speed increases, however, the bands become wider, particularly at medium and high load conditions. This indicates that controller tuning has a more noticeable impact on efficiency when the system operates at higher power levels, where the controllable loss components, especially those associated with the x - y currents, represent a larger share of the total losses.

The effect of λ_{xy} on efficiency is examined in detail in Figure 9. The efficiency exhibits a non-monotonic behavior with respect to this WF. The lowest values are obtained at both extremes of the tested range ($\lambda_{xy} = 0.01$ and $\lambda_{xy} = 5$), whereas intermediate values, particularly in the range 0.5–1, yield the highest efficiency. This result reflects two competing mechanisms. For very low values, the controller does not sufficiently suppress the current components in the x - y subspace; since these components do not contribute to torque production but circulate through the stator windings, they increase copper losses with-

out producing useful work. For excessively high values, the controller over prioritizes the secondary subspace at the expense of current tracking in the fundamental α - β plane, degrading overall energy conversion.

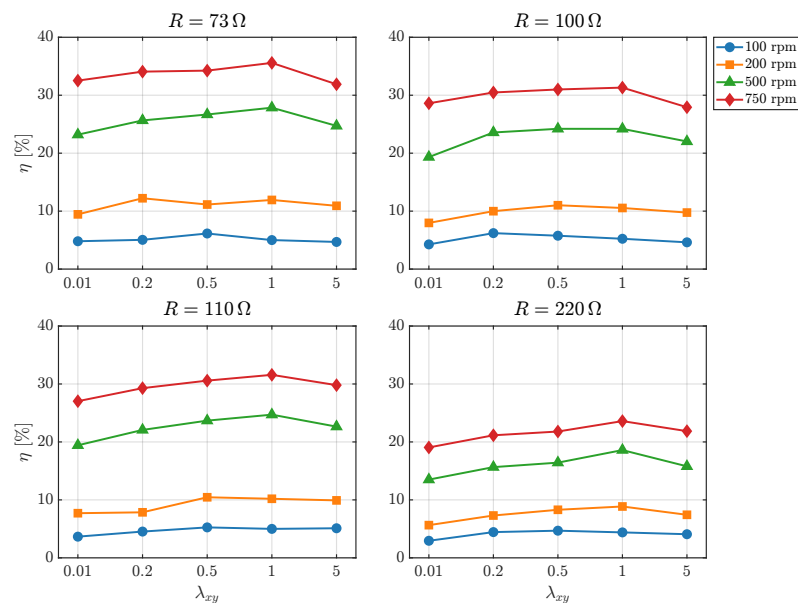


Figure 9. Effect of the control tuning parameter λ_{xy} on the efficiency.

In contrast, the effect of λ_{sc} on efficiency is much less pronounced, as shown in Figure 10. Although increasing λ_{sc} clearly reduces the switching frequency, this reduction does not translate into a significant change in overall efficiency. The efficiency curves remain nearly flat across the tested values of λ_{sc} , with only minor variations. This can be attributed to the fact that the switching penalty modifies the voltage vector selection process enough to reduce f_{sw} , but not enough to alter the current tracking behavior significantly. Consequently, the associated reduction in switching losses is too small to produce a measurable effect on the global energy balance.

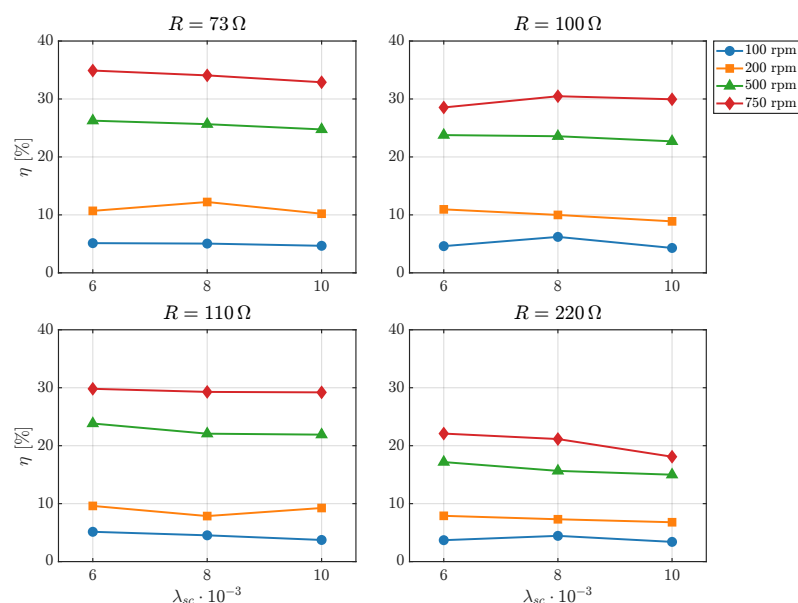


Figure 10. Effect of the control tuning parameter λ_{sc} on efficiency.

Table 2 summarizes the optimal WF combination and the corresponding efficiency range for each operating point. The optimal value of λ_{xy} falls consistently within the

range 0.5–1, with a clear shift toward 1 at medium and high speeds. The optimal λ_{sc} is predominantly $6 \cdot 10^{-3}$, confirming its limited influence on efficiency. The last column reports the efficiency difference between the best and worst WF combinations, which quantifies the potential improvement achievable through proper tuning. This difference increases with speed, reaching up to 7.64 percentage points at 500 rpm, and remains above 4 percentage points at 750 rpm for all load conditions.

Table 2. Optimal WF and efficiency range for each operating point for each load.

R [Ω]	ω [rpm]	λ_{xy}	$\lambda_{sc} [\times 10^{-3}]$	$\eta_{\min}$ [%]	$\eta_{\max}$ [%]	$\Delta\eta$ [pp]
73	100	0.5	8	4.30	6.14	1.84
	200	0.2	8	8.84	12.22	3.39
	500	1.0	6	22.46	28.39	5.92
	750	0.5	6	31.66	36.30	4.64
100	100	0.2	8	3.66	6.20	2.54
	200	0.5	8	7.75	11.02	3.26
	500	1.0	6	19.10	24.65	5.55
	750	1.0	6	27.14	32.47	5.33
110	100	1.0	6	3.20	6.08	2.88
	200	1.0	6	7.70	11.15	3.45
	500	1.0	6	18.60	26.25	7.64
	750	1.0	6	27.03	32.28	5.25
220	100	0.5	6	2.95	5.04	2.09
	200	0.5	6	5.64	11.32	5.67
	500	1.0	10	12.69	18.69	6.00
	750	1.0	6	17.80	24.26	6.46

Finally, to validate the dynamic performance of the optimal WF selection for the case of $\omega = 500$ rpm, with $\lambda_{xy} = 1$ and $\lambda_{sc} = 6 \times 10^{-3}$, a speed reference step test was carried out. Figure 11 shows the measured mechanical speed response, ω_m , to a step change in the speed reference, ω_m^* , from 0 to 500 rpm. It can be observed that the measured speed accurately tracks the reference, reaching the new operating point without noticeable overshoot.

Figure 12 depicts the corresponding stator current components in the synchronous reference frame, i_{ds} and i_{qs} , together with their respective references, i_{ds}^* and i_{qs}^* . During the transient, i_{qs} rapidly increases to its maximum value to generate the required torque to achieve the desired speed step, whereas i_{ds} remains regulated around its reference value, ensuring constant flux production throughout the maneuver. As the mechanical speed converges to the reference value, i_{qs} gradually decreases to the level required to sustain the new steady-state operating point. These results confirm the ability of the proposed control scheme to achieve accurate speed reference tracking using the optimal WF.

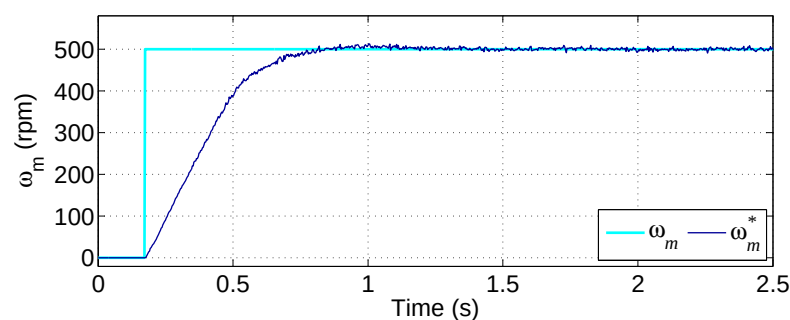


Figure 11. Mechanical speed response to a step change in the speed reference using the optimal WF ($\lambda_{xy} = 1$, $\lambda_{sc} = 6 \times 10^{-3}$) for $R = 100 \Omega$.

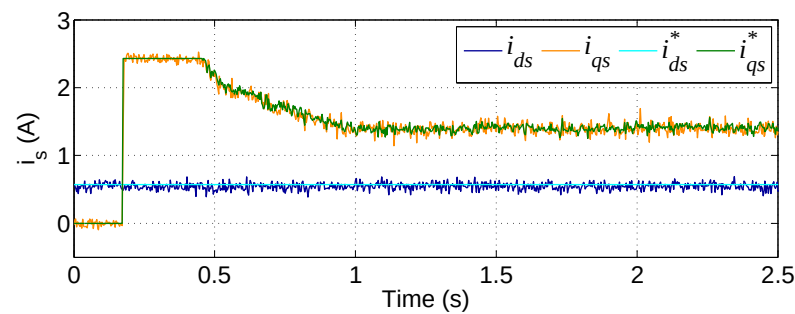


Figure 12. Stator current response (i_{ds} , i_{qs}) to a step change in the speed reference using the optimal WF ($\lambda_{xy} = 1$, $\lambda_{sc} = 6 \times 10^{-3}$) for $R = 100 \, \Omega$.

6. Discussion

The experimental results show that drive efficiency is governed by two distinct factors, namely the operating point, defined by mechanical speed and load resistance, and controller tuning through the WF λ_{xy} and λ_{sc} . Their relative importance, however, differs considerably. The operating point has the dominant influence, while the controller tuning plays a secondary but non-negligible role that becomes increasingly relevant at medium and high power levels.

The dominant effect of speed can be understood from the energy balance of the motor–generator set. At low speed, the output power is small and fixed losses (mechanical friction, iron losses and converter losses) consume a large fraction of the input power, keeping efficiency below 6% regardless of load or controller settings. As speed increases, the generator delivers substantially more power, reducing the relative weight of these losses and improving efficiency. Similarly, lower load resistance allows higher output currents and therefore higher useful power, which explains the consistent efficiency improvement observed at lower resistance values, particularly at 500 and 750 rpm.

The influence of λ_{xy} on efficiency can be explained by a specific physical mechanism. The current components in the x - y subspace do not contribute to torque production in a five-phase machine; however, they circulate through the stator windings and produce additional copper losses. When λ_{xy} is too low, these parasitic currents are insufficiently suppressed, resulting in increased losses and reduced efficiency. As λ_{xy} increases, the x - y currents are progressively attenuated, leading to improved efficiency. Beyond a certain threshold, however, further increases provide limited additional reduction of e_{xy} while introducing two adverse effects: the degradation of the fundamental α - β current tracking and an increase in switching frequency, which raises inverter losses.

The degradation of the fundamental plane tracking is confirmed by the results shown in Figure 6 and Table 2, where $e_{\alpha\beta}$ remains stable for λ_{xy} between 0.01 and 1 but increases sharply at $\lambda_{xy} = 5$, reaching values up to four times higher at 750 rpm. This degradation is mainly determined by mechanical speed and shows limited sensitivity to load resistance, indicating that higher speeds intensify the trade-off between suppressing secondary subspace currents and maintaining accurate regulation of the torque-producing components. Consequently, excessive emphasis on secondary subspace control eventually compromises the overall current regulation performance. The interaction between these mechanisms results in the non-monotonic efficiency profile observed in Figure 9, with an optimal region around $\lambda_{xy} = 0.5$ – 1 .

The behavior of λ_{sc} is different. This factor directly penalizes switching transitions and, as expected, increasing λ_{sc} reduces the average switching frequency. While this reduction is potentially beneficial in terms of semiconductor stress, electromagnetic emissions, and thermal management, its impact on efficiency within the tested range is marginal. This can be explained by the relatively low magnitude of the switching losses compared to the

copper and iron losses that dominate the energy balance of the system. As a result, λ_{sc} can be regarded as a parameter primarily aimed at improving the switching behavior of the converter rather than optimizing efficiency.

These findings have a direct practical implication for controller tuning. The analysis of the optimal WF summarized in Table 2 reveals that λ_{xy} concentrates in the range 0.5–1 across all tested conditions, with a shift toward $\lambda_{xy} = 1$ at medium and high speeds, while λ_{sc} is predominantly 6×10^{-3} . Rather than treating WF selection as a purely control-oriented task focused on current tracking or switching frequency, the results demonstrate that the global energy performance of the drive should also be considered. In particular, the choice of λ_{xy} should not be based solely on minimizing e_{xy} , since the value that minimizes the secondary plane error does not coincide with the one that maximizes efficiency.

7. Conclusions

This work presented an experimental analysis of the influence of the operating point and the cost function WFs on the efficiency of a five-phase IM drive controlled by FCSMPC. The system efficiency is primarily determined by the operating point. Mechanical speed has the dominant effect, with higher speeds increasing the generator output power and reducing the relative contribution of fixed losses, raising efficiency from below 6% at 100 rpm to approximately 36% at 750 rpm. Lower load resistance values further improve efficiency by increasing the useful output power, with this effect being most pronounced at medium and high speeds.

Among the controller parameters, λ_{xy} exhibits the strongest impact on efficiency. Its effect follows a non-monotonic behavior explained by two competing mechanisms, namely the insufficient suppression of parasitic x - y currents at low values and degradation of fundamental plane tracking at excessively high values. Experimental results show that $e_{\alpha\beta}$ increases significantly when $\lambda_{xy} = 5$, particularly at high speeds, while remaining stable in the range 0.01–1. The highest efficiency was obtained for $\lambda_{xy} = 0.5$ –1, where a suitable balance between secondary current suppression and fundamental tracking is achieved. Notably, this range does not coincide with the value that minimizes e_{xy} , highlighting the need to consider energy efficiency, rather than tracking performance alone, during the tuning process.

The WF λ_{sc} effectively reduces the switching frequency, which may benefit converter lifetime and electromagnetic compatibility, but its impact on overall efficiency is limited within the tested range. This result suggests that λ_{sc} can be adjusted primarily based on switching related criteria without significantly affecting energy performance.

Overall, the results demonstrate that WF tuning should not be treated as a purely control-oriented task. Intermediate values of λ_{xy} combined with moderate values of λ_{sc} provide a practical compromise between current tracking, switching behavior and energy efficiency, improving the overall system performance across the tested operating conditions. Future work will investigate the interaction effects between the two weighting factors by systematically analyzing their combined variation. Moreover, additional values in the vicinity of the optimal point identified in this study will be explored to refine the optimization process and better understand its local behavior.

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Abbreviations

The following abbreviations are used in this manuscript:

CCSMPC	Continuous Control Set Model Predictive Control
DC	Direct Current
FCSMPC	Finite Control Set Model Predictive Control
FSMPC	Finite State Model Predictive Control
IM	Induction Motor
MPC	Model Predictive Control
PWM	Pulse Width Modulation
VSI	Voltage Source Inverter
WFs	Weighting Factors

References

- Holtz, J. A predictive controller for the stator current vector of ac machines fed from a switched voltage source. *IPEC-Tokyo* **1983**, 1983, 1665–1675.
- Linder, A.; Kennel, R. Model predictive control for electrical drives. In *Proceedings of the 2005 IEEE 36th Power Electronics Specialists Conference*; IEEE: Dresden, Germany, 2005; pp. 1793–1799.
- Shahid, M.B.; Jin, W.; Abbasi, M.A.; Husain, A.R.B.; Munir, H.M.; Hassan, M.; Flah, A.; Souissi, A.S.E.; Alghamdi, T.A. Model predictive control for energy efficient AC motor drives: An overview. *IET Electr. Power Appl.* **2024**, *18*, 1894–1920. [\[CrossRef\]](#)
- Vijayan, A.K.; da Costa Gonçalves, P.F.; Batkhishig, B.; Nahid-Mobarakeh, B.; Emadi, A. Integrated FCS-MPC with Synchronous Optimal Pulse-Width Modulation for Enhanced Dynamic Performance in Two-Level Voltage-Source Inverters. *Electronics* **2025**, *14*, 3757. [\[CrossRef\]](#)
- Arahal, M.R.; Barrero, F.; Ortega, M.G.; Martin, C. Harmonic analysis of direct digital control of voltage inverters. *Math. Comput. Simul.* **2016**, *130*, 155–166. [\[CrossRef\]](#)
- Zhou, Z.; Su, M. Efficient Predictive Control for Dual Three-phase PMSM Drive with Vector Transformation. *IEEE Trans. Power Electron.* **2024**, *39*, 13630–13642. [\[CrossRef\]](#)
- Yu, B.; Song, W.; Yang, K.; Guo, Y.; Saeed, M.S. A computationally efficient finite control set model predictive control for multiphase PMSM drives. *IEEE Trans. Ind. Electron.* **2021**, *69*, 12066–12076.
- Song, X.; Wang, H.; Ma, X.; Yuan, X.; Wu, X. Robust Model Predictive Current Control for a Nine-Phase Open-End Winding PMSM With High Computational Efficiency. *IEEE Trans. Power Electron.* **2023**, *38*, 13933–13943. [\[CrossRef\]](#)
- Arahal, M.R.; Barrero, F.; Satué, M.G.; Bermúdez, M. Fast finite-state predictive current control of electric drives. *IEEE Access* **2023**, *11*, 12821–12828. [\[CrossRef\]](#)
- Jayan, V.; Ghias, A.M. Weighting Factor Free Model Predictive Control for a Flying Capacitor Converter in a DC Microgrid. *IEEE Trans. Energy Convers.* **2021**, *37*, 1030–1041.
- Serra, J.; Cardoso, A.J.M. A simplified model predictive control for asymmetrical six-phase induction motors that eliminates the weighting factor. *Machines* **2022**, *10*, 1189. [\[CrossRef\]](#)
- Liu, X.; Qiu, L.; Wu, W.; Ma, J.; Fang, Y.; Peng, Z.; Wang, D. Neural predictor-based low switching frequency FCS-MPC for MMC with online weighting factors tuning. *IEEE Trans. Power Electron.* **2021**, *37*, 4065–4079.
- Deng, A.; Yang, W.; Hu, G.; Huang, W.; Xu, D. Reinforcement Learning based Weight-Tuning Model Predictive Control of Permanent Magnet Synchronous Motor. In *Proceedings of the 2024 IEEE 10th International Power Electronics and Motion Control Conference (IPEMC2024-ECCE Asia)*; IEEE: New York, NY, USA, 2024; pp. 317–322.
- Arahal, M.R.; Satué, M.G.; Barrero, F.; Ortega, M.G. Adaptive cost function FCSMPC for 6-phase IMs. *Energies* **2021**, *14*, 5222. [\[CrossRef\]](#)
- Savi, F.; Cutuli, G.; Nuzzo, S.; Barater, D. System Level, Comparative Loss Evaluation of Interior Permanent Magnet Drive with Traditional and Predictive Control Strategies. In *Proceedings of the 2025 IEEE Workshop on Electrical Machines Design, Control and Diagnosis (WEMDCD)*; IEEE: New York, NY, USA, 2025; pp. 1–6.
- Karimi, S.; Seifinejad, H.; Moradi, H.; Bagheri, H.; Hemmati, S. Constrained FCS-MPC-Based Current Controller for Voltage Source Converters: Application to Shunt Active Power Filters. *IET Electr. Power Appl.* **2025**, *19*, e70025. [\[CrossRef\]](#)

17. Yao, C.; Xu, S.; Sun, Z.; Ren, G.; Li, G.; Ma, G. Automatically Tuning of Weighting Factors for FCS-MPC in PMSM Drives Using Lightweight Neural Networkweight Neural Network. In *Proceedings of the 2023 IEEE International Conference on Predictive Control of Electrical Drives and Power Electronics (PRECEDE)*; IEEE: New York, NY, USA, 2023; pp. 1–6.
18. Satué, M.G.; Barrero, F.; Martínez-Heredia, J.M.; Colodro, F. Concurrent AI tuning of a double-loop controller for multi-phase drives. *Machines* **2024**, *12*, 899. [[CrossRef](#)]
19. An, X.; Liu, Z.; Chen, Q.; Liu, G. Robust virtual-vector model predictive control of permanent-magnet motor considering D-Q axis inductance parameter uncertainty. *IET Electr. Power Appl.* **2024**, *18*, 76–89. [[CrossRef](#)]
20. Lee, S.S.; Heng, Y.E. Current controller of three-phase VSI using FCS-MPC with optimal active and zero vector selection for balanced loss distribution and switching loss reduction. *Int. Trans. Electr. Energy Syst.* **2017**, *27*, e2250. [[CrossRef](#)]
21. Xue, C.; Song, W.; Feng, X. Finite control-set model predictive current control of five-phase permanent-magnet synchronous machine based on virtual voltage vectors. *IET Electr. Power Appl.* **2017**, *11*, 836–846. [[CrossRef](#)]
22. Peng, T.; Huang, X.; Yang, C.; Chen, Z.; Liu, Q.; Fan, X. Temperature control-based FCS-MPC of traction converters. In *Proceedings of the 2021 CAA Symposium on Fault Detection, Supervision, and Safety for Technical Processes (SAFEPROCESS)*; IEEE: New York, NY, USA, 2021; pp. 1–6.
23. Barotto, B.K.; Brunoni, J.P.; Flesch, R.C.C. Impact of the FCS-MPC Switching Reduction Weighting Coefficient on Losses and THD for a Three-Phase Inverter. In *Proceedings of the 2025 5th International Conference on Electrical, Computer, Communications and Mechatronics Engineering (ICECCME)*; IEEE: New York, NY, USA, 2025; pp. 1–6.
24. Marsal, E.; G Satué, M.; Colodro, F. Estudio energético de FSMPC para un motor de 5 fases. In *Proceedings of the Simposios del Comité Español de Automática (CEA)*; Comité Español de Automática (CEA): Barcelona, Spain, 2025.
25. Marsal, E.; Arahal, M.R.; Satué, M.G.; Rouzbehi, K. Effect of Weighting Factors in Energy Efficiency of Predictive Control of Multi-Phase Drives. *Appl. Sci.* **2025**, *15*, 12148. [[CrossRef](#)]
26. Rodas, J.; Barrero, F.; Arahal, M.R.; Martín, C.; Gregor, R. Online Estimation of Rotor Variables in Predictive Current Controllers: A Case Study Using Five-Phase Induction Machines. *IEEE Trans. Ind. Electron.* **2016**, *63*, 5348–5356. [[CrossRef](#)]
27. Martín, C.; Arahal, M.R.; Barrero, F.; Durán, M.J. Multiphase rotor current observers for current predictive control: A five-phase case study. *Control Eng. Pract.* **2016**, *49*, 101–111. [[CrossRef](#)]
28. Satué, M.G.; Arahal, M.R.; Ortega, M.G.; Ramírez, D.R. A simple rotor current estimation method for predictive control of multi-phase drives. *Int. J. Circuit Theory Appl.* **2022**, *50*, 4478–4491. [[CrossRef](#)]

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