Duty Cycle Optimization for Constant Switching Frequency in 3-Phase, 2-Level PWM Rectifiers, Using Modified Modulated Predictive Control

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Abstract: In order to improve the steady-state performance of the finite control set model predictive control (FCS-MPC), this paper proposes a simplified modulated predictive current control scheme for a three-phase two-level PWM rectifier. In the proposed scheme, the optimal vector is evaluated initially similarly to the conventional finite control set model predictive control (FCS-MPC). Later, an additional stage is included to compute the duty cycle of the optimal voltage vector obtained from stage 1 by using the simplified relationship between the cost-function values and the duty cycle. Compared to the prior art the proposed approach can effectively improve the THD of the input current with less complexity. The proposed approach retains the fast dynamic response of the FCS-MPC with an improved steady-state performance. In addition, the fixed switching frequency operation is also obtained.

Keywords: PWM rectifier; modulated MPC; THD; fixed switching frequency; Three-Phase Rectifier; Load Change Transient Response

1 Introduction

In the landscape of modern power electronics, three-phase 2-level (2-L) PWM rectifiers have taken a prominent role due to their extensive applicability across various industrial and energy sectors. These converters stand out for their efficiency and versatility, enabling enhanced power quality management, unity power factor operation, and the capability for bi-directional power flow. Such attributes are crucial in various applications [1], from renewable energy integration to advanced manufacturing systems [2], underscoring their importance in current and future power systems infrastructure [3].

Amidst the diversity of control strategies formulated to optimize the performance of these rectifiers, Finite Control Set Model Predictive Control (FCS-MPC) has emerged as a particularly influential approach. Celebrated for its rapid response to dynamic system changes, the ability to incorporate nonlinear objectives directly into the control strategy, and relative ease of implementation, FCS-MPC has been highlighted in [4] for its incorporation of nonlinear objectives and implementation simplicity, while the demonstration of its effectiveness in power converter control can be found [5]. Its application has been broad, covering various configurations of PWM rectifiers, each demonstrating the approach's efficacy in real-world scenarios [6].

The operational principle of FCS-MPC is grounded in the use of a cost-function that evaluates potential voltage vectors, selecting the optimal one for application in the immediate next sampling interval. This process, while effective, introduces challenges, particularly regarding the handling of real-time implementation delays. For instance, the delay compensation mechanisms to enhance stability in grid-connected converters has been explored in [7]. Such delays necessitate sophisticated delay compensation mechanisms to ensure system stability and performance[8], complicating the control scheme further [9]. Those distinct approaches to mitigate real-time implementation delays also found in [10].

The literature reveals various attempts to refine predictive control strategies for power converters. Notably, deadbeat predictive control has been adapted for grid-connected converters to streamline the prediction equation and improve the selection process for the optimal voltage vector [11]. Similarly, efforts to address grid voltage distortions through predictive control have been documented, demonstrating the versatility and adaptability of predictive control frameworks [12]. Yet, despite these advancements, FCS-MPC's application has encountered limitations, notably its variable frequency operation and less than optimal steady-state performance compared to traditional linear controllers, such as PI control with Space Vector Modulation (SVM). The pursuit of enhanced controller performance has historically led to the adoption of higher sampling frequencies or the use of larger input filter inductances—solutions that are not without cost implications [3].

In response to these challenges, the academic and industrial communities have proposed various alternative strategies. The description of the integration of discrete space vector modulation (DSVM) with FCS-MPC to enhance steady-state performance depicted in [13]. Similarly, the alternative DSVM approaches, while also discussing the resulting increase in system complexity [14]. Modulated MPC emerged as another solution, promising improved steady-state performance and fixed switching frequency operation, albeit with higher computational demands [15]. Continuous Control Set MPC (CCS-MPC) also offers performance enhancements but is hindered by its requirement for offline implementation and the complexity of its tuning and problem formulation processes [16].

In light of these considerations, this paper presents a modified predictive current control scheme for PWM rectifiers that fundamentally simplifies the computational landscape of FCS-MPC. By recalibrating the approach to calculate the duty cycle of the active voltage vector through a direct mathematical relationship with the cost-function values of the optimal voltage vector, this approach retains the quick dynamic responsiveness intrinsic to FCS-MPC while substantially improving steady-state performance regarding input current Total Harmonic Distortion (THD). A cornerstone of this approach is its achievement of fixed switching frequency operation - a notable leap over traditional FCS-MPC implementations - accomplished without imposing additional computational burdens. This attribute of fixed switching frequency operation is not merely a technical enhancement; it streamlines the design of filtering components, augments system predictability, and assures compatibility with prevailing power electronics standards, facilitating a smoother integration into existing control architectures.

Moreover, by guaranteeing compatibility with standard control platforms, the proposed approach substantially broadens its applicability and adoption potential, offering a refined solution that seamlessly integrates into a diverse array of application contexts without necessitating specialized adjustments to hardware or software configurations. The validation of this approach through rigorous simulation studies within the Matlab-Simulink environment underscores its efficacy in diminishing the THD of the input current while steadfastly maintaining a fixed-switching frequency operation, thereby setting a new benchmark for control strategies in the realm of PWM rectifiers and potentially transforming the landscape of power electronics control.

The proposed control scheme targets three major objectives:

(1) Accurate tracking of the grid current reference to achieve unity power factor

(2) Minimizing the input current total harmonic distortion (THD)

(3) Guaranteeing constant switching frequency operation

The underpinning optimization problem is to find, at each sampling instant, the duty cycle(s) for the candidate voltage vectors that minimize a quadratic current tracking error (between reference and predicted grid current) plus a penalty term for THD, all subject to system dynamic constraints and feasible duty cycle ranges.

The paper is organized as follows: Section 2 presents the system model. Section 3 describes the proposed MMPC-based predictive control strategy. Section 4 details the optimization algorithm. Section 5 discusses simulation results. Section 6 concludes and outlines future work.

2 System Modeling

The controller is designed based on a discrete-time state-space model, obtained via Euler forward discretization of the rectifier dynamics (see Eqns. 4-8).

For the predictive control strategy, the model of the system is required in discrete form.

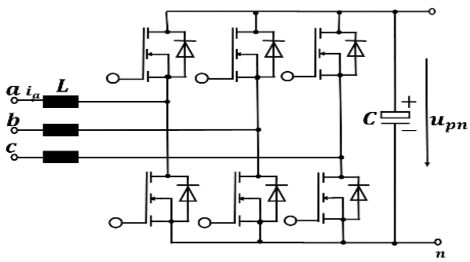

Figure 1
PWM Rectifier Topology

The control objective is defined to meet the input current tracking to its reference value with unity power factor operation. The converter topology is shown in Fig. 1. The dynamics of the input current is required to predict the future behavior of the input current. MPC technique then evaluates the cost-function value based on this model and selects the optimal converter switching state which provides the minimum cost-function value [17].

$$\bar{v} = R\bar{i} + L\frac{d\bar{i}}{dt} + \bar{v}_g \tag{1}$$

Where the complex space vectors are calculated from the instantaneous phase quantities as:

$$\bar{v} = \frac{2}{3}\left(v_{an} + v_{bn}\bar{a} + v_{cn}\bar{a}^2\right), \quad \bar{i} = \frac{2}{3}\left(i_a + i_b\bar{a} + i_c\bar{a}^2\right) \tag{2}$$

$$\bar{v}_g = \frac{2}{3}\left(v_{ga} + v_{gb}\bar{a} + v_{gc}\bar{a}^2\right) \quad \text{and} \quad \bar{a} = e^{j2\pi/3} \tag{3}$$

Here, $\bar{v}_g$ is the grid voltage, where $\bar{v}$ is the converter voltage vector and $\bar{i}$ is the input current of the converter. As mentioned earlier the model of the system needs to be discretized as the control implementation is referred to digital implementations. Thus, for discretization, Euler's backward discretization approach is chosen. The current derivative can be expressed as:

$$\frac{d\bar{i}}{dt} \approx \frac{\bar{i}(k) - \bar{i}(k-1)}{T_s} \tag{4}$$

It is assumed that the output voltage of the inverter and grid voltage are constant during a small sampling time T_s, thus the current slope is constant. The grid current is obtained by replacing (4) in (1):

$$\bar{i}(k) = \frac{1}{RT_s + L}\left[L\bar{i}(k-1) + T_s\bar{v}(k) - T_s\bar{v}_g(k)\right] \tag{5}$$

The future grid current is calculated by shifting the discrete time one step forward in (5).

$$\bar{i}(k+1) = \frac{1}{RT_s + L}[L\bar{i}(k) + T_s\bar{v}(k+1) - T_s\bar{v}_g(k+1)] \tag{6}$$

or for the separate $\alpha\beta$ components the instantaneous current behavior under static stationary coordinates can be derived as:

$$i_\alpha(k+1) = \frac{1}{RT_s + L}[Li_\alpha(k) + T_s v_\alpha(k+1) - T_s v_{g\alpha}(k+1)] \tag{7}$$

$$i_\beta(k+1) = \frac{1}{RT_s + L}[Li_\beta(k) + T_s v_\beta(k+1) - T_s v_{g\beta}(k+1)] \tag{8}$$

3 Proposed Approach

The voltage vectors for the converter under study are shown in Fig. 2 along with the six sectors. The conventional FCS-MPC computes the following cost function for all the converter voltage vectors for the power converter and selects the optimal switching state based on the cost-function minimization.

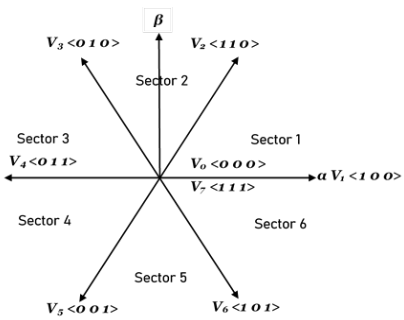

Figure 2

Space vector diagram for the two-level VSI

The cost function is defined as the square of the error current magnitude at the second prediction horizon:

$$g = |\bar{i}^*(k+1) - \bar{i}(k+1)|^2 = [i_\alpha^*(k+1) - i_\alpha(k+1)]^2 + [i_\beta^*(k+1) - i_\beta(k+1)]^2 \quad (9)$$

where $\bar{i}^*$ is the reference current. The reference current is provided by the outer loop linear PI DC link voltage controller. The converter switching state which gives the minimal value of the cost function is selected as the optimal switching state and applied as actuators. In[18] demonstrated that improved vector selection in model predictive torque control can significantly enhance system performance, a concept that aligns with the optimization process adopted in this work. However, the FCS-MPC does not provide a constant switching frequency operation and also the performance of the FCS-MPC is not comparable with other linear controllers such as Direct Power Control with SVM (DPC-SVM). Thus, it is desirable to develop a control technique that can retain some key advantages of the FCS-MPC and also provides improved steady-state performance with constant switching frequency. Thus, in this paper, a modified MPC technique is proposed which can improve the performance of the FCS-MPC and operate at a fixed-switching frequency. With the proposed approach one active vector and one zero vector will be applied instead of a single vector application. The aim is to obtain a fixed-switching frequency operation with improved performance.

The duty cycle for the active vector to be applied is d_{opt} and the duty cycle for the zero-vector is d_0. Thus, the following expressions hold true:

$$d_{opt} + d_0 = 1 \quad (10)$$

In this formulation, the primary optimization variables are the duty cycles assigned to the active vector d_{opt} and the zero vector d_0 , as they directly determine the vector application durations within each sampling period.

If the error associated with each voltage vector related to the cost function is denoted as ε then for the application of a duty cycle d_{opt} the weighted average error can be written as $d_{opt}\varepsilon$. The relationship between the cost-function values with the duty cycle is obtained using the minimization of this error as reported in [19]. Considering the Karush–Kuhn–Tucker condition and the Lagrange multiplier approach for the optimization problem an approximate relation between the duty cycle and cost-function can be obtained. The duty cycle is found to maintain a reciprocal relationship with the cost function value. Thus,

$$d_0 \propto \frac{1}{g_0}, \quad d_{opt} \propto \frac{1}{g_{opt}} \quad (11)$$

This can be written as,

$$d_0 = \frac{\kappa}{g_0}, \quad d_{opt} = \frac{\kappa}{g_{opt}} \tag{12}$$

Now,

$$\frac{\kappa}{g_0} + \frac{\kappa}{g_{opt}} = 1 \tag{13}$$

$$\Rightarrow \quad \kappa = \frac{g_{opt} \cdot g_0}{g_{opt} + g_0} \tag{14}$$

Thus,

$$d_{opt} = \frac{g_{opt} \cdot g_0}{(g_{opt} + g_0) \cdot g_{opt}} = \frac{g_0}{g_{opt} + g_0} \tag{15}$$

This allows computing the optimal duty cycle for the optimal active voltage vector d_0 can be computed as:

$$d_0 = 1 - d_{opt} \tag{16}$$

The overall process is depicted with a flow diagram shown in Fig. 3.

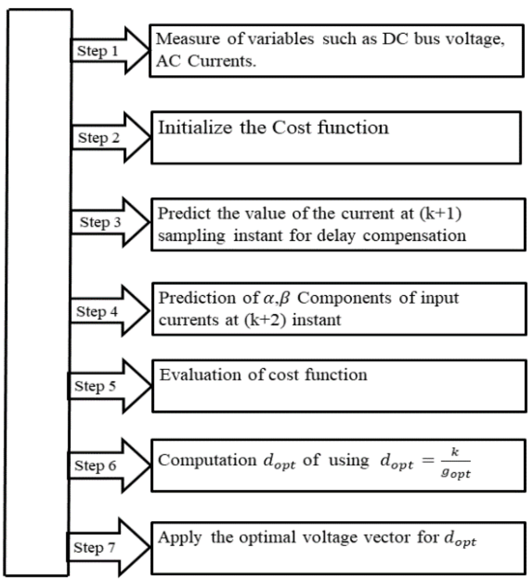

Figure 3

Flow diagram showing the implementation process of the proposed approach

4 Optimization Approach in Modified Modulated Predictive Control (MMPC)

System stability in the proposed MMPC framework is inherently promoted by penalizing large current errors in the cost function and by limiting duty cycle selections to physically admissible values ($0 < d_{opt}$, $d_0 < Ts$, $d_{opt} + d_0 = T_s$). These constraints prevent infeasible or unstable switching actions. Future studies may incorporate explicit Lyapunov-based constraints or robust margins if formal guarantees are needed under more challenging operating conditions [20].

The MMPC algorithm strives to optimize the converter performance by determining, at each control cycle, the most effective application intervals (duty cycles) for both the optimal active voltage vector and a corresponding zero vector. This dual-vector approach enables constant switching frequency operation, unlike conventional FCS-MPC [21]. Similarly [22], proposed a fixed switching frequency strategy for finite-control-set MPC by reconstructing the cost function, which provides a theoretical basis for the approach adopted here.

The algorithm proceeds as follows at each sampling instant:

1. Prediction

The discrete-time state-space model of the PWM rectifier, as detailed in Section 2 (see Eqns. 4–8), is used to predict future grid currents for all candidate switching vectors under evaluation.

2. Cost Function Evaluation

For each candidate vector, the algorithm computes the cost function:

$$g = \left\| i_g^*[k+1] - i_g[k+1] \right\|^2 + \lambda \cdot \mathrm{THD}(i_g[k+1]) \tag{17}$$

where $i_g^*[k+1]$ is the reference current vector for the next step $i_g[k+1]$ is the predicted current and λ is a tunable non-negative weight for THD in the cost function. Typically, λ may be set to zero to focus on current tracking, or given a value to penalize harmonic distortion in applications where waveform quality is paramount.

3. Duty Cycle Calculation

Once the cost for each vector is evaluated, identify the optimal active voltage vector, denoted by the lowest cost value (g_{opt}), and a zero vector (g_0). The application durations (duty cycles) for these vectors are then calculated using the reciprocal cost-law relationship:

$$d_{opt} = T_s \cdot (g_0 / (g_{opt} + g_0))$$

$$d_0 = T_s - d_{opt} \tag{18}$$

This ensures longer application duration for vectors with smaller cost, while always meeting $d_{opt} + d_0 = T_s$

4. Switching Actuation

Apply the optimal active and zero vectors for their computed durations within the current sampling period. This process results in a fixed switching frequency, in contrast to the variable frequency typical of standard FCS-MPC.

5. Iterative Implementation and Feedback

The above process is repeated at every sampling interval. Real-time measurement and feedback render the scheme robust to system changes, including load fluctuations and grid disturbances.

Advantages: Compared to conventional FCS-MPC, this MMPC approach:

- Guarantees a fixed switching frequency
- Reduces input current THD (see Section 5 for quantitative results)
- Retains fast transient response
- Does not increase computational burden due to its simple, closed-form duty calculation

A block diagram summarizing this process is provided in Figure 3

5 Simulation Results

In order to provide support to the theoretical approach simulation study is performed in a Matlab-simulink environment. The parameters used in the simulation are listed in Table 1.

Table 1

Parameters for the simulation

Parameters	**Value**
Grid line-to-line voltage (rms)	415 V
Fundamental frequency	50 Hz
Filter inductance	8 mH
Filter parasitic resistance	0.1 Ω
DC-Bus voltage	600 V
DC capacitor	10 µF
Nominal load resistance	50 Ω
Control sampling time	40 µs

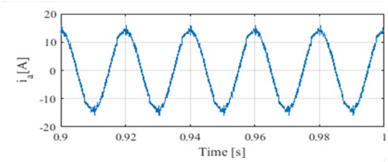

Figure 4: Input current of FCS-MPC with 50 Ω load.

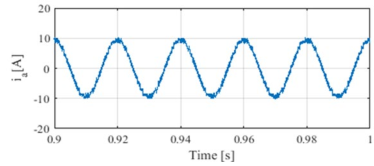

Figure 5: Input current of FCS-MPC with 75 Ω load.

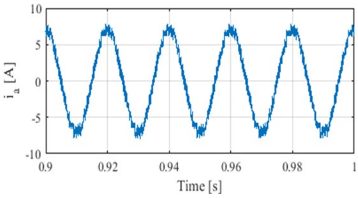

Figure 6: Input current of FCS-MPC with 100 Ω load.

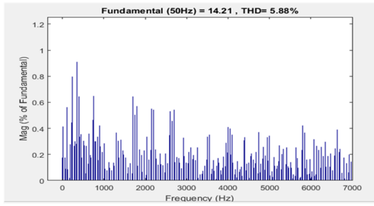

Figure 7: FFT spectrum of FCS-MPC with 50 Ω load.

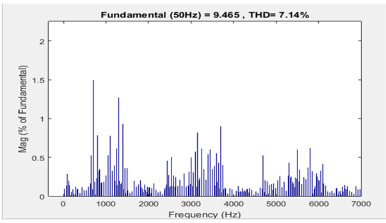

Figure 8: FFT spectrum of FCS-MPC with 75 Ω load.

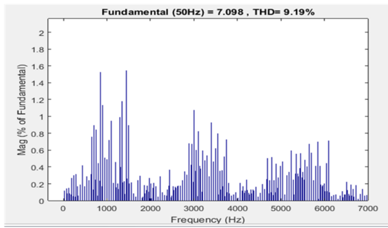

Figure 9: FFT spectrum of FCS-MPC with 100 Ω load.

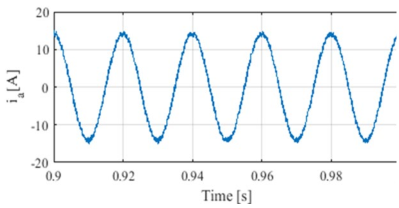

Figure 10: Input current of the proposed approach with 50 Ω load.

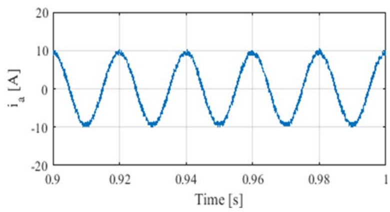

Figure 11: Input current of the proposed approach with 75 Ω load.

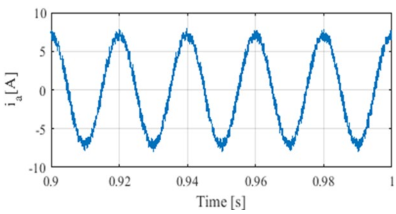

Figure 12: Input current of the proposed approach with 100 Ω load.

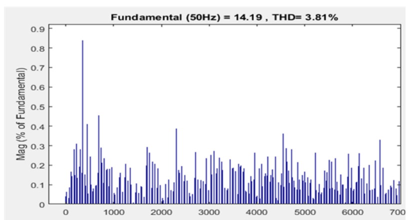

Figure 13: FFT spectrum of the proposed approach with 50 Ω load.

The proposed approach can significantly improve the Total Harmonic Distortion (THD) of the input current. Phase current time functions for different loading conditions are shown in Fig. 4-6 for FCS-MPC and in Fig. 7-9 for the proposed approach and the corresponding FFT spectra for various load can be seen in Fig. 10-12 for FCS-MPC and Fig. 13-15 for the proposed approach.

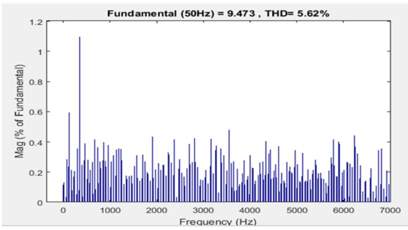

Figure 14: FFT spectrum of the proposed approach with 75 Ω load.

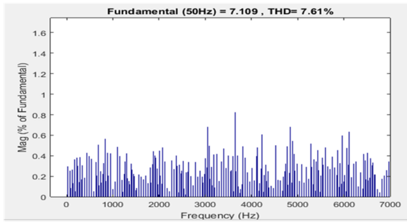

Figure 15: FFT spectrum of the proposed approach with 100 Ω load.

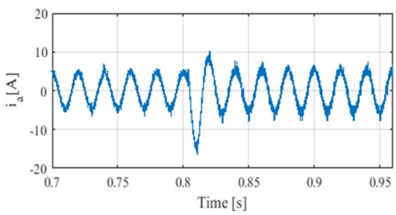

Figure 16: Transient response with the FCS-MPC.

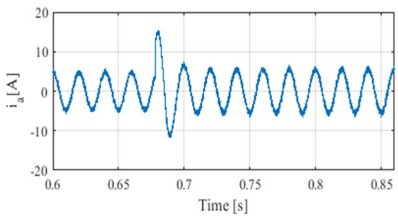

Figure 17: Transient response with the proposed approach.

The THD values and their relative change are summarized in Table 2. At the rated load the THD decreased from 5.88% to 3.81%. At light load, the THD decreased from 9.19% to 7.61%. It can be seen the relative change in THDs is larger at larger load, which can be observed also in the time functions. Looking at the spectra of the two approaches in Fig. 10-15, the FCS-MPC has some frequency ranges with higher peaks especially at lighter loads comparing to the proposed approach, where the spectra seems more evenly distributed.

It can be seen that the steady-state THD performance of the converter is improved with the proposed approach compared to the FCS-MPC. Moreover, the application of the duty cycle implies the fixed-switching frequency operation and solves one of the major problems of the FCS-MPC. However, It has to be noted that the THD alone is not a fair measure in the comparison of a constant switching frequency approach to a variable switching frequency one. The average number of state changes in the rectifier during a fundamental period for the FCS-MPC also need to be counted in further research work to decide with more confidence the superiority of the proposed approach.

The transient performance with the FCS-MPC and the proposed approach is shown in Fig. 16 and Fig. 17 for a load change from 75 Ω to 100 Ω, respectively. It can be seen that both approaches perform similarly, the transients disappear in a single fundamental period. Note, that the relatively large transients that the current peaks reach in both case approximately 15 A are not from the studied current controls but from the outer quite slow linear voltage control loop which provided the reference current.

Table 2
Comparison of input current THD of FCS-MPC and the proposed approach

Rload(Ω)	THD1 [%] FCS-MPC	THD2 [%] proposed	THD2/ THD1
50	5.88	3.81	0.648
75	7.14	5.62	0.787
100	9.19	7.61	0.828

Conclusions

This paper proposes a simplified duty cycle-based predictive control, to improve the steady-state performance of the Finite Control Set Model Predictive Control (FCS-MPC). A significant improvement of this approach, is in the capacity to maintain a constant switching frequency throughout the operation of the power converters. Compared to the current state of the art, the proposed approach is simple and effectively reduces the Total Harmonic Distortion (THD) of the input current. Simulation results are presented, which confirms the effectiveness of the proposed approach, as compared to the FCS-MPC.

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