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الجمهـــورية الجـــزائرية الديمقــــراطية الشعبيــــة

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وزارة التــــعليم العــــالــــي و البحث العلمــــي

MINISTRY OF HIGHER EDUCATION AND SCIENTIFIC RESEARCH

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Unconventional Control of Uncertain Disturbed Nonlinear System

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TABLE OF CONTENTS

Chapter 4: Fuzzy Logic Control

ABBREVIATIONS

SISO	Single-Input Single-Output
MIMO	Multiple-Input Multiple-Output
SMC	Sliding Mode Control
HOSMC	High Order Sliding Mode Control
SOSMC	Second Order Sliding Mode Control
FLC	Fuzzy Logic Controller
MPC	Model Predictive Control
TS	Takagi-Sugeno
LQG	Linear-Quadratic-Gaussian
IDCOM	Model Predictive Heuristic Control
DMC	Dynamic Matrix Control
QDMC	Quadratic Program Model Predictive
HIECON	Hierarchical Constraint Control
SMCA	Setpoint Multivariable Control Architecture
SMOC	Shell Multivariable Optimizing Controller
PCT	Predictive Control Technology
VSCS	Variable Structure Control System
APFC	Active Power Factor Correction
THD	Total Harmonic Distortion
PF	Power Factor
PLL	Phase Locked Loop

LIST OF FIGURES

LIST OF TABLES

General Introduction

❖ Control Problematic

The Automatic is an interdisciplinary field of engineering and mathematics, is interested in the behavior of dynamic systems. One of the main goals of the engineer is to design a system, called regulator, capable of controlling a physical system (also called physical process) given, the interest is to obtain a self-regulated processes requiring a minimum of human intervention to operate: for example, the controller speed of a car's role is to keep the vehicle at a given speed despite disturbances such as road conditions or wind.

In practice, the engineer control is faced with two problems:

- ✓ The problem of assessing, if the regulator ensures the desired system behaviour: it is the analysis problem;
- ✓ The problem of designing a regulator to ensure the desired system behaviour: it is the synthesis problem.

To solve these problems for a system and required properties, the engineers are implementing generic methods developed by researchers.

A general method for solving a problem of analysis or synthesis is always defined in a specific context, characterized in part by the nature of the mathematical model chosen to represent the real system (which is never exactly known) and secondly, by the nature of the desired behavior, defined in the specification.

❖ System and model.

The mathematical model describing the relationship between signals that the system exchange with the environment. The model is generally obtained by modeling from the physics laws and parameters identification through experiments. It stills an imperfect representation of the physics process, which by definition can never be known perfectly.

The signals that the system exchanges with the environment are two kinds:

•The input signals, which are typically the control input signals and the perturbation signals (those that can not change): reference signals, disturbances reject, measurement noise.

•The output signals, which are the measured signals (those that are known) and the control signals (those that we want to enslave).

In practice, the signals are measured by sensors and the control signals are issued by actuators. In this context, the objective of the control is to act on the control input so that the output of the controlled system has a certain predefined behaviour, despite the disturbance.

The most common control techniques are the feedback and closed loop controls, is said closed loop when the output of the process is taken into account to calculate the input: the controller then acts to limit the error between the measurement and desired set from the set of available measurements.

❖ Quantitative Specifications

In Automatic, the specifications comprise several types of specifications. First, the system should be stable. It is already a difficult point because there are several notions of stability. Next, it must have certain behavior: we talk about performance that can be quantitative or qualitative. The specifications Quantitative are generally must have certain properties that output signals in response to a certain class of input signals, expressed in the time domain (rapidity, precision, ...) and for linear systems, frequency (bandwidth, gain ...)

for qualitative performances: they are fundamental properties such as convergence of the output to a steady state single. Finally, an implicit specification is robustness: the correction must ensure the desired performance despite uncertainties. Some non-quantitative concepts, such as stability, certain qualitative performance and robustness, are particularly sensitive to formalize.

❖ **Effectiveness of the method**

A method must solve a given system specifications an analysis problem: "Does the system have the properties of the specification?" or synthesis problem: "find if there is a regulator to the system ensuring the properties of specifications ". A method must provide a convenient way to properly address these issues:

❖ **Specifications satisfaction**.

Methods are necessarily based on a choice of formalization (mathematical) of constraints specifications: or, in practice it is not easy to translate all the specifications and some are neglected. Unable consider the translation of certain specifications, a method may be more or less appropriate with regard to actually requested specifications.

❖ **Ease of implementation**.

A method must be a tool for the engineer without too theoretical knowledge advanced to treat industrial problems: this requirement a criterion for comparing existing approaches. For example, the setting implement "classical" methods requires great expertise: In these approaches, the engineer must indeed fix the corrective structure (PI, PID, multi-loop, ...) and set the parameters to meet specifications. They help with graphic criteria (diagrams Bode, Black-Nichols, Nyquist). The process is difficult because there is no method for the choice of the structure and to these settings. In contrast, with the progress computing power of computers, the research effort was directed towards the development methods based on algorithms, such as FLC, SMC, and MPC, Where the regulator are constructed optimally and automatically.

❖ **Effectiveness of digital tools**.

Methods based on algorithms are interesting because they facilitate the process synthesis for the engineer by systematically. An important point is the effectiveness of the algorithm: it must be completed in a reasonable time; typically polynomial (that is to say, the calculation time is a function polynomial of a characteristic size of the system, such as his control). A class particularly interesting problems is constituted by the optimization problems convex.

❖ **Motivation**

If the parameters of a system can be obtained precisely, then its control would be a relatively straightforward problem and model-based approaches such as PID and pole placement could be used. However, in real-time industrial systems, it is often the case that there exist considerable difficulties in obtaining an accurate model. Even when the model is sufficiently accurate, there are many other uncertainties for example due to the precision of the sensors, noise produced by the sensors, environmental conditions of the sensors, and nonlinear characteristics of the systems and actuators. Then, not only does the performance of the model-based approaches drastically decrease, but the complexity of the controller design also increases. In such cases, the nonlinear approaches are generally preferred both for modeling and control purposes. The most common nonlinear controls are sliding mode control, model predictive control and fuzzy logic controller, which presented by its theories background and application in this thesis. For nonlinear system we take a single phase active power factor correction AC/DC boost converter such as an application of our proposed nonlinear controls.

- **Thesis Outline**

Chapter 1: Presents a review on the available literature nonlinear control techniques. This chapter explains the history development of nonlinear control theories. Among them, Fuzzy logic, Model predictive control, and sliding mode control are the best control methods for nonlinear systems due to its fast dynamic response, good stability, and implementation simplicity in real time.

Chapter 2: This chapter presents a basic introduction and useful references for readers who are not familiar with model predictive control. We briefly introduce the theory of model predictive control, then the predictive controls in linear and nonlinear cases are detailed, later an application of finite state predictive control for APFC is presented.

Chapter 3: It describes in detail the structure of high order sliding mode control and a concisely reviews the history of SMC. The application of high order sliding mode control technique for single phase active power factor correction is analyzed, simulated, and implemented via dSPACE 1104. The result shows that the proposed controller offers better steady state performance and transient response.

Chapter 4: in this chapter the fundamental theories of fuzzy logic controller has been presented briefly and an application of FLC for single phase active power factor correction is presented. The performance of the developed control has been evaluated in terms of harmonic mitigation and DC link voltage regulation. The detailed simulation and real-time results are presented to validate the proposed research.

Chapter 1: Background and Literature Review

1.1 Introduction

The historical process in which human beings know and change the world is always a gradual developing process from low class to top grade, from simplicity to complexity, from surface to inner, and the same to the aspect of control area, the early research on control system is all linear. For example, Watt steam-engine regulator, adjusting the liquid level and so on. It was limited by the current knowledge of human being on the natural phenomena and the ability to deal with the practical problems, for the linear system's physical description and its mathematical solving is much easier to achieve. In addition, it has already formed a perfect series of linear theories and research methods. By contrast, for nonlinear systems, except for minority situations, so far there hasn't been a feasible series of general methods. Even if there are several methods that will only solve problems belonging to one category, they can't be applied generally. Accordingly, that is to say, how we study and dispose the nonlinear system control is still staying at the elementary stage on the whole. Besides, from the viewpoint of the accuracy we need to control the system, it will get a satisfactory result within limits to utilize the linear system theories to deal with the most technology problems in engineering at present. Therefore, the nonlinear factors of one real system are often neglected on purpose or substituted for various linear relations.

To sum up, that is the principal cause which prompts the linear system theories to develop quickly and become correspondingly perfect day by day and restricts the nonlinear system theories from getting much attention and developing over a long term. The present system has become more and more complicated, integrative and intelligent, while the request on its performance has become more and more strict. The research on the motion laws and analytical methods of the nonlinear system has already become a significant branch in the automatic control theories. Strictly speaking, the real systems mostly belong to nonlinear system. By contrast, linear systems are ideal models in order to simplify the mathematic problems. The contrast between linearity and nonlinearity is that the last one is a "non propriety". Yet, this main characteristic is a negation of propriety and cannot be used to have any unification in the methods or techniques used to analyze and control such systems. As a result, it determines the complexity on research. For nonlinear systems, in the past the researchers adopted the method of approximate linearization to linearize them, such as Taylor unfolding method, Jacobian matrix method and so on. These methods should be applied to nonlinear systems under the better initial value and the higher accuracy, which are effective to dispose the systems with the working point moving on a small scale. At the same time, the working point in practical systems is always moving. If the point moves beyond the scale or some component works in the nonlinear area, the output will bring strange phenomenon which cannot be explained by linear theories. In addition, there are still some running behaviors such as asynchronous suppression, bifurcation, chaos, strange attractors from complicated systems which belong to nonlinear phenomenon in substance. Obviously, they can't be analyzed by means of linear models. Therefore, the nonlinear problems in mathematics should be solved with the theories and methods in the nonlinear science. During the last two decades of the 20th century, differential geometry theory and its applications have developed fast, so that the research on nonlinear system control theories and their applications got a breakthrough improvement. With the help of differential geometry theory, necessary and sufficient conditions [Hermann, 1977; Isidori, 1996] of controllability and observability and were set up systematically via the research on the dependency relationships among the state, input and output of nonlinear system, which were studied by the basic tools such as Lie bracket. The research models of differential geometry control theory get rid of partial linearization and movement on a small scale and realize the global analysis and synthesization on the dynamic systems. Nonlinear system control theory based on differential geometry theory realizes the linearization of nonlinear systems by means of state feedback and coordinate transformation under some conditions on the control. The point that doesn't neglect any high order nonlinear item in the linearization progress is different from traditional approximate linearization methods. So the linearization is not only accurate but also global, that is to say, it can be applied to the whole scale with definitions [Howze, 1973]. Among others, the linearization based on differential geometry theory in the algorithm still need the accurate nonlinear system models. However, it's impractical to use an accurate mathematic model to describe the dynamic character for a system. So it's inevitable to exist the factors such as parameter uncertainty, time-varying parameters, time delay, external disturbance which will have an influence on controller design. Accordingly, it's also considerably crucial to research on the robustness of nonlinear systems [Huangfu, 2010].

In the recent years, the fuzzy control, nonlinear H^{∞} robust control, expert system, neural nets, adapting control, predictive control, stepping control, sliding mode control and etc. have become the main research aspects. In addition, there are still several research results from some literatures which combine the control technologies together to improve them. Meanwhile, they are all accomplished under the more restrictive conditions on mathematic models. Among the researches on various nonlinear control techniques, each one has its own property. to react on the perturbation of system parameters and external elements, while for it's of strong robustness and its control structure design is simple to come true, sliding mode, and fuzzy logic, and predictive have been an indispensable branch in the system robustness control theories.

1.2 The developing history and present state of nonlinear control theories

1.2.1 The developing history of nonlinear control theories [Huangfu, 2010]

In the 1930s to the 1940s, so many scholars such as Nyquist, Bode, Weiner, Nichols, Routh, Hurwitz strived to construct classic control theory based on frequency domain method and root locus technique. They adopted Laplace conversion as the mathematic tool and mainly studied the linear time-invariant system of SISO. They also converted the differential equation and difference equation which describe the system into complex number field in order to get transfer functions of the system. Generally speaking, they used feedback control to form the so-called closed loop control system. But there were still several obvious limitations in classic control theory. Especially, the theory is very difficult to apply in time-varying systems and nonlinear systems availably and to show the much deeper characteristics. Consequently, the above promote modern control theory to develop. In modern control theory, the multi-input multi-output system was studied thoroughly. And then it is pretty significant to construct basic theory which depicts the essences of control systems such as controllability, observability, realization theory, decomposition theory and so on. Meanwhile, it prompts control to develop from a class of engineering design methods to a new science. Based on state-space method, modern control theory adopts linear algebra and differential equations as the main mathematical tools to analyze and design control systems. The state-space method is essentially a time domain approach, which not only describes the external characteristics of the system, but also describes and reveals the internal states and performances of the system. The goals of analysis and synthesis are to reveal the inherent laws of the system and then to realize the optimization of the system in a certain sense. In principle, it can be a single variable, linear, time-invariant and continuous, and it can also be multi-variable, nonlinear, time-varying and discrete. At the same time, its existence is to solve many practical control problems at a high level from theory to application and to promote nonlinear control, adaptive control, robustness control, etc. to be an independent science branch with fruitful achievements. Before the development of nonlinear control theory, nonlinear controller has been applied in industry, such as a variety of relay control because of its reliable structure and good performance. Early, the study of nonlinear system control has made some significant progress. Main methods are describing function method, phase-plane method and Lyapunov method. These methods have been widely used to solve the practical problems of nonlinear systems. However, these methods still cannot become a general method to analyze nonlinear systems.

To summarize the study results in the historical stage, the researching problems principally center on the absolute stability of the system, which limit the nonlinear term to a fan-shaped domains and allow a linear function to replace the nonlinear function.

1.2.2 Present state of nonlinear control theories [Huangfu, 2010]

Since the 1970s, nonlinear control system theory and its application research have achieved a breakthrough development. The successful application of modern mathematical tools such as modern differential geometry and differential algebraic theory played a key role in the area. For the input and output response of a nonlinear system, Slotine, Khalil, Isidori and others adopted state feedback approach and used Lie algebra to linearize it accurately. Feedback linearization: its basic idea is to use algebraic transformation to convert the motion characteristics of a nonlinear system all or partially into linear dynamic characteristics. As a result, it can be analyzed by well-known linear control theory. This approach is completely different from approximate linear method. The difference is that, feedback linearization is achieved by means of the rigorous state conversion and feedback rather than the linear approximation of dynamic characteristics. Feedback linearization can be considered as a method that transforms the original system model into the form of a relatively simple equivalent model. The design method of feedback linearization also has some limitations. For example, it does not apply to all nonlinear systems, when it only applies to the system which is a smooth nonlinear system with precise mathematical model. When parameters are uncertain or model dynamic characteristics are not created, the system robustness will not be guaranteed. In order to overcome the above shortcomings, people are launching the ongoing active research to process nonlinear essence; in this thesis we are introducing robust controls, such as fuzzy logic, sliding mode control, and predictive control in order to increase the robustness of the system with uncertain parameters.

1.3 The basic principle of sliding mode control theory

The concept of sliding mode control first appeared in Russian literature in the fifties of the twentieth century. The former Soviet Union experts Emelyanov first proposed the concept of sliding mode variable structure. In fact, before that, 1932, V. Kulebakin used in variable structure control and DC generator for an aircraft in 1934; Nikolski used the relay to operate the ship trajectory, these can be considered as earlier "Sliding Mode Control" [Utkin, 1999]. Later, Utkin has written an English summary of papers on the sliding mode control [Utkin, 1977]. Then, the sliding mode control theory was widely disseminated to the different areas. Seventies, sliding mode variable structure system with its unique advantages and characteristics attracted western scholars' attention, which was subsequently a number of scholars from different theoretical point of view, using a variety of mathematical tools for their in-depth research. They make the sliding mode control theory gradually developed into an independent research branch. In the linear system control theory, after the single-input single-output (SISO) and multiple-input multiple-output (MIMO) system establishing standard type, the sliding mode control strategy had been further in-depth study [Edwards, 1998; Utkin, 1992; Utkin, 1999]. In which, R. A. DeCarlo designed the sliding mode controller for multi-variable nonlinear system [DeCarlo, 1988]; Later, the theory of nonlinear systems differential geometry had been developed [Isidori, 1995]. This theory will soon be applied to sliding mode control and related fields. The sliding mode control itself is a nonlinear controller, so its application is not limited to linear systems, also applies to nonlinear systems. If a nonlinear system using approximate linearization into a linear system, then designing sliding mode controller, the effect is clearly not as good as designing a controller for nonlinear system directly. However, for the nonlinear systems, it should be converted to standard affine form usually, which is not all nonlinear systems can be accomplished. Proceeding from the practical problems, Slotine, who has designed a sliding mode based on input-output decoupling controller [Slotine, 1983; Slotine, 1993], who's sliding mode surface is composed by the Hurwitz polynomial on output error and error derivatives. Differential algebraic theory first appeared in Fliess research results in [Fliess, 1990a; Fliess, 1990b]. He has opened up a new direction for nonlinear systems sliding mode control, who proposed a general algorithm about nonlinear system converting into a controllable standard form. The algorithm used the dynamic system state and input output decoupling. Sira-Ramirez presented a controllable standards design approach based on sliding mode control, using this method designed a special sliding surface [Sira-Ramirez, 1992; Sira-Ramirez, 1993]. X.Y. Lu and S.K. Spurgeon also raised the standard model based on general sliding mode control strategy of nonlinear control system [Lu, 1998]. However, the drawback of sliding mode controller is the high frequency oscillations (chattering phenomenon), which is a major obstacle for the implementation of standard SMC. For these reasons, the high order sliding mode control (HOSMC) is a new method proposed in last few decade [Levant, 1985; Levant, 2003; Levant, 2005] to overcomes the main drawbacks of classical SMC, its provides a smooth control, good performance yielding to less chattering in real time implementation, and better convergence while preserving the robustness properties. In conventional SMC design, the control aim is to move the system state into sliding surface S(t)=0. A sliding mode is said ``r^{th} order sliding mode'' if $S(t) = \dot{S}(t) = \cdots = S^{(r-1)} = 0$. In HOSMC, the purpose is to force the state (error) to move on the switching surfaces and to keep its (r-1) first successive derivatives null. More specifically, second order sliding mode control aims for $S(t) = \dot{S}(t) = 0$, that is, the controller's aim is to steer to zero at the intersection of S(t) and $\dot{S}(t)$ in the state space [Bouafassa, 2014b].

1.4 The basic principle of fuzzy logic control theory

One way to avoid the complexity of the mathematical representation of a nonlinear system is to use a fuzzy model [Tanaka, 2001]. With this concept, human expertise reflects the behavior and state relations, inputs and outputs of the process by using the fuzzy rules [Takagi, 1985]. In effect, progress and investment in fuzzy theory from its introduction by Zadeh [Zadeh, 1965] demonstrated that fuzzy system can be used as a universal approximator [Kosko, 1994; Zeng, 2000; Wang, 1994; Tanaka, 2001; Bouafassa, 2014a]. Taking into account both premise and conclusion parties form a fuzzy rule, there are two essential families of fuzzy model: Mamdani [Mamdani, 1974] and Takagi-Sugeno (TS) [Takagi, 1985; Sugeno, 1988; Tanaka, 2001]. The second family differs remarkably; although it keeps the fuzzy formalism in premise part that the inference is just an interpolation of the consequent. As result, after defuzzification, we obtain a nonlinear arithmetical function connecting inputs and outputs of the system. This property is used to represent nonlinear system interconnection models affine and linear around some operating points, which are associated with weighting functions called activation functions. There are two ways to determine these functions either by identification from a real system [Gasso, 2002], or by using a knowledge model that will provide its exact representation in a solid portion of the state space [Wang, 1994; Tanaka, 2001]. Theoretically fuzzy system gives us reason to approximate any nonlinear system, in practice this representation is related to a number of knowledge rules. Whatever the number, large or small, modeling errors are taken into account.

Since the introduction of fuzzy theory, there was a small community of researchers who have invested in the fuzzy control. Despite the success of fuzzy controllers in the industrial field, they were often criticized for lack of results on the stability and explanations on the robustness of systems controlled by fuzzy logic [Boukezzoula, 2000]. Only from the 90s the stability and robustness have been proposed for fuzzy logic controller [Wang, 1994; Tanaka, 1998]. Then, several approaches and evidence have emerged, including fuzzy classic called nowadays fuzzy logic type-1 [Wang, 1996; Mansouri, 2009; Chaoui, 2001; Essounbouli, 2006; Wang, 1994; Tseng, 2001] and fuzzy logic type-2 [Chafaa, 2007; Hussain, 2011; Manceur, 2012; Castillo, 2008; Mendel, 2013; Mendel, 2013; Karnik, 1999]. The fuzzy logic type-1 has a lot application such as automatic control, data classification, decision analysis, expert systems, time series prediction, robotics, and pattern recognition.

Type-2 fuzzy set was initially proposed as an extension of classical (type-1) fuzzy sets. Type-2 fuzzy sets are very useful in circumstances where it is difficult to determine an exact membership function for a fuzzy set; hence they are very effective for dealing with uncertainties. However, type-2 fuzzy sets are more difficult to use and understand than type-1 fuzzy sets. Even in the face of these difficulties, type-2 fuzzy logic has found applications in many fields. Fuzzy logic type-2 systems consist of rules, where fuzzy sets used are type-2. By definition, these fuzzy sets include uncertainty in their membership functions. We can say that the logic type-2 fuzzy is a generalization of conventional fuzzy logic in the sense that the uncertainties are not just limited to linguistic variables (words), but also taken into account in the definition of the functions belonging.

1.5 Model predictive control

Model predictive control has had an exceptional history with early intimations in the academic literature coupled with an explosive growth due to its independent adoption by the industries process where it proved to be highly successful in comparison with alternative methods of multivariable control. It's phenomenal success in the industries process and was mainly due to its conceptual simplicity and its ability to handle easily and effectively complex systems with hard control constraints and many inputs and outputs. Several publications provide a good introduction to theoretical and practical issues associated with MPC technology. [Rawlings, 2000] provides an excellent introductory tutorial aimed at control practitioners. [Allgower, 2000] present a more comprehensive overview of nonlinear MPC and moving horizon estimation, including a summary of recent theoretical developments and numerical solution techniques. [Mayne, 2000] provide a comprehensive review of theoretical results on the closed-loop behavior of MPC algorithms. Past reviews of MPC theory include those of [Garcıa, 1989; Richalet, 1978; Muske, 1993; Rawlings, 1994]. Several books on MPC have been published recently [Camacho, 2004; Grüne, 2011; Magni, 2011; Rodriguez, 2012].

Fig.1.1 shows an evolutionary tree for the most significant industrial MPC algorithms, illustrating their connections in a concise way.

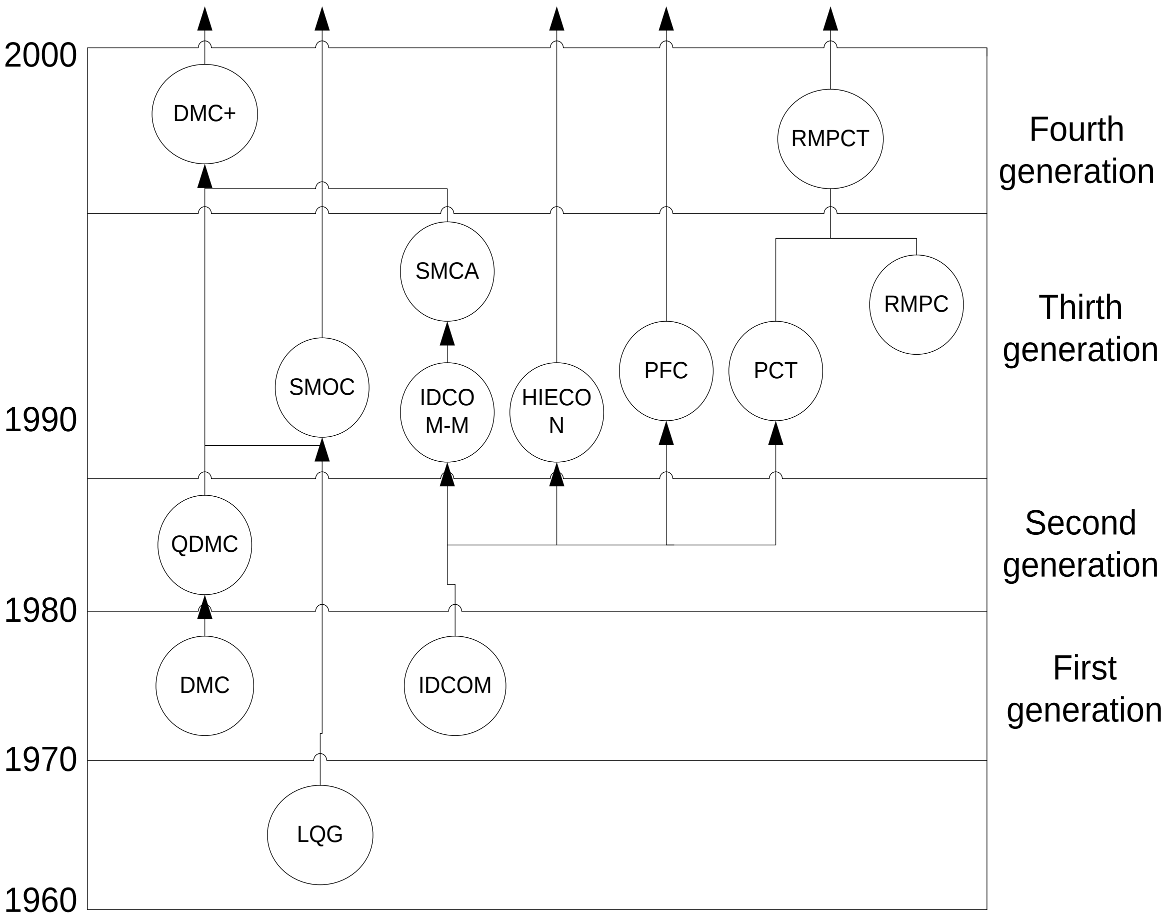


Fig. 1.1 Approximate genealogy of MPC algorithms.

1.6 The main research contents of the thesis

Based on the research of the current nonlinear control theory applied to nonlinear systems, it chiefly launches an in-depth study and exploration in allusion to several facing issues of nonlinear control theory and gives the corresponding research findings and results.

The typical goals of process control are:

• Disturbance rejection to decrease variability in the key variable.

• Stable and safe operation.

Chapter 2: Model Predictive Control

2.1 Introduction

Model Predictive Control (MPC) is an advanced control methodology which uses explicitly the system model to predict the future evolution of the process. The basic idea of MPC consists in computing an optimal control sequence over a prediction horizon at each decision instant, by minimizing some given cost function expressing the control objective. The first control signal is scheduled to b e applied to the system during the next sampling period and this optimization process is successively repeated at each sampling time. MPC is known in the industrial world, especially, in petrochemical sector due to the slow dynamics of this system. With its ability to take into account the constraints, control of multivariable systems and the possibility to use different model structures, MPC is now widely recognized as one of the most powerful control techniques to solve many control problems. Over time, many improvements have been made on this technique, including the work of Mayne and Michalska [Mayne, 1990]. Today it is also suitable for fast control systems. The prediction horizon keeps being shifted forward and for this reason MPC is also called receding horizon control (RHC). Fig. 2.1 illustrates a simplied view of MPC.

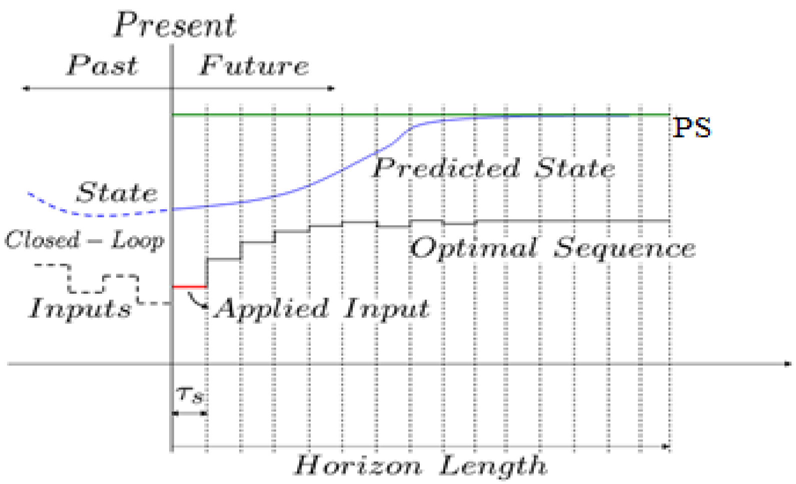


Fig. 2.1 Schematic view of the MPC strategy.

2.2 Predictive approaches

Our goal in this section is to give a detailed of predictive control to allow the reader to get an idea of the different approaches in the field, both for linear predictive control or nonlinear one. The presentation adopted here, which is to separately expose the academic and industrial cases for each part of the predictive control, linear or not, is only made for clarity. Far from us the idea that there it would be a distinct boundary between academic and industrial case for the linear predictive control, which it would be absolutely distinct from the nonlinear.

2.2.1 Linear predictive control

2.2.1.1 Industrial case

In 1962, the close relationship between the optimal control problem and programming Linear were recognized, first by Zadeh [Zadeh, 1962]. In 1963 Propoi [Propoi, 1963] proposes a receding horizon approach which is the heart of all the predictive control algorithms (MPC). It is known as the "Open-Loop Optimal Feedback [García, 1989]. In 1967, Lee [Lee, 1967] proposes a close algorithm to MPC [Qin, 2003]. Rediscovered in 1976 following the work of Richalet [Richalet, 1976] as the Model Predictive Heuristic Control (MPHC) [García, 1989]. Predictive control experienced during recent decades a growing rise to industrial applications. Since the method IDCOM for IDentification-COMmand [Richalet, 1978], several methods have been proposed for the industrial applications of MPC such as Cultler and Ramaker in 1979, which have developed within Shell Oil, an unconstrained multivariable control algorithm named "Dynamic Matrix Control (DMC)." This algorithm has been improved following the work of Prett and Gillette [Prett, 1979] that allow take into account the nonlinearities and constraints of the system. Other methods have also been proposed and have found various application fields nearly in aerospace, industry and petrochemicals, [García, 1989; Qin 2003].

2.2.1.2 Academic case

From the academic point of view, other predictive control methods based on adaptive control have been proposed. Thus, we cite the work of Peterka [Peterka, 1984] and the algorithm known enough Generalized Predictive Control (GPC) [Clarke, 1987]. However, all these methods have been developed in the part of the discrete linear predictive control. We give a few more details on the generalized predictive control in the following.

❖ **The approach GPC (Generalized Predictive Control)**

Proposed in 1987 [Clarke, 1987], this method is based on a model CARIMA (Controlled Self-Regressive and Integrated Moving-Average) whose equation is given by:

$$A(q^{-1})y(t) = B(q^{-1})\,u(t-1) + C(q^{-1})\delta(t)/\Delta \tag{2.1}$$

Where A and B are polynomials in (q^{-1}) such as:

$$\begin{aligned} A(q^{-1}) &= 1 + a_1 q^{-1} + \ldots.. + a_{na} q^{-na} \\ B(q^{-1}) &= b_0 + b_1 q^{-1} + \ldots.. + a_{nb} q^{-nb} \end{aligned} \tag{2.2}$$

and the operator of differentiation Δ is equal to $1\text{-}q^{-1}$, see [Clarke, 1987; Tuffs, 1985]. The prediction output vector is based on a Diophantine equation given by:

$$1 = E_j q^{-1} A\Delta + q^{-1} F_j\left(q^{-1}\right) \tag{2.3}$$

Where E_j and F_j are polynomials defined from $A(q^{-1})$ and the prediction interval j. Hence, the equation for the predicted output is:

$$y\left(t+j\,|t\right) = G_j \Delta u(t+j-1) + F_j y(t) \tag{2.4}$$

With

$$G_j\left(q^{-1}\right) = E_j B = B\left(q^{-1}\right)\left[1 - q^{-1} F_j\left(q^{-1}\right)\right]/A\left(q^{-1}\right)\Delta \tag{2.5}$$

GPC control law is developed to minimization of a quadratic criterion built on the error between the output and the reference signal and the weighted control. However, no stress on the input or on the output is taken into account in this method. It was not until 1993 with the work of Camacho, where the constraints are taken into account in the GPC [Camacho, 1993]. In addition, other methods based on the continuous-time systems have been developed for predictive control. One of them, named Continuous-time Generalized Predictive Control (CPMF) was proposed in 1987 [Gawthrop, 1987]. However, its basic algorithm appears only two years later. A brief statement mono-variable case is given in the following.

❖ **The CGPC method (Continuous-time Generalized Predictive Control)**

It is based on system models SISO strictly proper:

$$Y(s) = \frac{B(s)}{A(s)}U(s) + \frac{C(s)}{A(s)}V(s) \tag{2.6}$$

Where A, B and C are polynomials according to the Laplace variables. The variables Y(s) U(s) and V(s) denote, respectively, output, control and disturbance.

The prediction of the output is made from developing *Maclaurin* series truncated [Demircioglu, 1991]. It is given by:

$$\hat{y}(t+T) = y(t) + \sum_{k=1}^{N_y} y_k(t)\frac{T^k}{k!} \tag{2.7}$$

Where

$$y_k(t) = \frac{d^k \hat{y}(t+T)}{d(t+T)^k}(T=0) = \frac{d^k y(t)}{dt^k} \tag{2.8}$$

and N is the prediction order.

The CGPC control law is obtained from the minimization of a quadratic criterion built on the prediction error between the output and the reference and the developing Taylor series of the weighted control signal such that,

$$J = \int_{T_1}^{T_2} [y_r^*(t,T) - \omega_r^*(t,T)]^2 dT + \lambda \int_{0}^{T_2-T_1} [u_r^*(t,T)]^2 dT \tag{2.9}$$

With

$$y_r^*(t,T) = y_r^*(t,T) - y(t) \quad and \quad u_r^*(t,T) = \sum_{k=0}^{N_u} u_k(t)\frac{T^k}{k!} \tag{2.10}$$

More details are given in [Demircioglu, 1989]. The CGPC was proposed for both mono-variable systems [Demircioglu, 1991] and multivariable [Demircioglu, 1992a]. Some studies of this technique takes into account the constraints proposed in [Demircioglu, 1999]. Also, some properties of the closed-loop stability guaranteed for CGPC are given in [Demircioglu, 1992b]. They are mainly based on two points [Kwon, 1977]:

- The first is to force the state of the system to zero when the prediction horizon is reached,

-The second point, suggested adding a quadratic weighted in the final state of the system (when the prediction horizon is reached) for quadratic starting criterion.

Thus, when the weight tends to infinity, the second point coincides with the first one. However, if it tends to zero, this is leads to treating the problem of CGPC with his initial criterion, [Demircioglu, 1992a]. Some comparative studies have been made between the linear continuous-time predictive control and the discrete time [Demircioglu, 2000]. Robust stability analysis of the CGPC is given in [Wang, 2006].

2.2.2 Nonlinear predictive control

The nonlinear model predictive control (NMPC) is a robust control technique, because it can work with uncertainties and disturbances. This property comes from the fact that the NMPC is very close to optimal control. Some results on the inherent robustness of NMPC are given in [Magni, 1997; Chen 1982]. Other methods taking into account directly uncertainties and disturbances are based on min-max formulation. We will not give details of these methods control, the main ones:

- Robust NMPC solves a min-max problem in open loop, [Lall, 1994].
- Robust NMPC through the optimization of a state feedback controller [Kothare, 1996].

- Robust NMPC using optimization "multi-objective" [Darlington, 2000].
- The NMPC control-based H^{∞} [Magni, 2001].

According to work of Allgöwer and Findeisen, the key points of the nonlinear model predictive control (NMPC) are as follows:

- ✓ Direct use of nonlinear models for the prediction;
- ✓ Explicit consideration of the constraints on the input and state;
- ✓ Minimization a cost function;
- ✓ predicted Behavior is generally different from the behavior in closed loop;
- ✓ The need for real-time solution of an optimal control problem in open loop for the application;
- ✓ Availability of system states to measure or estimate the future prediction.

2.2.2.1 Industrial case

The NMPC has many applications in the industrial world. However, the last one hundred, are much less numerous than the linear predictive control (4500 industrial applications [Qin, 2003]. For More details the reader is referred to [Qin, 2003]

2.2.2.2 Academic case

Many efforts have been made in recent years to adapt the nonlinear model predictive control systems, see [Allgöwer, 1998]. However, it will never reach the simplicity of the linear predictive control [Mayn, 1997].

❖ **General problem of nonlinear model predictive control**

Consider the nonlinear system as follow:

$$\dot{x}(t) = f(x(t), u(t)) \tag{2.11}$$

With $x(0)=x_0$, subject to the constraints of input state. To Solve a NMPC problem (with constraints), we should solve the optimal control problem in open loop following finite horizon.

$$\min_{\hat{u}} J(x,\hat{u},T_p) = \int_t^{t+T_p} \left(\|\hat{x}(\tau;x(t),t\|_Q^2 + \|\hat{u}(\tau)\|_R^2 \right) d\tau \tag{2.12}$$

Subject to

$$\begin{cases} \dot{\hat{x}}(t) = f(\hat{x},\hat{u}) \\ \hat{x}(t;x,t) = x(t) \end{cases} \tag{2.13}$$

And

$$\begin{cases} \hat{x}(\tau;x(t),t) \in X \\ \hat{u}(\tau) \in U. \end{cases} \tag{2.14}$$

The previous cap variables are those estimated.

The resolution of this problem raises two major obstacles:

- ✓ Stability for systems constrained finite horizon;
- ✓ Cumbersome numerical computation. Indeed, a nonlinear optimization problem has resolved "online" and it is not guaranteed to find a global optimum, or at worst cases, even local, see [Chen, 2000].

To overcome these drawbacks, several methods have been proposed.

On the first hand, to solve the stability problem for a system constrained to finite horizontal, the first interesting result was presented in [Mayne, 1990]. They have introduced a terminal equality constraint in the criterion

$$\hat{x}(t,T_p) = 0 \tag{2.15}$$

Or, they force the state of the system to zero at the end of the prediction horizon T_p. This return to the criterion gives reason to the wise: *"When we do not know where we are going, we return where we come"*. Similar results were proposed in [Rawlings, 1993]. An illustration of this idea is given in Fig. 2.2.

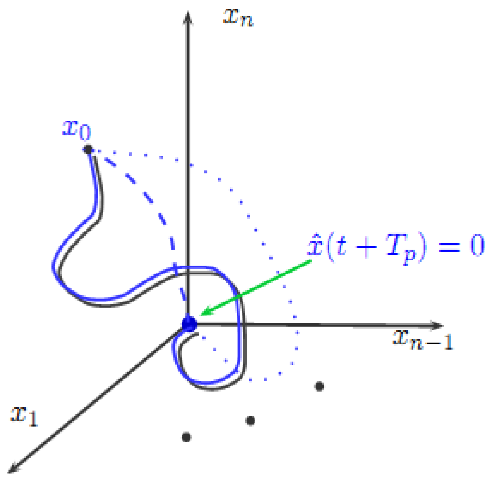


Fig. 2.2 Dynamics of the system with a terminal equality constraint.

However, in terms of numerical computation, to solve a problem of dynamic optimization with equality constraint is very heavy, it is not possible to solve in a finite time [Chen, 2000]. In addition to this, the terminal constraint imposed is summed up in a very small area: point. To avoid this problem, [Michalska, 1993] proposed to expand the terminal area, see Fig. 2.3, by relaxing the terminal equality constraint in profile of a constraint terminal inequality, always on the state. It is such that:

$$\hat{x}(t, T_p) \leq 0 \tag{2.16}$$

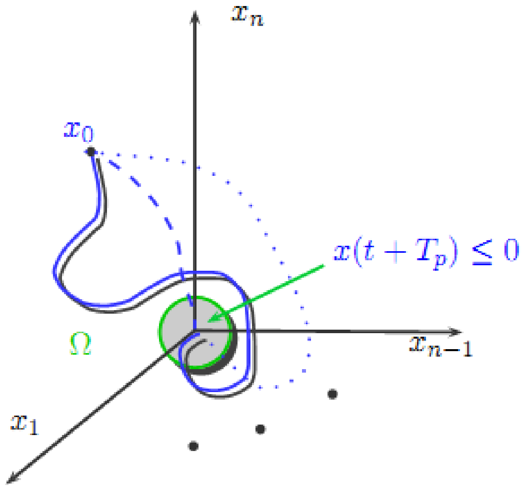


Fig. 2.3 Dynamics of the system with a terminal constraint inequality.

With this method, the state space of the system considered is divided into two parts:

- ✓ A first part in which the predictive control law is applied, corresponding outside of the terminal Ω, whether $R^n \backslash \Omega$
- ✓ And a second part, which is the terminal Ω, where a feedback control law of linear state is applied.

Following their work, Chen and Allgöwer [Chen, 1998] proposed to introduce a penalty terminal $\left\|\hat{x}(t+T_p;x(t),t)\right\|_P^2$ in the criteria to ensure the stability of the system in the terminal area. This method is called Quasi-Infinite Horizon Nonlinear Model Predictive Control. It guarantees the stability of system loop.

❖ **QIH-NMPC (Quasi-Infinite Horizon NMPC)**

This is a technique based on linearization by approximation of nonlinear system considered. With this method and in each sampling interval, the optimization nonlinear problem resolved "online" with constraints on input and terminal inequality constraints.

A side from the above methods, the stability of the NMPC, other methods based on a reverse optimal approach (that is to say such that the control law is also optimal if the optimal control problem has been turned into a problem with an infinite horizon) were examined by Magni and Sepulchre, see [Magni, 1997].

Although other rigorous results for the stability of the NMPC have been established, they are not implementable in practice [Magni, 1997].

On the other hand, the digital calculation of the length of NMPC generates two sub-problems implementation: a delay problem and a global minimum problem since in general, the nonlinear constrained optimization problem is non-convex problem [Chen, 2000]. Faced with such a situation and following the works of Chen and Allgöwer [Chen, 1998], Scokaert [Scokaert, 1999] proposed an alternative for discrete systems where they emphasize that the feasibility implies stability for MPC with linear state feedback. The problem of delays calculations was investigated by many authors including Pierre and Pierre, [Pierre, 1995; Rattan, 1989]. Other proposals have been made for linear systems; see [Wissel, 1997]. They have proposed a generalized predictive control delay whose control depends more on past and present measures. Although the computational burden is heavier for NMPC only the algorithm proposed by Ronco [Ronc, 1999] called Open-Loop Feedback Intermittent Optimal controller (OLIFO) considers delay calculation for its implementation, [Chen, 2000]. Also, the delay measurement is considered in the work of Chen and Allgöwer [Chen, 1998]. [Chen, 1999b] proposed an algorithm to resolve the problems of stability and computational burden posed by NMPC. With this method, optimization "online" is not necessary and the stability is guaranteed. Furthermore, it is shown by the dynamic inversion control method is a special case of this algorithm MPC. However, this method is only applicable to nonlinear systems dynamic stable zeros and well defined degree relative, [Chen, 2000].

Another approach has been proposed for solving optimization problems "online". Indeed, [Lu, 1995; Soroush, 1997] have set the control signal order at zero, which is to make the constant control effort in an interval of prediction given. However, this method has a limit as the developing Taylor series can only be made up to one order equal to the relative degree. Another method was proposed in 1998 by Siller-Alcalá, under the name of Nonlinear Continuous-Time Generalized Predictive Control (NCGPC) [Gawthrop, 1998]. This is a method of control that is based on the Taylor series and is robust. In addition, it allows avoid optimization problems "online" since the predictive control law is calculated lated "offline". Only the state's actions are made to update the control law. Also, this technique is done in a closed loop and is very close to the linearization method input/output state feedback [Gawthrop, 1998]. A brief review is given in the following.

❖ **NCGPC (Nonlinear Continuous-time Generalized Predictive Control)**

The objective with this control method is to make the asymptotic track of output y(t) on the reference while minimizing a quadratic criterion built on error prediction between the reference and the output. As we will see the algorithm is based on a nonlinear model affine in control as:

$$\dot{x} = f(x) + g(x)u \tag{2.17}$$

The prediction of the output is made from its development in Taylor series. Its equation given by: y (t + τ)

$$\hat{y}(t+\tau) = \sum_{k=0}^{\rho} \frac{\tau^k}{k!} y^{(k)}(t) \tag{2.18}$$

It is the same for the reference. The NCGPC control law is developed from the minimization of a quadratic criterion analytically. In addition to its robust character, NCGPC solves the systems control problems with non minimum phase, but also systems whose relative degree is poorly defined. Approach more generally about this control technique for a class of nonlinear systems using approximations is given in [Chen, 2004]. Furthermore, studies have been conducted on the PID controllers (proportional-integral-derivative) Nonlinear predictive for SISO systems [Chen, 1999b] and MIMO [Feng, 2002]. A comparative study between LQR control (Linear Quadratic Regulator) and NCGPC; watch between them that the closed loop stability is obtained for the NCGPC with a finite horizon prediction and a coordinate’s change is adequate. In addition, as shown by Chen, [Chen, 2003], whatever the prediction horizon fixed, the loop system is unstable if its degree relative is strictly greater than four. We will return. Indeed, does not imply optimality stability.

2.3 Description of the proposed model predictive control for PFC

2.3.1 Modeling of the system

The basic model of the boost converter is defined according to the state of switch M. When the switch M is turned on (u=1), the voltage across the transistor is equal to zero and the diode is closed. Fig. 2.4.b shows the equivalent circuit of the boost converter in the ON state. As soon as the switch M is turned off (u=0), the voltage across the diode is equal to zero, Fig. 2.4.c shows the equivalent circuit in this operation mode [Kessal, 2011]. The state space model for the boost converter governed the real switched system can be expressed as follow [Kessal, 2014; Bouafassa, 2015].

$$\begin{cases} C\dfrac{dV_0}{dt} = (1-u)i_L - i_0 \\ L\dfrac{di_L}{dt} = V_{in} - (1-u)V_0 \end{cases} \tag{2.19}$$

When a non-controlled rectifier is connected to the source voltage to ensure the converted AC/DC as shown in Fig. 2.4, the current drawn from the source i_s will be very distorted (high THD) and not in phase with the input voltage Vs, which increases the reactive power and lead to a low input power factor. To solve this problem, one possibility is to add a PFC circuit using two control loops. The dc-bus voltage is sensed and compared with its reference value V_{ref}. The obtained error is used as input for the voltage loop controller, the output of the voltage loop controller I_{max} multiplied by |sinωt| obtained from PLL is the reference current i_{ref}, the reference current is compared with the inductor current i_L. The obtained error is used as input for the current loop controller that calculates the duty cycle.

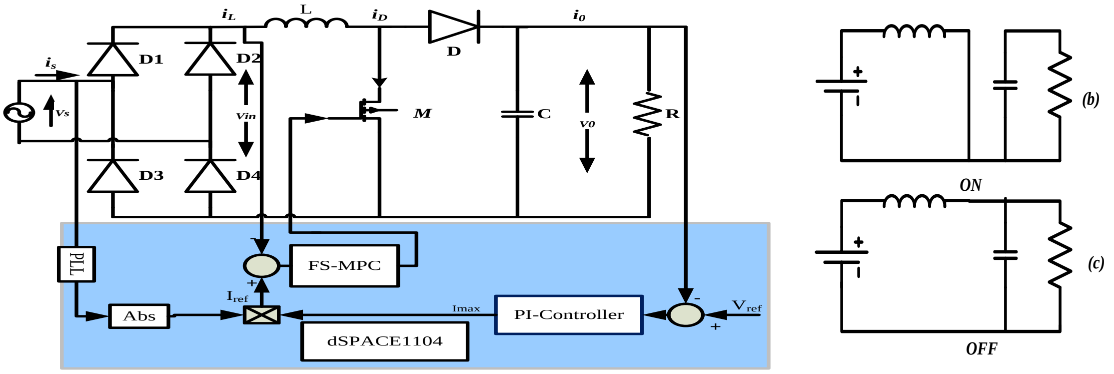


Fig. 2.4 PFC pre-regulator with FS-MPC.

2.3.2 Predictive current control

The main characteristic of FS-MPC (Finite state model predictive control) technique is the use of the system model for predicting the future behaviour of the variables to be controlled for each of the valid switching states. The controller uses this information to obtain the optimal control action.

The discrete time model of a single active power factor correction is used to derive Eq. (19), considering the sampling period T_s, when the switch is turned *on* or *off* the predicted control variables is given by

$$i_L(k+1) = i_L(k) + \frac{T_s}{L} v_{in}(k) \tag{2.20}$$

$$i_L(k+1) = i_L(k) + \frac{T_s}{L} \left(v_{in}(k) - v_o(k) \right) \tag{2.21}$$

The behaviour of the controlled variables i_L can now be predicted for the next sampling interval t(k+1), in order to obtain control actions for both the present time and a future period. One-step horizon predictive controller inputs measured values of i_L, V_{in}, and V_o estimating the future behaviour of the controlled variables based on the evaluation of a cost function. The determination of the cost function is a key factor in FS-MPC represents the deviation of the controlled variables from the desirable values of the reference current and is expressed as

$$J = \left| i_L(k+1) - i_{ref} \right| \tag{2.22}$$

The cost function assures the tracking of the inductor current i_L from the reference current i_{ref} provided by outer voltage loop. For each sampling step the cost function is evaluated twice for each switching state. Evaluation of cost function for different switching states determines the control actions for the next time instant. Fig. 2.5 presents the FS-MPC process. The dotted line corresponds to the reference current. At the sampling time t(k) the FS-MPC has to decide between S_0 and S_1 on basis of minimizing the cost function, the black line corresponds to the finally performed actions, while the faded line are discarded choices. All steps of the proposed FS-MPC method are presented in Fig. 2.6.

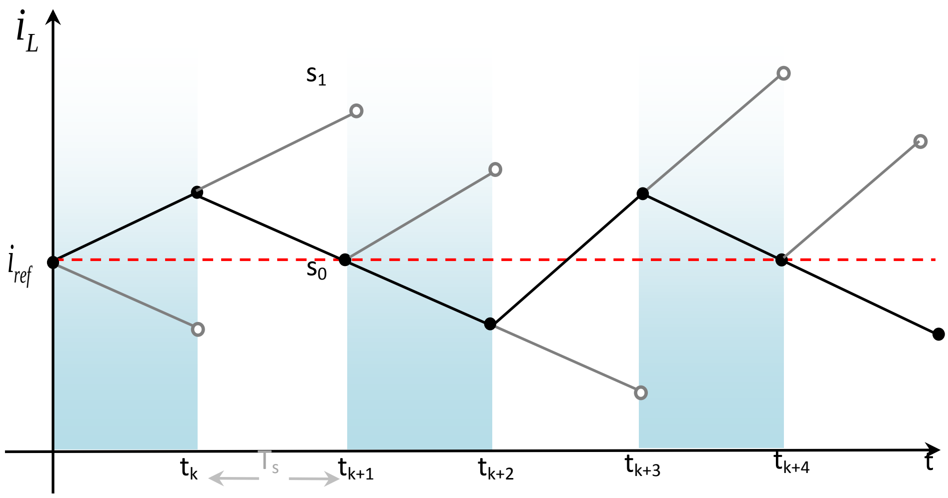


Fig. 2.5 Diagram of the FS-MPC process.

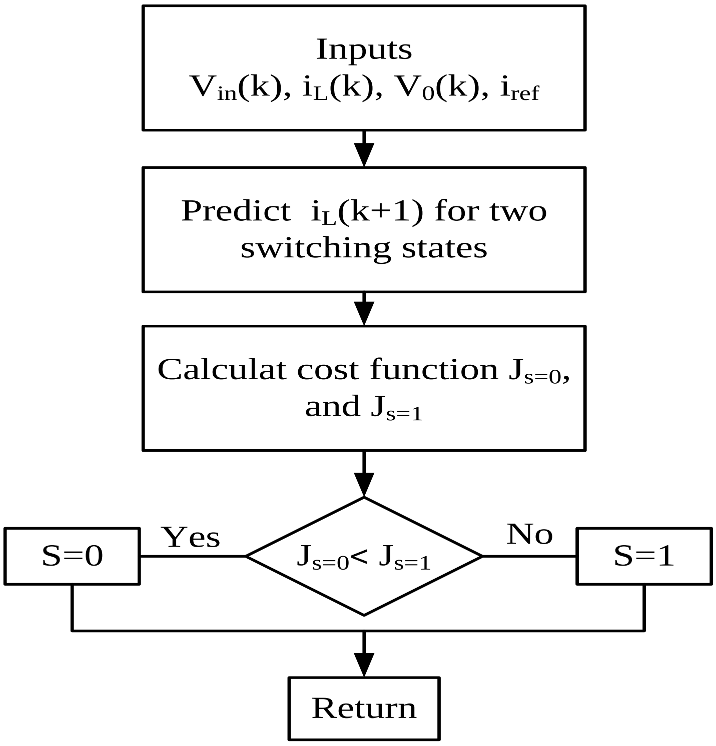


Fig. 2.6 Flowchart of FS-MPC.

2.3.3 Voltage loop controller

The block diagram of a PI controller is shown in Fig. 2.7. The dc-link voltage is sensed and compared with its reference value; the obtained error is directly used by PI regulator. The output of the PI controller must be restricted to a definite values to avoid overshoots/undershoots.

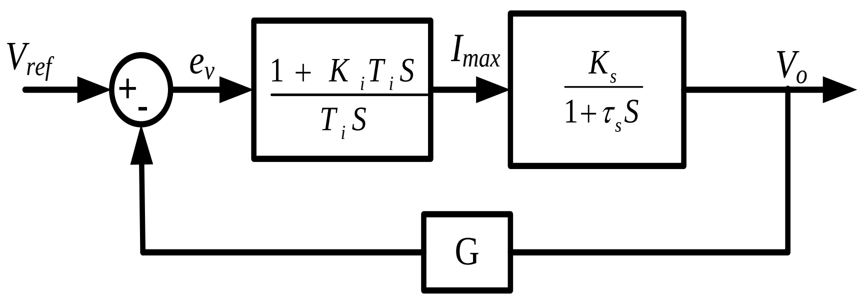


Fig. 2.7 PI-Controller.

Where: $\frac{1+K_i T_i S}{T_i S}$ is the transfer function of the PI controller.

The system from Fig. 2.4 is modelled as a first order system and can be written as

$$\frac{V_0}{I_{max}} = \frac{V_{SM}}{4V_{ref}} \frac{R}{1+\frac{RC}{2}S} = \frac{K_s}{1+\tau_s S} \tag{2.23}$$

For optimal performance of the PI controller the parameters K_p and K_i are take the values 0.1 and 100 respectively.

2.3.4 Simulation results

The circuit was simulated and tested for different loads and output voltage, Fig. 2.8 show the input voltage and current waveforms along with the output voltage simulation results for fixed output voltage of 100V and load (100 Ω), the simulation result confirm that the line current is sinusoidal with nearly unity power factor. Fig. 2.9 shows the measured waveform of the output voltage response during output voltage transient test. The output voltage was stepped up from 100V to 130V (Peak) for about 3sec before is stepped back to 100V. Fig. 2.10 shows the measured waveform of the output voltage response during load transient test. For the transient test, the load was stepped from 100Ω to 150Ω and kept for about 1sec before it was stepped back to 100Ω. It can be noted that the proposed controller has less overshoot/ undershoot and faster settling time. As can be seen from both Figs. 2.9 and 2.10 that the output voltage has reasonable overshoot during input voltage and load transient changes.

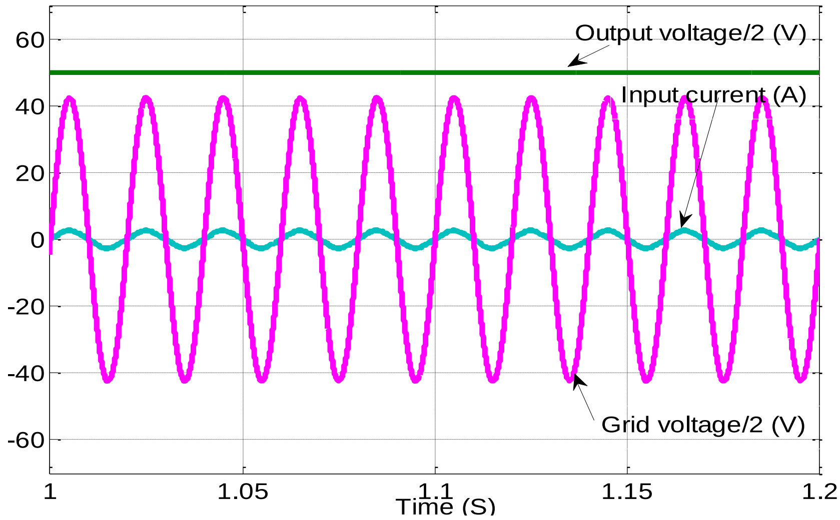


Fig. 2.8 Signal waveforms in the steady state.

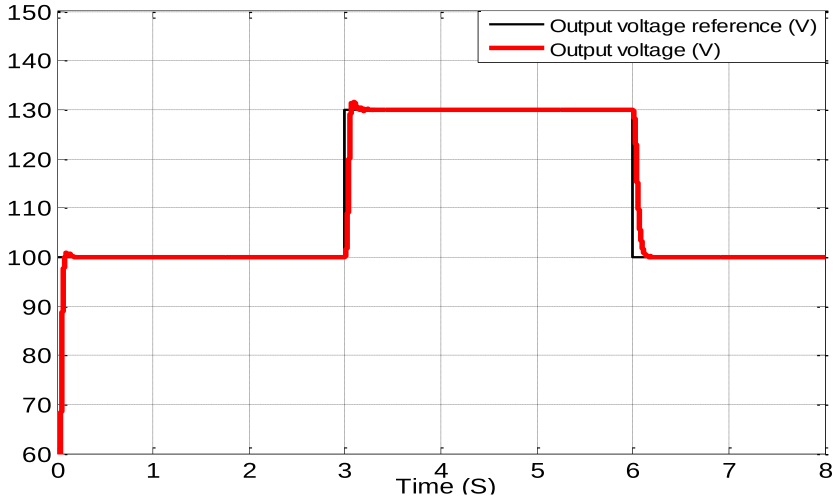


Fig. 2.9 Output voltage variation.

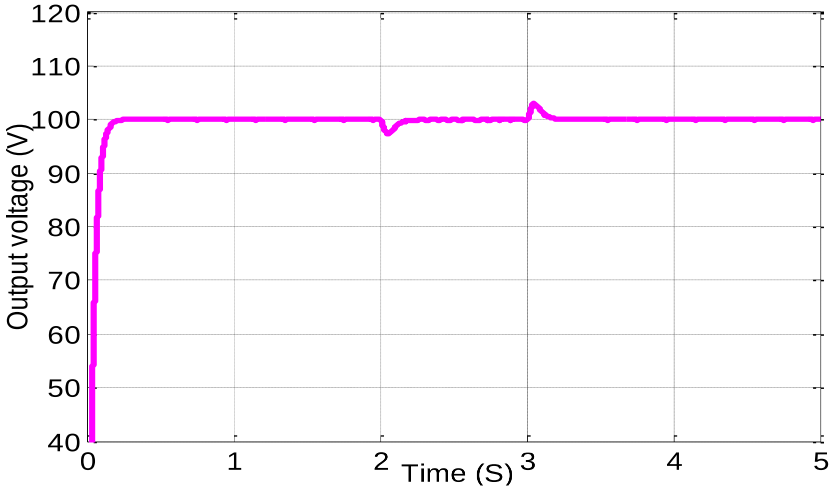


Fig. 2.10 Transient of the step change of R.

2.3.5 Experimental validation

In order to examine the robustness of the proposed method an experimental setup using the dSPACE 1104 (Digital Signal Processor) controller board is given, which shows in Fig. 2.11, in LAS laboratory, Setif-1 University, Algeria. For the real time implementation of the proposed controller, the third-order Bogacki-Shampine solver has been chosen with fixed step at 40e^{-6}. The controller was executed at 20 KHZ. An inverter (SEMIKRON, 20KVA, 1200V, 50A) used as a rectifier. The variation in the load is obtained by connecting or disconnecting parallel load.

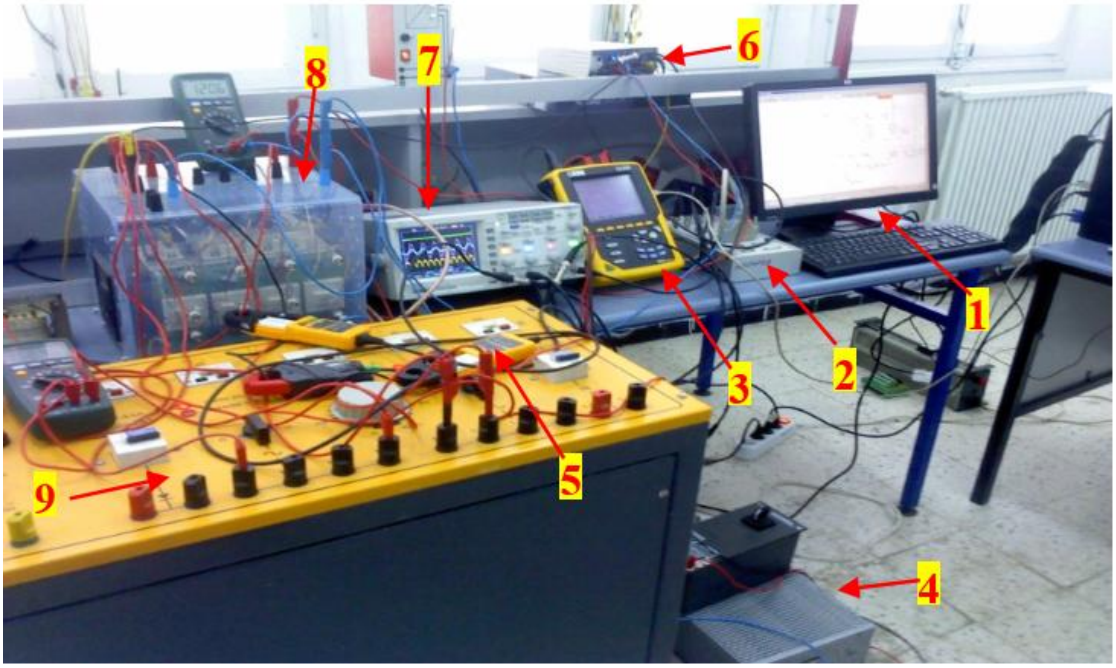


Fig. 2.11 Experimental test bench: (1: PC, 2: dSPACE I/O connectors, 3: Power analyzer, 4: Load, 5: Current sensor, 6: Voltage sensor, 7: Scope, 8: Inverter, 9: Transformer.

Using the developed test bench, various tests where conducted to verify the performance of the proposed method.

Test 1: In this test, no PFC controller was implemented, the experimental results are shown in Fig. 2.12 and 2.13, it is observed that the performance of the system without PFC is not satisfactory since a THD more than 50% is observed in the input current having a low power factor around 0.79, the input current is not in sinusoidal waveform. These results confirm the importance of PFC for the conservation of energy in the power converters.

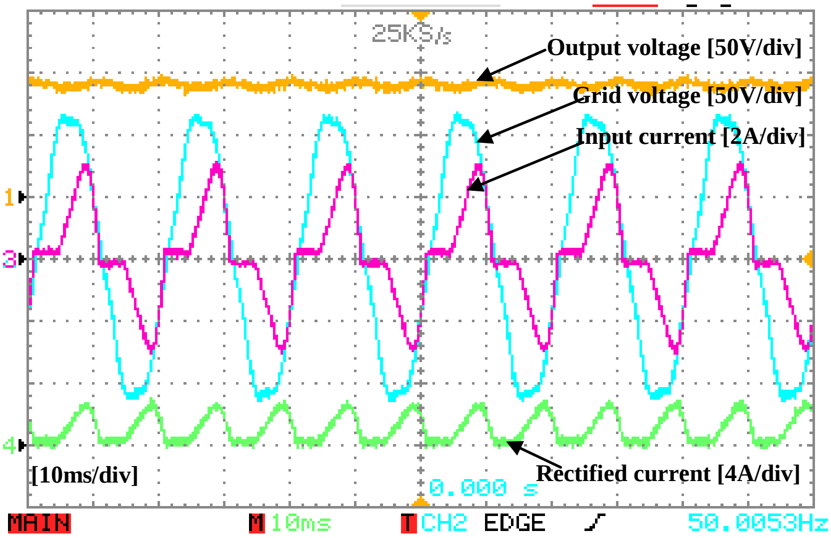


Fig. 2.12 Experimental results in steady state without PFC.

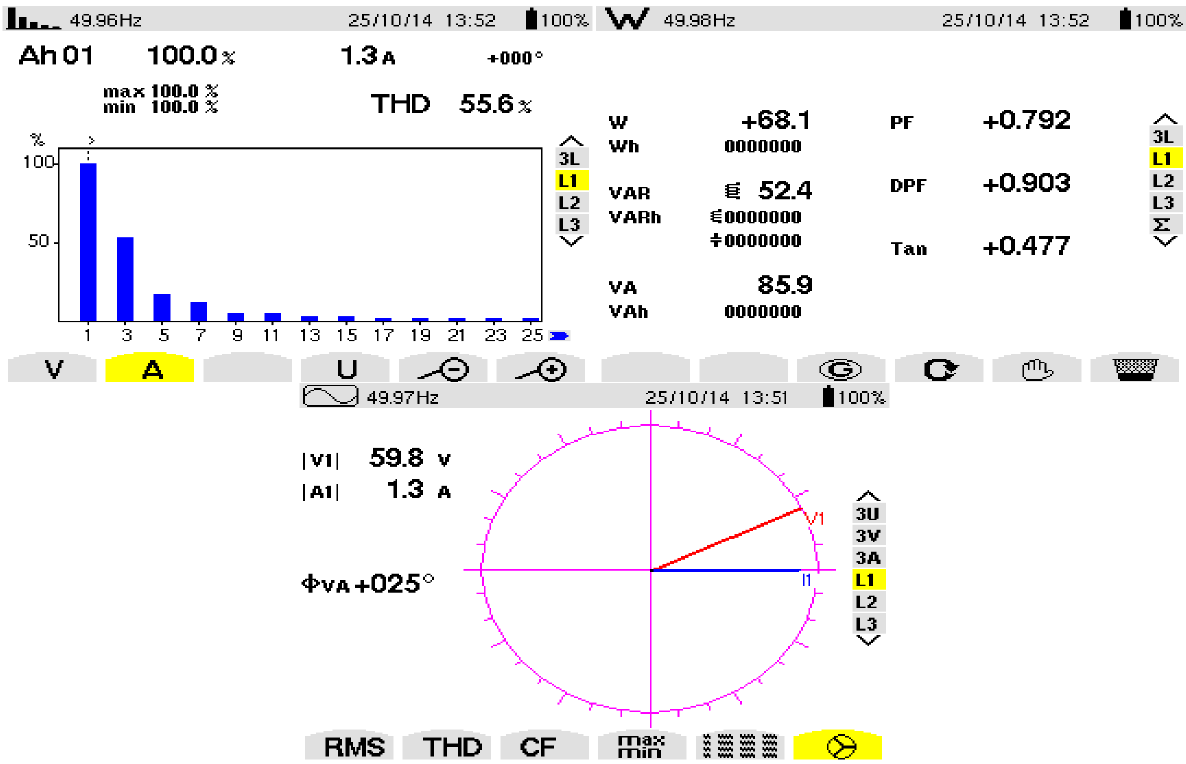


Fig. 2.13 Experimental values in steady state without PFC.

Test 2 - In this test, the PFC act was implemented at nominal load and output voltage (100 Ω, 100V), Figs. 2.14 and 2.15 show the experimental results, which highlight the effect of PFC. Reduction of the THD approximately 4%, and PF is nearly 0.996. Also, the output voltage is still maintained close to its desired reference, and the input current has a sinusoidal shape and in phase with the grid voltage.

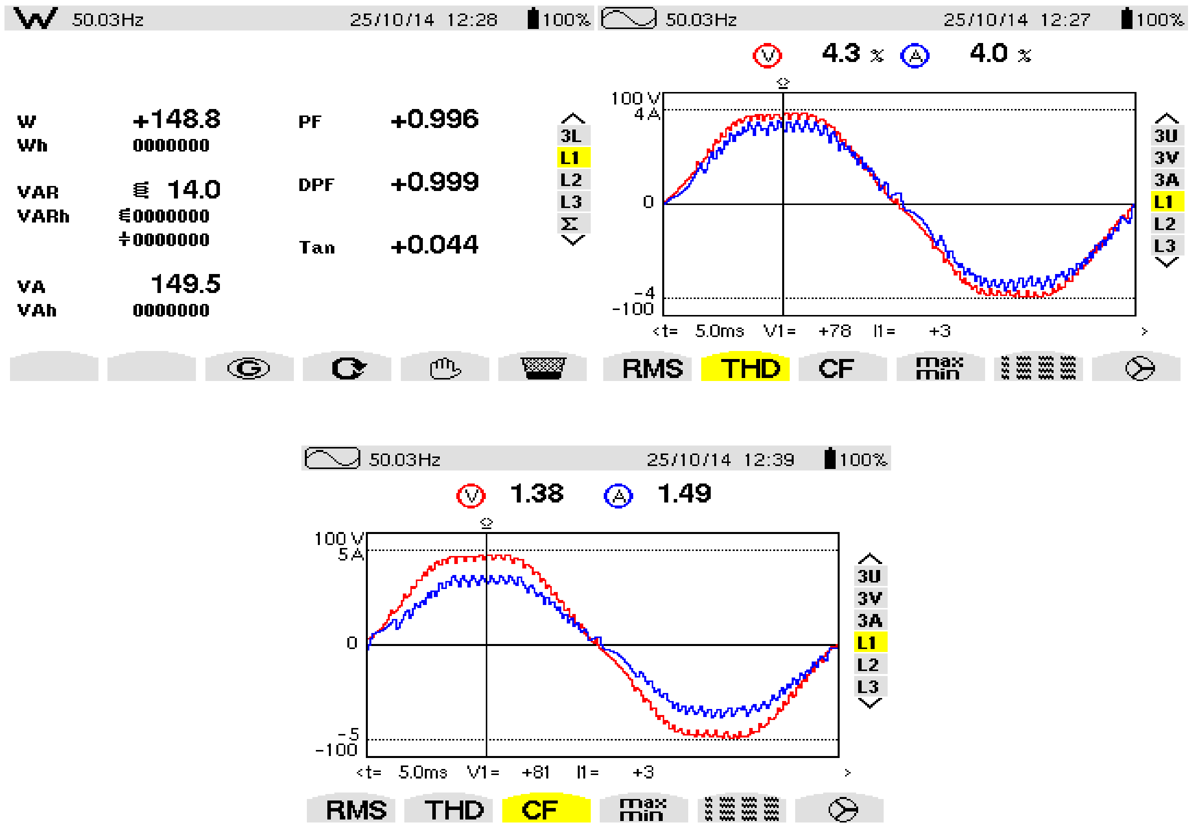


Fig. 2.14 Experimental measurement with PFC: PF, THD, and CF.

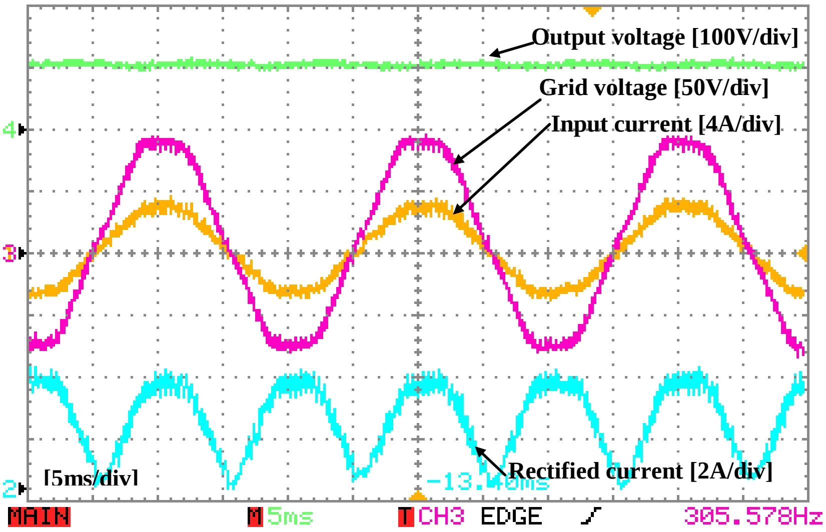
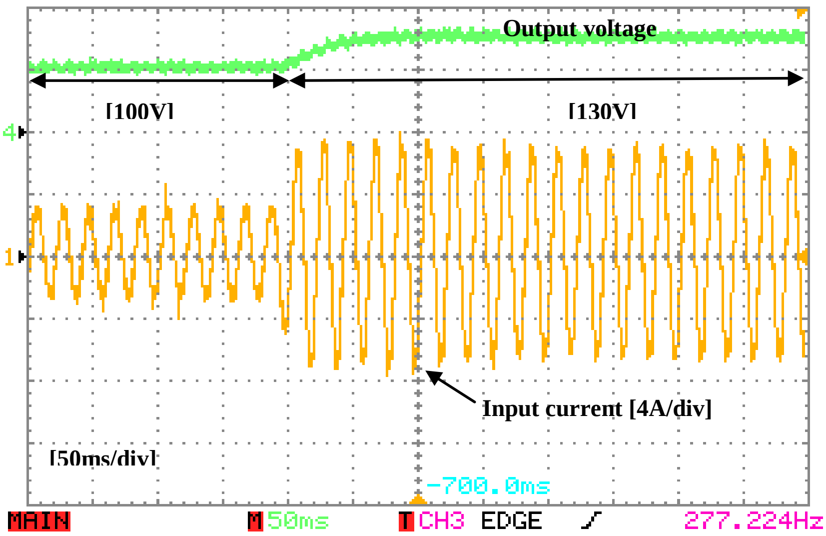
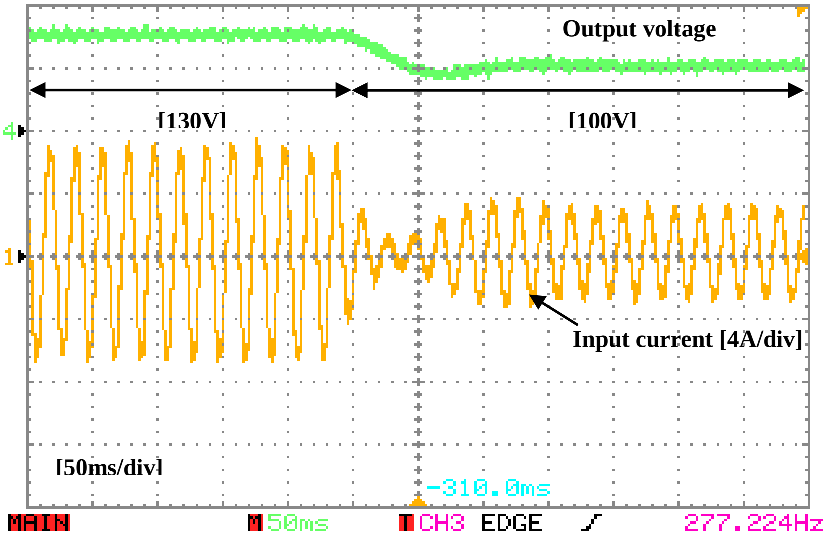


Fig. 2.15 Experimental signal waveforms in the steady state.

Test 3- in this test the reference output voltage is changed from 100V to 130V and vice versa at fixed load of 100Ω. Fig. 2.16 shows the transient input current and output voltage, the new value of the output voltage has been reached after 0.1s and need only 0.075sec to return to previous value again. Also, the input current is sinusoidal with nearly unity power factor.



(a)



(b)

Fig. 2.16 Transient output voltage changes: (a) increasing from 100V to 130V, **(**b) decreasing from 130V to 100V.

Test 4- Here, the circuit was tested under different loads at fixed source voltage of 100V. After reached the steady state a step change in the load from 100Ω to 150Ω and from 150Ω to 100Ω has been taken place. The experimental results are shown in Fig. 2.17. As it may be seen that the output voltage still maintained constant with less overshoot/ undershoot.

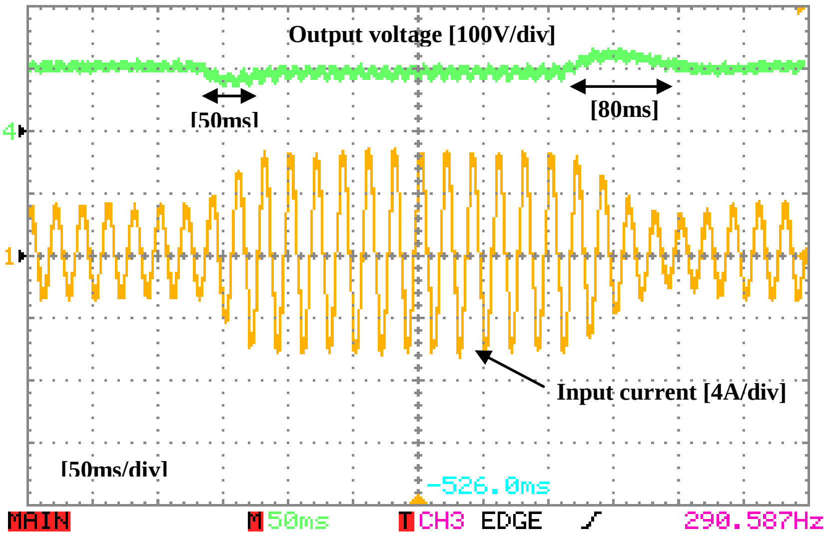


Fig. 2.17 Transient load changes.

2.4 Summary

In this chapter, we gave a short presentation of the control theory for systems both linear and nonlinear. We have also presented the state of the art predictive control for each of these two types of systems in highlighting industrial and academic case. We proposes a simple, low-cost, and powerful control based on finite state model predictive controller (FS-MPC) for a single phase active power factor correction (SAPFC), the proposed control can achieve high performances under different loads and output voltage variations. The method works in discrete time domain with a minimum time delay and capable of achieving a unity power factor in AC current. The proposed controller presents a high performance in the steady and transient states, where it does not get influenced during the parameters fluctuation. The output voltage follows its reference value perfectly, the THD is measured and required the IEEE 519 (THD < 5.0%), the PF is measured as 0.996.

Chapter 3: Higher Order Sliding Mode Control

3.1 Introduction

The Variable Structure Control System (VSCS) is a class of systems where the control law is intentionally changed during the control process according to some well-defined rules that depend on the state of the system. The Sliding Mode Control (SMC) is a particular type of the VSCS, which is characterized by a feedback control laws and switching function, of the current system behaviour and produces as an output the particular feedback controller, it should be used at that instant of time. A Variable Structure System (VSS) is a combination of subsystems, where each subsystem has a fixed control structure and is effective for specified regions of the system behavior. The instances at which the varying of the structure occurs are determined by the current state of the system. Also, its robustness to disturbances and parametric uncertainties make unnecessary a precise knowledge about the system. In addition, the VSC can be implemented by the power electronics already existing in the system. But the stress increase due the chattering. There is a certain difficulty about the VSC design, concerning the definition of a sliding surface with guaranteed proprieties of attractiveness and stability, VSC has been used on various applications of power electronics for regulating the output voltage.

3.2 Sliding mode control

The Sliding Mode Control (SMC) is a nonlinear control method that changes the dynamics of a nonlinear system by application of a high frequency switching control that forces the system to slide along the cross section of the normal system's behavior.

The sliding mode control offers several advantages over the other control methods which are: low sensitivity to plant parameters uncertainty, greatly reduced-order modeling of plant dynamics, finite-time convergence (due to discontinuous control law), stability, robustness, good dynamic response, and simple implementation.

3.2.1 Concept of ideal and real sliding mode

Before introducing different approaches, it is necessary to define two important concepts, the ideal and real sliding mode. Called an ideal sliding mode on a surface all trajectories of the state vector *x*, which is maintained on the sliding surface.

Any movement of the state vector *x* being made in the immediate vicinity of the sliding surface is called real sliding mode. The existence of real sliding mode is due to a technological limit imposed by a non-zero switching time and neglected system dynamics. The ideal sliding mode is an extension of the real sliding mode for a switching time tending to zero.

3.2.2 Fillipov approach

The use of discontinuous control laws in the sliding mode control leads to the study of EDs (differential equations) in discontinuous terms. Different approaches exist for resolving the singularity of differential equations with discontinuous terms. Fillipov [Filippov, 1988] suggests an approach that is used in the sliding mode control, this approach is presented below.

Let us consider the following dynamical system:

$$\dot{x} = f\ (\mathrm{x}, \mathrm{t}) \tag{3.1}$$

Where $x \in R^n$, f is a field vector defined on R^n and t represents the time.

The function f is continuous but has discontinuities on the sliding set $\Sigma = \{x \in R^n \mid \sigma(x, t) = 0\}$. Being defined that the system (3.1) has a variable structure, that is to say, it is defined so that the surface is attractive, and its dynamics can be summarized by the following two structures:

$$\dot{x} = \begin{cases} f^{+}(\mathrm{x}, \mathrm{t}) & \text{if} \quad \mathrm{x} \in \Omega^{+} = \left\{ x \in R^n \middle| \sigma(x) > 0 \right\} \\ f^{-}(\mathrm{x}, \mathrm{t}) & \text{if} \quad \mathrm{x} \in \Omega^{-} = \left\{ x \in R^n \middle| \sigma(x) < 0 \right\} \end{cases} \tag{3.2}$$

The dynamics of the system (3.2) is defined over the state space except the sliding set Σ. By Filippov theorem presented below, if in each of σ(x, t) = 0, the following condition is verified:

$$f_N^+ > 0 \quad and \quad f_N^- < 0 \tag{3.3}$$

Where f_N^+ and f_N^- are projections of $f^+(x,t)$ and $f^-(x,t)$ on the normal sliding set Σ oriented Ω^- towards Ω^+, then the sliding set Σ is attractive for the state vector *x* as each side of the surface, $\dot{x}$ is directed towards the latter. It is then deduced the following relationships:

$$\begin{aligned} f_N^+ > 0 &\Rightarrow \sigma(x,t) > 0 \text{ and } \dot{\sigma}(x,t) < 0 \; where \; \sigma(x,t)\dot{\sigma}(x,t) < 0 \\ f_N^- < 0 &\Rightarrow \sigma(x,t) < 0 \text{ and } \dot{\sigma}(x,t) > 0 \; where \; \sigma(x,t)\dot{\sigma}(x,t) < 0 \end{aligned} \tag{3.4}$$

Inequality $\sigma(x,t).\dot{\sigma}(x,t) < 0$ called *attractiveness* condition to the sliding set Σ. Any control that checks this inequality, regardless to the parametric variations of the model, it is a control that ensuring the convergence of the trajectory of the state vector *x* to the set Σ.

If at each point on the sliding surface condition (3.3) is verified, then according to Filippov's theorem there exists a unique solution for the system (3.1) on the sliding set Σ, which is a linear combination of two structures defined by equation (3.2).

❖ **Filippov Theorem**:

Consider the following system:

$$\dot{x} = f\,(\mathrm{x},\mathrm{t}) \tag{3.5}$$

Where *f* is continuous except on the sliding set Σ, then the dynamics of the system (3.5) on the sliding set Σ is written as follow:

$$\dot{x} = \alpha(\mathrm{t})f_0^+ + (1-\alpha(\mathrm{t}))f_0^- \tag{3.6}$$

Where

$$f_0^+ = \lim_{\sigma \to 0^+} f^+(x,t) \; and \; f_0^- = \lim_{\sigma \to 0^-} f^-(x,t) \tag{3.7}$$

With

$$\alpha(t) = \frac{\langle \nabla\sigma, f^{-}(x,t) \rangle}{\langle \nabla\sigma, f^{-}(x,t) - f^{+}(x,t) \rangle} \tag{3.8}$$

The $\langle a,\ b \rangle$ denotes the scalar product of two vectors a and b.

Note 1. The requirement of attractiveness is not sufficient to satisfy already convergence the system in finite time to the set Σ. In order to avoid an asymptotic convergence, this condition is usually substituting by the condition said η- attractiveness [Slotine, 1991]: $\sigma(x,t).\dot{\sigma}(x,t) < -\eta|\sigma(x,t)|$, where η> 0, and the convergence time is bounded by $\frac{|\sigma(0)|}{\eta}$, where $\sigma(0) = \sigma(x(t = 0))$.

3.2.3 Utkin approach

Utkin proposes to replace the discontinuous control, which ensures the sliding over the surface, by a similar equivalent continuous control (denoted u_{eq}) having the same properties. In the Utkin approach, the discontinuity of the system is not from the discontinuous nature of the system, as Fillipov proposed in his theorem (see previous section), but the control itself. In these works on the concept of equivalent control [Utkin, 1976; Utkin, 1977; Utkin, 1999], Utkin consider the following continuous system:

$$\dot{x} = f(x, u) \tag{3.9}$$

Where $x \in R^n$ unlike to the Fillipov formulation in equation (3.5), f modelled in (3.9) a continuous system with a discontinuous input control $u(x)$, which satisfies attractiveness condition.

The control has the form as follow:

$$u(x) = \begin{cases} u^{+} & \text{if} \quad x \in \Omega^{+} \\ u^{-} & \quad\;\; x \in \Omega^{-} \end{cases} \quad \text{with} \quad u^{+} \neq u^{-} \tag{3.10}$$

Utkin approach in the dynamics of the system (3.9) with control (3.10) can be replaced by an equivalent continuous control u_{eq}, so that the trajectory of the state vector x is maintained on sliding set Σ whose dynamic is written in the following form:

$$\dot{x} = f\left(x, u_{eq}\right) \tag{3.11}$$

Fillipov and Utkin approaches are different because Fillipov considers discontinuous nature of the system, while Utkin consider a system continuously with a discontinuous control. Despite this difference these two approaches meet in the case of input linear systems that consider in [Utkin, 1999; Lopez, 2000].

Consider the following nonlinear system affine:

$$\dot{x} = f\left(x, u\right) = f\left(x\right) + g\left(x\right)u\left(x\right) \tag{3.12}$$

Where $x \in R^n$, *u (x)* is a structure variable control. The functions *f(x)* and *g(x)* are sufficiently differentiable and defined on R^n. To keep the system (3.12) on the switching surface, we applying a control, which forced the trajectory of state vector x to be tangent to $\sigma(x,t)$ that is to say $\dot{\sigma}(x,t) = 0$. The equivalent control u_{eq} is the algebraic solution of the following equation:

$$\frac{d\sigma(x,t)}{dt} = 0 \tag{3.13}$$

Condition (3.13) gives along the trajectory of (3.12) the result follows:

$$\frac{d\sigma(x,t)}{dt} = \nabla\sigma \frac{\delta x(t)}{\delta(t)} = \nabla\sigma\left[f\left(x\right) + g\left(x\right)u_{eq}\right] = 0 \quad with\ \nabla\sigma = \frac{\delta\sigma(x)}{\delta x} \tag{3.14}$$

From the above equation the equivalent control u_{eq} is written as follows:

$$u_{eq} = -\left[\nabla\sigma g\left(x\right)\right]^{-1} \nabla\sigma f\left(x\right) \tag{3.15}$$

Where the existence of $[\nabla\sigma g(x)]^{-1}$ is a necessary condition of the equivalent control existence, it is corresponds to a *transversality* condition presented in the next section.

It is noticed that the equivalent control is defined entirely from the system model (3.12) and the parameters of the sliding surface. The uniqueness of the equivalent control u_{eq} is guaranteed, in the case of a linear input system.

3.2.4 Sira-Ramirez approach

Sira-Ramirez proposes a reformulation of the two previous approaches through the use of differential geometry, which allow completing Fillipov and Utkin approaches and provide a better understanding a core of sliding mode control.

Consider the nonlinear SISO system affine in input (3.12). Defined locally for all $x \in O \subset R^n$, where O is open in R^n whose intersection with the sliding surface is not empty. The functions $f(x)$ and $g(x)$ are two function differentiable on R^n. $\forall x \in O, g(x) \neq 0$. u is an control defined in R, $u: O \to R$. We have $\sigma(x,t)$ continues as $\sigma(x,t): O \to R$. The sliding set is redefined locally in the open O by:

$$\Sigma_O = \{ x \in O \,|\, \sigma(x,t) = 0 \} \tag{3.16}$$

The set Σ_O represents a sub variety of open O of dimension n -1, with $\nabla\sigma(x,t) \neq 0$, at least locally in this open.

Consider that the control u is a variable structure control (3.10) under the condition of attractiveness, then it allows the trajectory of the state vector x of the system (3.12) to converge to the set Σ_O and the system is said an ideal sliding mode.

The system (3.12) is in sliding mode if:

$$\lim_{\sigma \to 0^+} L_{f+gu^+}[\sigma(x,t)] < 0 \;\; and \;\; \lim_{\sigma \to 0^-} L_{f+gu^-}[\sigma(x,t)] > 0 \tag{3.17}$$

Where L is the Lie derivative with $L_{f+gu}[\sigma(x,t)] = \langle \nabla\sigma, f(x) + g(x)u \rangle$.

Condition (3.17) is equivalent to (3.3), this means that in the vicinity of the sliding surface, the system field vector (3.12) is slide to the sliding surface.

3.2.5 Transversality condition

Now let the equivalent control presents in Utkin approach, and existence condition through the use of differential geometry. In ideal sliding mode, the equivalent control u_{eq} checks for the system (3.12), the following condition:

$$\begin{aligned} &\sigma(x)=0 \\ &\dot{\sigma}(x)=\nabla\sigma(f(x)+g(x)u_{eq})=\langle\nabla\sigma, f(x)+g(x)u\rangle=L_{f+gu_{eq}}\left[\sigma(x)\right]=0 \end{aligned} \tag{3.18}$$

The equivalent control it is a solution of:

$$L_{f+gu_{eq}}\left[\sigma(x)\right]=L_f\left[\sigma(x)\right]+u_{eq}(x)L_g\left[\sigma(x)\right]=0 \Rightarrow u_{eq}=-\frac{L_f\left[\sigma(x)\right]}{L_g\left[\sigma(x)\right]} \tag{3.19}$$

So that the equivalent control exists, it is necessary that $Lg[\Sigma(x)]\neq 0$. This condition is called a *transversality* condition, and indicates that for obtaining a state vector slide on the sliding set Σ, it is necessary that the field vectors of *g (x)* is not tangent to the whole Σ.

$$L_g\left[\sigma(x)\right]=\langle\nabla\sigma, g(x)\rangle\neq 0 \tag{3.20}$$

$L_g[\sigma(x)]$ is not only different from zero, it has a constant sign, which depends on the sign of: u^+-u^- (3.10) and the orientation of the sliding surface [Sira-Ramirez, 1988].

Transversality condition can be readily understood from the geometric perspective. However, as the sliding surface is defined by the user, it is always possible to define for checking the transversality condition (3.20).

3.2.6 Relative degree in sliding mode

In this section we give the definition of the relative degree of any system related to a sliding surface and its relation to the existence conditions of a conventional sliding mode.

The notion of relative degree is important in first and high order sliding mode controller. The system below (3.21) has a relative degree r with respect to the function of constraint σ(x) at x_0 if and only if at this point and in its vicinity the following condition is satisfied [Isidori, 1989].

$$\begin{gathered}\forall x \in B_{\epsilon}(x_0),\ \ L_g L_f\left[\sigma(x)\right]=L_g L_f^2\left[\sigma(x)\right]=.....=L_g L_f^{r-2}\left[\sigma(x)\right]=0 \\ and\ \ L_g L_f^{r-1}\left[\sigma(x_0)\right]\neq 0\end{gathered} \tag{3.21}$$

Where $B_{\epsilon}(x_0)$ is a ball of center x_0 and ϵ radius. More simply, the relative degree of a system is the number minimum of σ(x, t) derivations with respect the time to show the control.

The following example shows a practical case for a linear system and surface.

Example: Consider the following linear model and surface:

$$\begin{bmatrix}\dot{x}_1 \\ \dot{x}_2\end{bmatrix}=\begin{bmatrix}0 & 1 \\ a_1 & a_2\end{bmatrix}\begin{bmatrix}x_1 \\ x_2\end{bmatrix}+\begin{bmatrix}0 \\ b\end{bmatrix}u \tag{3.22}$$

$$\sigma(x)=\begin{bmatrix}c_1 & c_2\end{bmatrix}\begin{bmatrix}x_1 \\ x_2\end{bmatrix}\ \ with\ c_i\geq 0 \tag{3.23}$$

Where a_1, a_2 and b ≥0.

The system has a relative degree equal to 1 with respect to the sliding surface σ. Because the first derivative of the sliding surface with respect the time, the control appears explicitly:

$$\dot{\sigma}(x)=\begin{bmatrix}c_1 & c_2\end{bmatrix}\begin{bmatrix}\dot{x}_1 & \dot{x}_2\end{bmatrix}^T=c_1\dot{x}_1+c_2\dot{x}_2=c_1x_1+c_2\left(a_1x_1+a_2x_2+bu\right) \tag{3.24}$$

Where c_2=0, then the previous system has a relative degree equal to 2, because the control appears explicitly in the second derivative of the sliding surface σ.

3.2.7 Components of sliding mode control

The equivalent control is designed to maintain the trajectory of the state vector on the surface when it is already there. It is useful to emphasize that cannot converging the state vector from its initial position towards the sliding surface. Moreover, the corresponding control is an inversion of the system model. Therefore, this control is valid only if the model used to determine accurately describes the physical system, which is impossible in the case of an uncertain system model. For this, it is customary to associate an equivalent control, discontinuous control high frequency subsequently noted Δu. The control Δu in first hand allows the convergence of the state vector to the sliding surface and compensates by the other hand the uncertainties external disturbances where are not taken into account by the model used for synthesis the equivalent control.

The control strategy described above is written as follows:

$$u = u_{eq} + \Delta u \tag{3.25}$$

Δu can take different forms [Harashima, 1986; Lopez, 2000], the most simple is:

$$\Delta u = -K sign(\sigma(x)) \tag{3.26}$$

Where $K > 0$. For a system $\dot{x} = f(x, t, u) + \mu$, where μ represents an uncertainty bounded recovery by M , just as $K > M$ for a convergence of the system in finite time to the sliding set Σ [Lopez, 2000].

The equivalent control u_{eq} can be seen also as a way in the case of an uncertain system, minimizing the high frequency component of a conventional sliding mode control represented by Δu, since the major drawbacks of this later are described below. So that the control (3.25) or (3.26) holds the trajectory of the state vector on Σ, the discontinuous control must continually switch at infinite frequency between u^+ and u^-. The switching control is called *chattering*.

The chattering is a major drawback in conventional sliding mode control, which shown in Fig.3.1, because it is excite the high components frequencies of the system that are likely to destabilize the system and reduce a life time of many actuators.

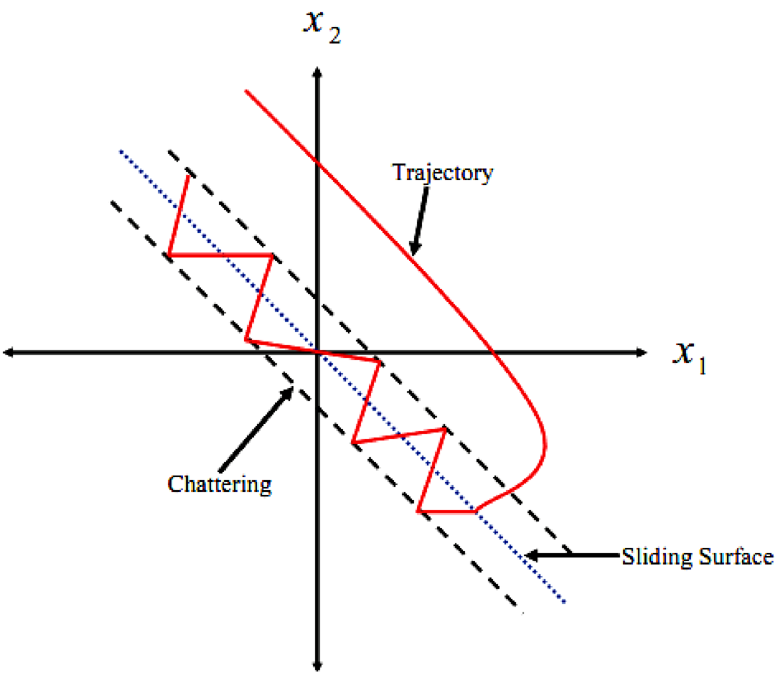


Fig. 3.1 Chattering effect.

There are many approaches have been proposed to reduce or eliminate the chattering [Slotine, 1986; Harashima, 1986; Chettouh, 2008; Asada, 1986]. The principle is the same as chattering occurs in the vicinity of the sliding surface, the idea is to approach in this region the discontinuous control Δu by a continues control. The drawback of these approaches is that there is a degradation of the robustness [Utkin, 1999]; another approach to reduce the chattering and improve the convergence precision to the set Σ is the use of higher order sliding mode control. This family of controls is a generalization of the control (3.26), which is called classical sliding mode control.

3.3 Higher order sliding mode control

In the standard SMC, $\dot{\sigma}$, is discontinuous; this is the main reason why high frequency switching appears in the output signal (chattering effect), which causes problems in practical application. In order to avoid chattering, in this section a high order sliding mode control (HOSMC) is used [Levant, 1998; Levant, 2003]. HOSMC acts on the higher order time derivative of the system, instead of influencing the first derivative as it happens in

SMC [Levant, 2003]. Its principal characteristic is that it keeps the main advantages of the SMC, and removing the chattering effects. The sliding order is a number of continuous total derivatives of σ in the vicinity of the sliding mode. It fixes the dynamics smoothness degree.

The *r-th* order sliding mode (*r*-sliding) is determined by the equation:

$$\Sigma_r = \left\{ x \in 0 \middle| \sigma(x) = \dot{\sigma}(x) = \ldots = \sigma^{[r-1]}(x) = 0 \right\} \tag{3.27}$$

The major drawback of a first order sliding mode control is the chattering [Sira-Ramirez, 1988; Levant, 2000]. We present in the following a generalization of this type of control introduced in the eighties [Emelyanov, 1986], which allows to reduce chattering, improve the control precision of convergence for real sliding mode and will lift the restriction on the relative degree.

In this section, we present higher order sliding mode control, its relevance and its practical limitations compared to a first order sliding mode control. We are particularly interested for second order sliding mode controls.

Consider the non-linear SISO model affine in input as:

$$\begin{aligned} \dot{x} &= f(x) + g(x)u \\ \sigma &= \sigma(x) \end{aligned} \tag{3.28}$$

Where $x \in O \subset R^n$, O is open in R^n, $u \in R$ is a discontinuous function bounded, $\sigma(x)$, $f(x)$ and $g(x)$ are uncertain function sufficiently differentiable and $\sigma(x) \in R$. The purpose of the control is to force the trajectories of the system to evolve, in finite time, in the sliding set Σ_r (3.27). So, as r–1 is the first derivative of $\sigma(x)$ with respect to time are continuous. This mean that is not depend directly on the control u. A control u is a *r-th* order sliding mode control for a relative degree r with respect to σ, is as follows:

$$\frac{\delta\sigma(x)}{\delta u} = \frac{\delta\dot{\sigma}(x)}{\delta u} = \ldots\ldots = \frac{\delta\sigma^{[r-1]}(x)}{\delta u} = 0 \quad \mathit{and} \quad \frac{\delta\sigma^{[r]}(x)}{\delta u} \neq 0 \tag{3.29}$$

It is necessary for the following, to complete the above definitions by definitions of the concept of reducing the system provides the *r-th* order sliding mode control and the definition of *r-th* order real sliding mode. These definitions are derived from the work of Emelyanov and Levant [Emelyanov, 1986; Levant, 1993]

❖ **Definition.**

Consider $\sigma, \dot{\sigma}, \ldots \ldots \sigma^{[r-2]}$ are differentiable functions with respect to *x* and locally:

$$rang\left[\nabla\sigma,\ \nabla\dot{\sigma},\ldots.,\nabla\sigma^{[r-2]}\right]=r-1 \tag{3.30}$$

Equation (3.30) is called: *weak regularity condition*. If the set Σ_r is a differentiable manifold and if for all i = 1, · · ·,*r* -1 , the σ_i are regular varieties, while the weak regularity condition (3.30) can be extended to $\sigma^{[r-1]}$ (3.31).

$$rang\left[\nabla\sigma,\ \nabla\dot{\sigma},\ldots.,\nabla\sigma^{[r-2]}\right]=r \tag{3.31}$$

Equation (3.31) is called: *strong regularity condition*. If the regularity condition is satisfied, the *r-th* order sliding mode (3.27) establishes a dimension constraint *r* and dynamics resulting is reduced to the order *n-r*, where *n* is the order of the system.

It is useful to note that in the case of the conventional sliding mode (3.26), the order of the system changes from *n* to *n*-1.

3.3.1 Second order sliding mode control

We are interested now to second order sliding mode control. It is useful to notice that these controls are applied to the single phase power factor correction (this choice is presented in Sections 3.4). The goal is that the control *u* forced the trajectories of the system (3.28) to evolve in finite time on the sliding set Σ_2 (3.27). The Fillipov trajectories eligible is included in the tangent space to Σ_2. We deal in the following the system (3.28) only in the case of a relative degree equal to 1 and 2 with respect to the sliding surface σ. In order to establish the existence of second order sliding mode control algorithms, it is necessary that the following assumptions are checked [Perruquetti, 2002; Kunusch, 2009].

H.1 The control values belong to the closed set $U = \{u : |u| \leq U_m\}$, where U_m is a real constant.

H.2 There exists $u \in (0, 1)$ such that for any continuous function u(t) with $|u(t)| > u_1$, there is t_1, such that $\sigma(t)u(t) > 0$ for each $t > t_1$. Hence, the control $u(t) = -sign(\sigma(t_1))$, where t_0 is the initial value of time, provides hitting the surface in finite time.

H.3 Given $\dot{\sigma}$, the total time derivate of the sliding variable σ, there are positive constants σ_0, $u_0 < 1$, Γ_m, Γ_M, such that if $|\sigma| < \sigma_0$ then

$$0 < \Gamma_m \leq \frac{\delta}{\delta u}\dot{\sigma}(x,t,u) \leq \Gamma_M \qquad \forall u \in U,\, x \in X \tag{3.32}$$

H.4 In addition, a positive constant Φ has been computed, such that within the region $|\sigma| < \sigma_0$, the following inequality holds $\forall\, t, x \in X, u \in U$:

$$\left|\frac{\delta}{\delta t}\dot{\sigma}(t,x,u) + \frac{\delta}{\delta t}\dot{\sigma}(t,x,u).\big(f(x) + g(x,u)\big)\right| \leq \Phi \tag{3.33}$$

Therefore, the stabilization problem of the system (3.1) with input-output dynamics can be solved through the solutions of the following equivalent differential inclusion by applying SOSM:

$$\ddot{\sigma} \in [-\Phi, \Phi] + [\Gamma_m, \Gamma_M]\dot{u} \tag{3.34}$$

The hypothesis **H.2** establishes that from any point of the state space, it is possible to define a bounded control converging the state vector trajectory x in a finite time, towards linearity band. Assumptions **H.3** and **H.4** implies that $\ddot{\sigma}\,(x)$ is bounded for a fixed control u. According to the implicit function theorem, there is a continuous control u_{eq} that satisfied $\dot{\sigma} = 0$ [Utkin, 1992]. This control can be treated as equivalent control ensuring the surface invariance $\sigma(\mathrm{x}) = \dot{\sigma}(x) = 0$.

In the case of a second order sliding mode control, the control appears explicitly on $\ddot{\sigma}(x)$ *i.e*

$$\frac{\delta\sigma(x)}{\delta u}=\frac{\delta\dot{\sigma}(x)}{\delta u}=0 \quad and \quad \frac{\delta\ddot{\sigma}(x)}{\delta u}\neq 0 \tag{3.35}$$

In this case the assumptions described above can be applied

3.3.1.1 Twisting algorithm

The Twisting algorithm is proposed by Levant [Levantovsky, 1985; Emelyanov, 1986a; Emelyanov, 1986b]. It was developed for systems with relative degree equal to 1 and 2. The principle of this algorithm is simple, it is a integrating a function $sign(\sigma)$ alternating between two values $\alpha_m \; sign(\sigma)$ and $\alpha_M \; sign(\sigma)$. This alternation converges the state vector x to the singular point $(\sigma, \dot{\sigma}) = (0,0)$ with a geometric progression at each trajectory intersection of the state vector x in the state space with the axes σ and $\dot{\sigma}$. In the case of a system has a relative degree equal to 1, the algorithm can be written as form (3.36). In the case of a system has a relative degree equal to 2, the algorithm can be written as form (3.37).

$$\Delta\dot{u}(x)=\begin{cases} -\Delta u & if \;\; |u|>u_m \\ \alpha_m sign(\sigma) & if \;\; \sigma\dot{\sigma}\leq 0 \;\; |u|\leq u_m \\ \alpha_M sign(\sigma) & if \;\; \sigma\dot{\sigma}>0 \;\; |u|\leq u_m \end{cases} \tag{3.36}$$

$$\Delta\dot{u}(x)=\begin{cases} \alpha_m sign(\sigma) & if \;\; \sigma\dot{\sigma}\leq 0 \\ \alpha_M sign(\sigma) & if \;\; \sigma\dot{\sigma}>0 \end{cases} \tag{3.37}$$

3.3.1.2 Super-twisting algorithm

This algorithm has been developed for the control of systems has a relative degree equal to 1. Therefore, the discontinuity of the control acts on the first derivative of the surface. This control law was proposed by Emelyanov in [Emelyanov, 1986c] and studied by Levant in [Levant, 1993]. The Super-twisting does not use the information about the surface derivative, which is seen as an advantage. The super-twisting algorithm consists of two parts: a discontinuous part u_1 and a continuous part u_2. See the following equations:

$$u(t) = u_1(t) + u_2(t) \tag{3.38}$$

Where

$$\dot{u}_1(t) = \begin{cases} -u & if \ |u| > 1 \\ -w \ sign(\sigma) & if \ |u| \le 1 \end{cases} \tag{3.39}$$

$$u_2(t) = \begin{cases} -\lambda|\sigma_0|^{\rho} \, sign(S) & if \quad |\sigma| > \sigma_0 \\ -\lambda|\sigma|^{\rho} \, sign(S) & if \quad |\sigma| \le \sigma_0 \end{cases} \tag{3.40}$$

Where σ_0 is a boundary layer around the sliding surface.

Sufficient conditions of Super-twisting convergence are given by the following conditions [Levant, 1993]:

$$\begin{cases} w > \dfrac{\Phi}{\Gamma_m} \\ \lambda^2 \ge \dfrac{4\Phi}{\Gamma_m^2} \dfrac{\Gamma_M (w+\Phi)}{\Gamma_m (w-\Phi)} \\ 0 < \rho \le 0.5 \end{cases} \tag{3.41}$$

The Super-twisting exponentially stable for ρ = 1. The choice of ρ=0.5 ensures optimum speed of convergence to the set Σ_2 [Levant, 1998]. In [Levant, 2000] a hybrid between Twisting and Super-twisting is proposed for the control of a Aircraft.

3.3.2 Disadvantages of higher order sliding mode control

Although we have shown in the previous section the value of using higher order sliding mode control. However, it is necessary to emphasize the practical limits to the increase of the order of the control leads us to be satisfied with the second order sliding mode control (Twisting and Super-twisting). A control of *r-th* order requires to obtain, for each switch, the information on $r - 1$ derivative of σ. Although it is possible to reduce this information in the case of a real sliding mode to $r - 2$ derivatives of σ. For a control of *r-th* order sliding mode, the discontinuous control is applied to $\sigma^{[r]}$ this has the advantage of obtaining a continuous control of r-1 integrators. However, these integrators reduce the high-frequency components of the control necessary to compensate model uncertainties [Braikia, 2011]. This chain of integrators also like disadvantage of slowing the control, which results in a reduction of the performance of higher order sliding mode controls, and robustness. Finally, the use of second order sliding mode control to reduce the chattering and improve the accuracy of convergence in finite time of single phase power factor correction is presented below.

3.4 Description of the proposed control method for Power Factor Corrector (PFC)

The main contribution of this chapter is to introduce a novel control approach for PFC based on High order sliding mode control and an optimal predictive current control law and its experimental validation around dSPACE 1104. Further, the objectives are to achieve unity power factor, low THD, minimal digital hardware, and robustness guaranteed for different output voltage and load fluctuation. Fig. 3.2 illustrates the general scheme of the circuit with its control loops.

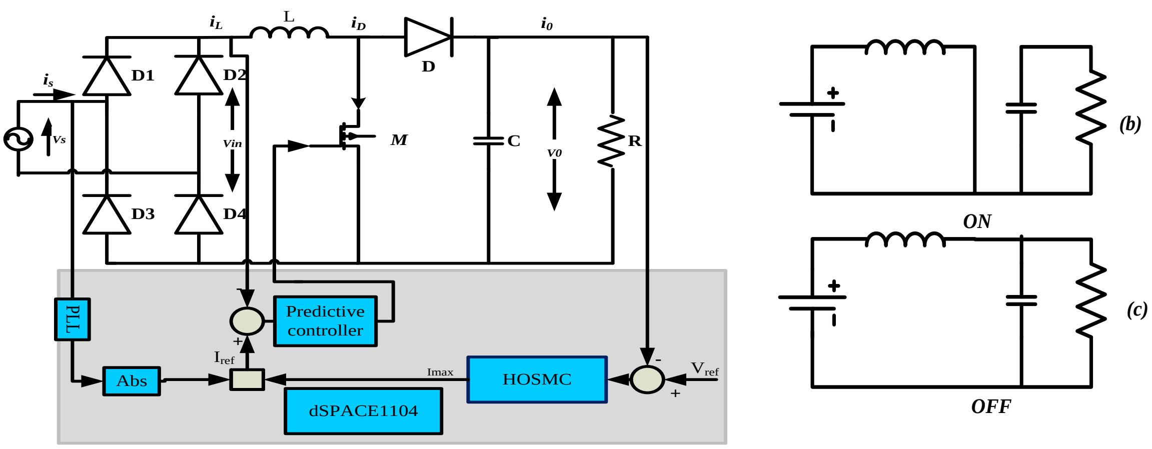


Fig. 3.2 PFC pre-regulator with HOSMC.

3.4.1 Voltage loop controller

To regulate the output voltage, the high order sliding mode controller (a second) based on super twisting algorithm has been used which has a specific operating mode. A sliding mode is said ``*r*th order sliding mode'' if $S(t) = \dot{S}(t) = \cdots = S^{(r-1)} = 0$. In HOSMC, the idea is to force the error to move on the switching surfaces and to keep its (r-1) first successive derivatives null. More specifically, SOSMC aims for $S(t) = \dot{S}(t) = 0$, which is, the controller's aim to steer to zero at the intersection of S(t) and $\dot{S}(t)$ in the state space [Eker, 2010]. The main feature of the proposed method is that the system presents a high robustness performance during the change of the parameters. The control is carried out based on the state variables used to build a switching surface, whose purpose is to force the dynamic system to follow this switching surface in finite time and reduce chattering effects.

Let’s the following tracking error as:

$$e(t) = V_{ref} - V_0 \tag{3.42}$$

The aim of control is that;

$$\lim_{t\to\infty}\|e(t)\| = \lim_{t\to\infty}\| V_{ref}(t) - V_0(t) \| = 0 \tag{3.43}$$

Then we will have:

$$\frac{de(t)}{dt} = \frac{dV_{ref}(t)}{dt} - \frac{dV_0(t)}{dt} = \frac{dV_{ref}(t)}{dt} - \frac{1}{C}\left(i_L - \frac{V_0}{R}\right) \tag{3.44}$$

Thus we have

$$\frac{d^2e(t)}{dt^2} = \frac{d^2V_{ref}(t)}{dt^2} - \frac{d^2V_0(t)}{dt^2} = \frac{d^2V_{ref}(t)}{dt^2} - \frac{1}{C}\left(\frac{di_L}{dt} - \frac{1}{R}\frac{dV_0}{dt}\right) \tag{3.45}$$

Using (3.41) and (3.45), the second derivative rewrite as follow:

$$\frac{d^2e(t)}{dt^2} = \frac{d^2V_{ref}(t)}{dt^2} - \frac{1}{C}\left(\frac{1}{L}(V_{in} - V_0) - \frac{1}{RC}\left(i_L - \frac{V_0}{R}\right)\right) \tag{3.46}$$

Let us consider the following surface [Manceur, 2012].

$$\sigma(e,t) = \left(\frac{\partial}{\partial t} + \lambda\right)^{(n-1)} e\,(t) \tag{3.47}$$

Where λ is a positive constant. It is clear that the degree related to the voltage in (3.41) is equal to 1; in this case the surface is chosen as the setting error. Therefore, the surface can be writing as:

$$\sigma(t) = V_{ref} - V_0 \tag{3.48}$$

To maintain the trajectory of the output voltage (V_0) in switching surface, an equivalent current (I_{eq}) control is applied considering the following invariance conditions:

$$\begin{cases} \sigma(t) = V_{ref} - V_0 = 0 \\ \dot{\sigma}(t) = \dfrac{dV_{ref}(t)}{dt} - \dfrac{dV_0(t)}{dt} = 0 \end{cases} \tag{3.49}$$

To prove the stability and ensure the stable convergence property of the proposed controller, the Lyapunov function is selected as

$$V(t) = \frac{1}{2}\sigma^2(t) + \frac{1}{2}\dot{\sigma}^2(t) \tag{3.50}$$

Where V(0)=0, V(t)>0 for σ(t) ≠ 0 and $\dot{\sigma}(t) \neq 0$

The stability is ensured if the derivative of the Lyapunov function is negative, and satisfies the following condition:

$$\dot{V}(t) < 0, \quad \sigma(t) \neq 0, \quad \dot{\sigma}(t) \neq 0 \tag{3.51}$$

Taking the first derivative of (3.50) yields

$$\begin{aligned} \dot{V}(t) &= \sigma(t)\,\dot{\sigma}(t) + \dot{\sigma}(t)\,\ddot{\sigma}(t) \\ &= \dot{\sigma}(t).[\sigma(t) + \ddot{\sigma}(t)] \end{aligned} \tag{3.52}$$

Using (3.42), and (3.46) yields

$$\begin{aligned} \dot{V}(t) &= \dot{S}(t)\left[V_{ref} - V_0 + \ddot{V}_{ref} - \frac{1}{C}\left(\frac{1}{L}(V_{in} - V_0) - \frac{1}{RC}\left(i_L - \frac{V_0}{R}\right)\right)\right] \\ &= \left[\dot{V}_{ref} - \frac{1}{C}\left(i_L - \frac{V_0}{R}\right)\right].\left[V_{ref} - V_0 + \ddot{V}_{ref} - \frac{1}{C}\left(\frac{1}{L}(V_{in} - V_0) - \frac{1}{RC}\left(i_L - \frac{V_0}{R}\right)\right)\right] \end{aligned} \tag{3.53}$$

For our case, V_{ref} is constant and its derivative is null, (3.53) can be represented as

$$\dot{V}(t) = -\frac{1}{C}\left(i_L - \frac{V_0}{R}\right)\left[V_{ref} - V_0 - \frac{1}{C}\left(\frac{1}{L}(V_{in} - V_0) - \frac{1}{RC}\left(i_L - \frac{V_0}{R}\right)\right)\right] < 0 \tag{3.54}$$

When the system is in the reaching phase with σ(t) ≠ 0, $\dot{\sigma}(t)$ and have V(t) is negative definite. From the above analysis, the derivative of the Lyapunov function is a negative definite, the system is globally asymptotically stable.

The global control is composed of the equivalent control (I_{eq}) and the super twisting algorithm terms. Now, lets us the super twisting algorithm, this later is used in order to stabilize the system, avoid chattering effects, and converges the system to the desired trajectory in finite time. The advantage of super twisting algorithm is that; it does not need any information on the time derivative of the sliding variable and maintains all the distinctive robust features of the SMC. The control law is composed by two parts defined by the following control law:

$$u(t) = u_1(t) + u_2(t) \tag{3.55}$$

Where

$$\dot{u}_1(t) = \begin{cases} -u & if\ |u| > 1 \\ -w\ sign(\sigma) & if\ |u| \le 1 \end{cases} \tag{3.56}$$

$$u_2(t) = \begin{cases} -\lambda|\sigma_0|^{\rho}\, sign(\sigma) & if\quad |\sigma| > \sigma_0 \\ -\lambda|\sigma|^{\rho}\, sign(\sigma) & if\quad |\sigma| \le \sigma_0 \end{cases} \tag{3.57}$$

Where σ_0 is a boundary layer around the sliding surface.

By adding the super twisting algorithm terms, we obtain the global control of voltage loop as follow:

$$I_{Max} = I_{eq} + \int_0^t \dot{u}_1 dt + u_2 \tag{3.58}$$

3.4.2 Current loop controller

Due to predictive control's suitable performance and flexible implementation on real time, it has been used to control the power converters for last few years [Rodriguez, 2012]. In this work, the predictive method has been used to control converter switching. Taking into account the state of the switch from the circuit shown in Fig. 3.2, the equations of the inductor current $i_L(t)$ for each state can be expressed as [Bibian, 2001; Azazi, 2014]:

When M is turned on:

$$L\frac{d\,i_L}{dt} = V_{in}(t) \qquad \text{for} \qquad t(k) \le t < t(k) + d(k).T_s \tag{3.59}$$

When M is turned off:

$$L\frac{d\,i_L}{dt}=V_{in}(t)-V_0(t) \quad \text{for} \quad t(k)+d(k).T_s \le t < t(k+1) \tag{3.60}$$

Where $V_{in}(t)$ and $V_0(t)$ are the input and output voltage respectively, t(k) and t(k+1) are the started time of the k^{th} and $(k+1)^{th}$ switching cycle respectively, d(k) and T_s are the duty cycle and switching period respectively. Since the switching frequency is more than the line frequency, (3.59) and (3.60) can be rewritten as

$$L\frac{i_L(t(k)+d(k).T_s)-i_L(t(k))}{d(k).T_s}=V_{in}(t(k)) \tag{3.61}$$

$$L\frac{i_L(t(k+1))-i_L(t(k)+d(k).T_s)}{(1-d(k)).T_s}=V_{in}(t(k))-V_0(t(k)) \tag{3.62}$$

The diagram presented in Fig. 3.3, represents the inductor current during one switching cycle [Azazi, 2014].

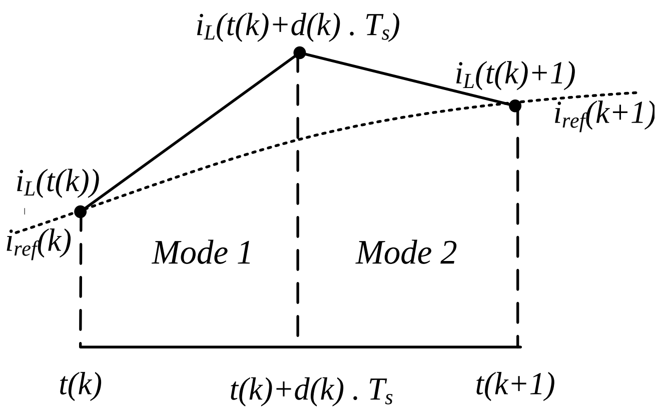


Fig. 3.3 Inductor current during one switching cycle.

At instant $t(k)+d(k)\,.\,T_s$, the inductor current can be derived from (3.61) as

$$i_L(t(k)+d(k)\cdot T_s)=i_L(t(k))+\frac{1}{L}\cdot V_{in}(t(k))\cdot d(k)\cdot T_s \tag{3.63}$$

In the start time of switching cycle t(k+1), the inductor current can be derived from (3.62) as follow:

$$i_L(t(k+1))=i_L(t(k)+d(k)\cdot T_s)+\frac{1}{L}\cdot(V_{in}(t(k))-V_0(t(k)))\cdot(1-d(k))\cdot T_s \tag{3.64}$$

Substituting (3.63) and (3.64), the inductor current can be written as

$$i_L(t(k+1)) = i_L(t(k)) + \frac{1}{L}\cdot(V_{in}(t(k))\cdot T_s - \frac{1}{L}\cdot V_0(t(k))\cdot(1-d(k))\cdot T_s \tag{3.65}$$

The discrete form of (3.65) can be expressed as

$$i_L(k+1) = i_L(k) + \frac{V_{in}(k)\cdot T_s}{L} - \frac{V_0(k)\cdot(1-d(k))\cdot T_s}{L} \tag{3.66}$$

From (3.66), the inductor current of the next switching cycles is calculated by the inductor current from the present switching cycle, input voltage, output voltage and duty cycle. For calculating the duty cycle d(k), (3.66) can be rewritten as

$$d(k) = \frac{L}{T_s}\frac{i_L(k+1) - i_L(k)}{V_0} + \frac{V_0(k) - V_{in}(k)}{V_0} \tag{3.67}$$

Through the boost parameters such as input voltage, output voltage and inductor current, the duty cycle d(k) for the actual switching cycle have been calculated. For the designed boost converter with PFC, the inductor current i_L(k+1) has been forced to follow the reference i_{ref} (k+1), which has a rectified sinusoidal form.

Substituting V_0, i_L(k+1) in (3.67) by its references V_{ref} and i_{ref}(k+1) respectively, the duty cycle can be expressed as [Azazi, 2014]:

$$d(k) = \frac{L}{T_s}\cdot\frac{i_{ref}(k+1) - i_L(k)}{V_{ref}} + \frac{V_{ref} - V_{in}(k)}{V_{ref}} \tag{3.68}$$

The reference current i_{ref} in (2.69) is calculated as follow:

$$i_{ref}(k+1) = I_{max}\,.|\sin(\omega_{line}\,.t(k+1)| \tag{3.69}$$

Where I_{max} is the peak value of the reference current, which is given by the output of the voltage loop controller.

All steps of the proposed method are shown in Fig. 3.4.

Step 1: Identification of the components parameters.

Step 2: Determination of suitable voltage and introduce the HOSMC to regulate it.

Step 3: Determination of I_{Max} through HOSMC, after that introduce the predictive control to regulate the current.

Step 4: Verify that all of duty cycles are calculated by the predictive control.

Step 5: Exploitation of the converter.

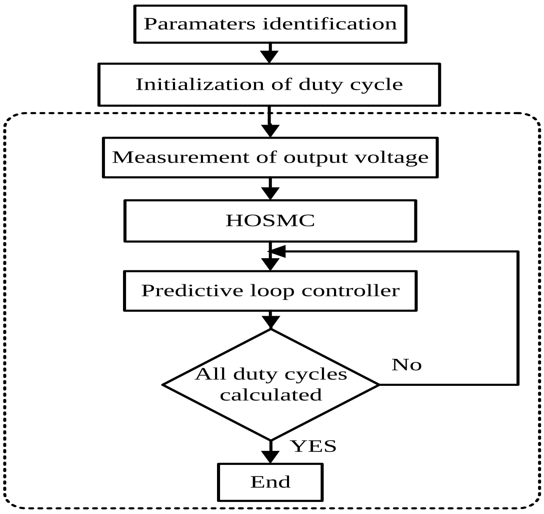


Fig. 3.4 Flowchart of proposed method.

3.4.3 Simulation results

The aim of the simulation is to improve the performance of the power factor correction boost converter using HOSMC and predictive controllers to reduce the harmonic distortion produced by the nonlinear load and to achieve a unity power factor under parameters variation. The parameters of the system are presented in appendix. Due to the limitation of the paper presentation, only the performance of the proposed method is presented in simulation section. The performance of PI has been presented in experimental section.

Fig. 3.5 illustrates the source voltage, the output voltage, the source current simulated waveforms, for the proposed method at nominal load and nominal source voltage (150V, 100Ω) in the steady state. From this figure, it can be seen that the obtained results with the proposed control are satisfactory and require the international norms. The input current is in sinusoidal form and in phase with the source voltage. The total harmonic distortion (THD) is found less than 4%, and the output voltage is maintained constant at desired value for a steady-state error of 1V.

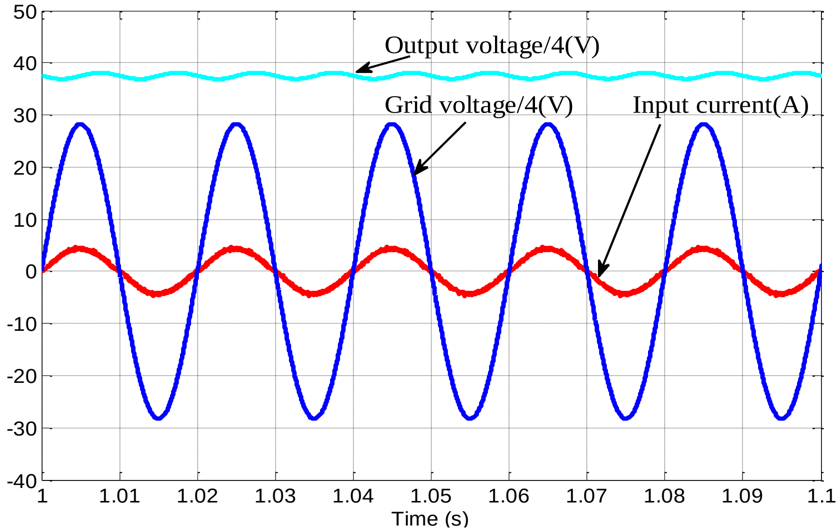


Fig. 3.5 Signal waveforms in the steady state.

In order to verify the robustness of the proposed method, the influence of the parameters changes (load resistor, reference output voltage) during steady and transient states based on the performance of the output voltage are studied by considering two cases.

1. Variation of ±50% on load resistor.

2. Variation of ±21% on reference output voltage.

Fig. 3.6 illustrates the transient response during the step change of R by keeping the reference output voltage fixed at 130V, the resistor load has been decreased 50% at t =2s, and has been increased 50% at t = 4s, from 100Ω to 50Ω and from 50Ω to 100Ω. After a short transient, the output voltage is maintained constant at its reference value during the load resistor variations, Also, from the figure an overshoot around 6V (4.6%) with time response around 0.18s in increasing load, and an overshoot around 7V (5.38%) with response time around 0.16s in decreasing load are observed.

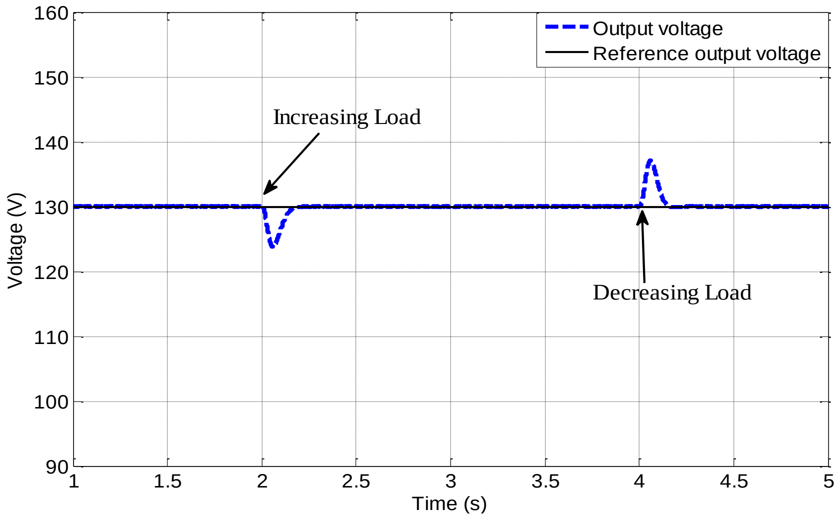


Fig. 3.6 Transient of the step change of R, from 100Ω to 50Ω, and from 50Ω to 100Ω.

Fig. 3.7 illustrates the step change of the reference output voltage, the reference output voltage has been increased 21% at t =2 s, and has been decreased 21% at t = 4s, from 140V to 170V and vice versa. After a short transient, the output voltage is maintained close to its reference with a short response time and a very little overshoot, from 140V to 170V, where the response time is around 0.11s with overshoot around 0.5V (neglected). During the reference output voltage variation from 170V to 140V the response time is observed around 0.18s with overshoot around 1V. Also, the stability has been ensured with our controller.

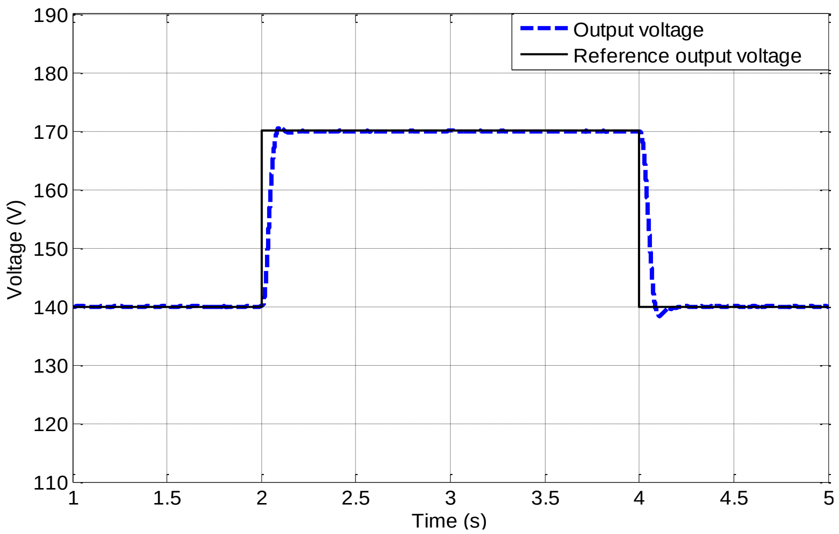


Fig. 3.7 Variation of output voltage.

3.4.4 Experimental validation

The experimental test bench was presented in the previous chapter.

Experimental Test 1—System with PFC: In this test, the PFC act was implemented at nominal loading condition with nominal output voltage (100Ω, 150V), Fig. 3.8 and Fig. 3.9 show the responses of the system in the steady state for the proposed method, the system exhibits excellent results with PFC. Reduction of the THD around 3.9%, the power factor close to unity is achieved, PF=0.993. The output voltage is maintained close to its reference after a short transient, the input current has a sinusoidal waveform and in phase with the source voltage.

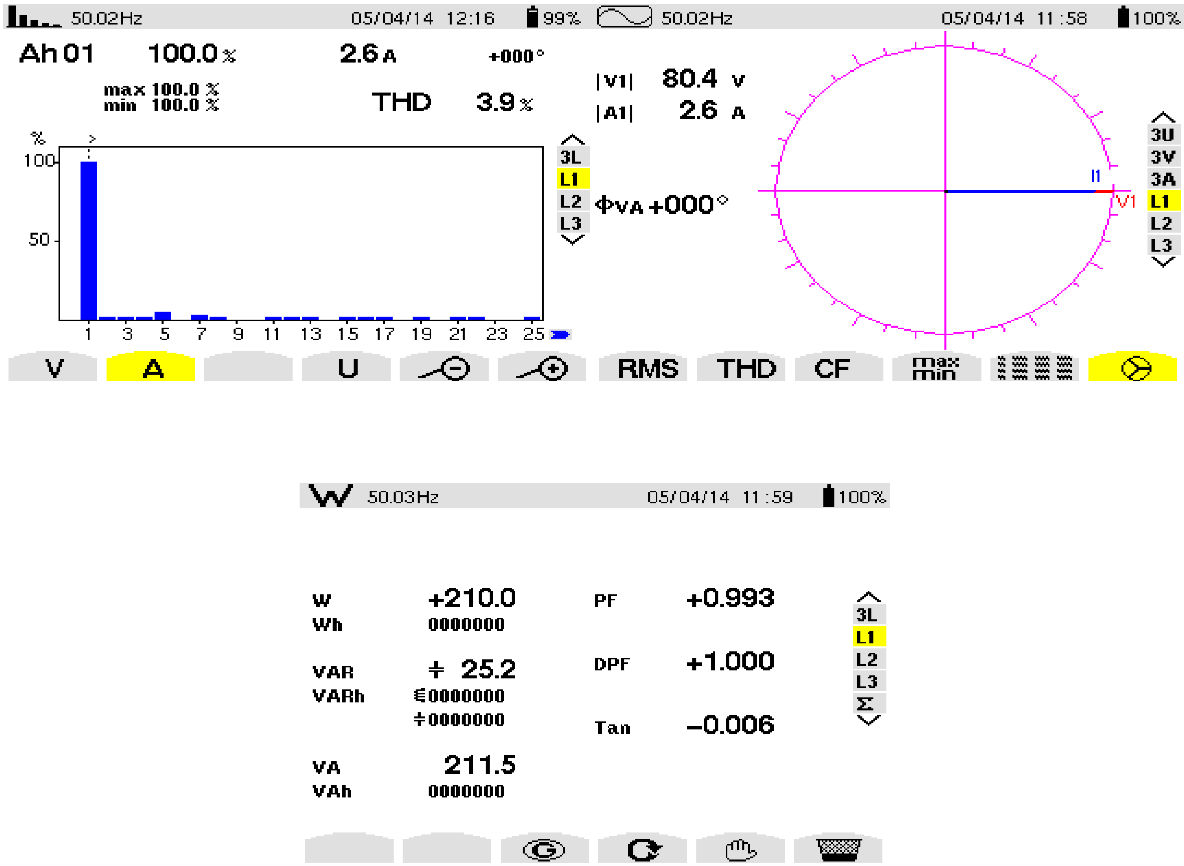


Fig. 3.8 Experimental values in steady state with APFC.

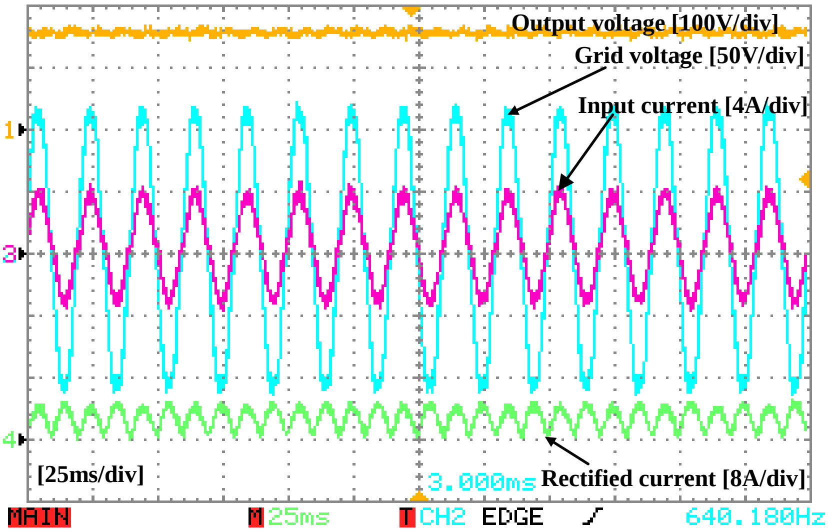


Fig. 3.9 Experimental results in steady state with APFC.

Experimental Test 2—Load Change: This test consist the step change of the load. Fig. 3.10 shows the output voltage during the step load changes by keeping the reference output voltage fixed at 130V. When the system has been reached the steady state a large step change in the load from 100Ω to 50Ω and from 50Ω to 100Ω has been taken place. After a short transient, the output voltage is maintained constant at its reference with a small fluctuation. The response time during load increament is 0.09s and load decreament is 0.1s for the proposed method, which are nearly similar to simulation values and much better than previous works.

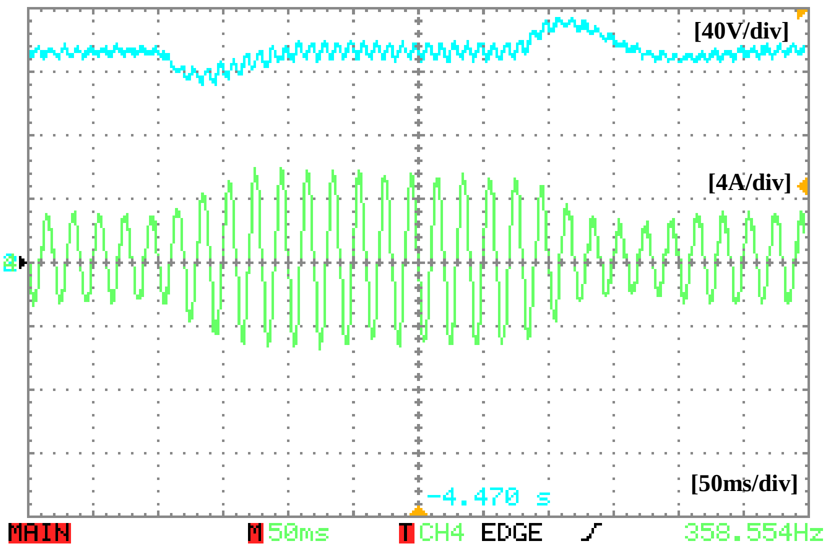


Fig. 3.10 Transient of the step change of R from 100Ω to 50Ω, and from 50Ω to 100Ω.

Experimental Test 3—Reference output voltage change: Another test has been performed by changing the reference output voltage, V_{ref} is changed from 140V to 170V and vice versa with constant load of 100Ω. The transient input current and output voltage are presented in Fig. 3.11 and Fig. 3.12. For the proposed method, the new value of the output voltage has been reached after 0.075s and need only 0.05s to return to previous value again. In this case also, the shape of the input current is found sinusoidal which is in phase with the source voltage during this change.

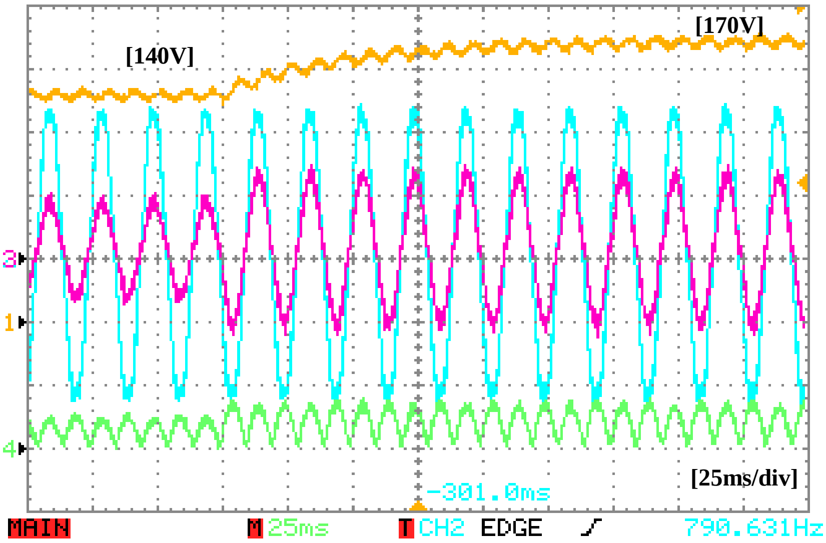


Fig. 3.11 Transient of the step change of V_{ref}, increasing from 140V to 170V.

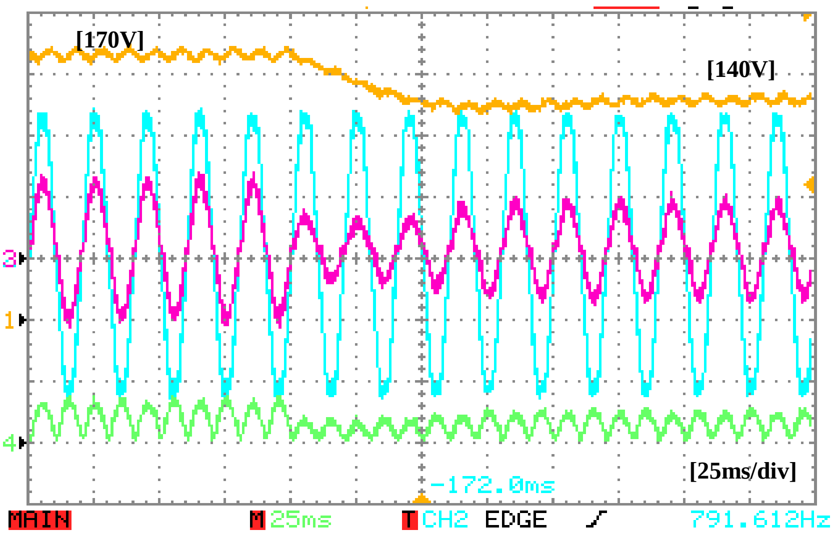


Fig. 3.12 Transient of the step change of V_{ref}, decreasing from 170V to 140V.

After the previous figures, we can be observed that there is a good accordance between the experimental and the simulation results. Is can be seen that the proposed method exhibits better performances compared to the previous works in steady state performance and transient responses.

3.5 Summary

In this chapter, basic concepts and properties of sliding modes have been discussed concisely. The main advantages of the sliding mode control technique are the simplicity in both implementation and design and the inherent robustness with respect to matched internal and external uncertainties and disturbances. Also, we presented in this chapter design, simulation and real time implementation of single-phase boost power factor correction converter by using two control strategies, higher order sliding mode controller (a second) based on super twisting algorithm and predictive techniques. The simulation results showed that the design of the PFC boost converter controller using the proposed controllers has enhanced the converter performance. The system does not get influenced during the parameters variation (load, reference output voltage) in the steady and transient states in the presence of PFC converter. The output voltage tracks its reference perfectly; the input current is in sinusoidal waveform and in same phase with the grid voltage. The THD of input current is measured and satisfy the international standards, THD<5.0%, the unity power factor is measured as 0.993; the system is stable during the changes of the reference output voltage and the load.

Chapter 4: Fuzzy Logic Control

4.1 Introduction

If PID control is inadequate – for example, in the case of higher-order plants, systems with a long dead time, or systems with oscillatory modes – fuzzy control is an option. But first, let us consider why one would not use a PID controller:

- ✓ The PID controller is well understood, easy to implement – both in its digital and analog forms – and it is widely used. By contrast, the fuzzy controller requires some knowledge of fuzzy logic. It also involves building arbitrary membership functions.
- ✓ The fuzzy controller is generally nonlinear. It does not have a simple equation like the PID, and it is more difficult to analyse mathematically; approximations are required, and it follows that stability is more difficult to guarantee.
- ✓ The fuzzy controller has more tuning parameters than the PID controller. Furthermore, it is difficult to trace the data flow during execution, which makes error correction more difficult.

On the other hand, fuzzy controllers are used in industry with success. There are several possible reasons:

Since the control strategy consists of if – then rules, it is easy for a plant operator to read. The rules can be built from a vocabulary containing everyday words such as ‘high’, ‘low’, and ‘increasing’. Plant operators can embed their experience directly.

The fuzzy controller accommodates many inputs and many outputs. Variables can be combined in an ‘if’ – then rule with the connectives ‘*and*’ and ‘*or*’. Rules are executed in parallel, implying a recommended action from each. The recommendations may be in conflict, but the controller resolves conflicts.

4.2 Fuzzy systems

Zadeh's was the first who explain why there is a need for a fuzzy system theory [Zadeh, 1965]. For most complex systems where few numerical data exist and where only ambiguous or imprecise information may be available, fuzzy reasoning offers a way to understand system behaviour by allowing one to interpolate approximately between observed input and output situations. The imprecision in fuzzy models is generally quite high.

Fuzzy logic is based on the way the brain deals with inexact information. Fuzzy systems combine fuzzy sets with fuzzy rules to produce overall complex nonlinear behaviour. Fuzzy systems are structured numerical estimators. They start from highly formalized insights about the structure of categories found in the real world and then express fuzzy 'if-then' rules as some expert knowledge. Being numerical model-free estimators and dynamical systems, fuzzy systems are able to improve the intelligence of systems working in an uncertain, imprecise, and noisy environment. Some of the information available in developing models of physical processes might be judgmental, perhaps an instinctive reaction on the part of the modeller, rather than hard quantitative information.

Fuzzy reasoning allows us to incorporate intuition into a problem. One prevalent way to convey information is our own means of communication: natural language. By its very nature, natural language is vague and imprecise; yet it is the most powerful form of communication and information exchange among humans. Despite the vagueness in natural language, humans have little trouble understanding one another's concepts and ideas; this understanding is not possible in communications with a computer, which requires extreme precision in its instructions.

Fuzzy systems have been shown to be capable of modelling complex nonlinear processes to arbitrary degrees of accuracy. They have attracted growing interest of researchers in various scientific and engineering areas. The number and variety of applications of fuzzy systems have been increasing, ranging from consumer products and industrial process control to medical instrumentation, information systems, and decision analysis.

4.3 Principle of fuzzy logic controller [Sivanandam, 2007].

Fig. 4.1 shows the block diagram of a typical fuzzy logic controller (FLC) and the system plant. There are five principal elements to a fuzzy logic controller, Fuzzification, knowledge base, inference, rule base, and defuzzification.

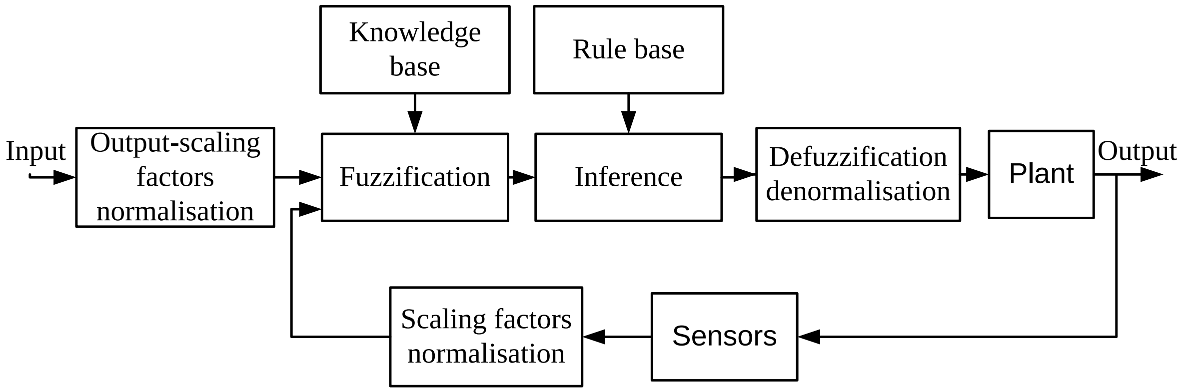


Fig. 4.1 Block diagram of FLC [Cirstea, 2002].

4.3.1 Fuzzification

Fuzzification is an important concept in FLC, it is the process where the crisp quantities are converted to fuzzy by identifying some of the uncertainties present in the crisp value, the conversion of fuzzy values is represented by the membership function where it classifies the element in the set, whether it is discrete or continuous. The membership functions can also be formed by graphical representations. The graphical representations may include different shapes. There are certain restrictions regarding the shapes used. The rules formed to represent the fuzziness in an application are also fuzzy. The "shape" of the membership function is an important criterion that has to be considered. There are different methods to form membership functions.

4.3.2 Knowledge base

The knowledge base consists of a database of the plant. It provides all the necessary definitions for the fuzzification process such as membership functions, fuzzy set representation of the input–output variables and the mapping functions between the physical and fuzzy domain

4.3.3 Rules base

Rules form the basis for the FL to obtain the fuzzy output. The rule-based system is different from the expert system in the manner that the rules comprising the rule-based system originate from sources other than that of human experts and hence are different from expert systems. The rule-based form uses linguistic variables as its antecedents and consequents. The antecedents express an inference or the inequality, which should be satisfied. The consequents are those, which we can infer, and is the output if the antecedent inequality is satisfied. The fuzzy rule-based system uses IF–THEN rule-based system, given by, IF antecedent, THEN consequent. The properties for the sets of rules are, Completeness, Consistency, Continuity, and Interaction.

4.3.4 Fuzzy Inference System

Fuzzy inference systems (FISs) is a major unit of an FLC. The decision-making is an important part in the entire system. The FIS formulates suitable rules and based upon the rules the decision is made. This is mainly based on the concepts of the fuzzy set theory, fuzzy IF THEN rules, and fuzzy reasoning. FIS uses "IF...THEN..." statements, and the connectors present in the rule statement are "OR" or "AND" to make the necessary decision rules. The basic FIS can take either fuzzy inputs or crisp inputs, but the outputs it produces are almost always fuzzy sets. When the FIS is used as a controller, it is necessary to have a crisp output. Therefore in this case defuzzification method is adopted to best extract a crisp value that best represents a fuzzy set.

The most important types of fuzzy inference methods are Mamdani's inference method, which is the most commonly seen inference method. It was introduced by Mamdani (1976) [Mamdani, 1976]. The second one is the so-called Takagi–Sugeno method. It was introduced by Sugeno (1985) [Takagi, 1985]. The difference between the two methods lies in the consequent of fuzzy rules. The first method use fuzzy sets as rule consequent whereas the second one employ linear functions of input variables as rule consequent. All the existing results on fuzzy systems as universal approximators deal with Mamdani fuzzy systems only and no result is available for TS fuzzy systems with linear rule consequent.

4.3.5 Defuzzification

Defuzzification means the fuzzy to crisp conversions. The fuzzy results generated cannot be used as such to the applications, hence it is necessary to convert the fuzzy quantities into crisp quantities for further processing. This can be achieved by using defuzzification process. The defuzzification has the capability to reduce a fuzzy to a crisp single-valued quantity or as a set, or converting to the form in which fuzzy quantity is present. Defuzzification can also be called as "rounding off" method. Defuzzification reduces the collection of membership function values in to a single sealer quantity. There are seven method used for defuzzifying the fuzzy output function. They are. Max-membership principle, Centroid method, Weighted average method, Mean-max membership, Centre of sums, Centre of largest area, and First of maxima or last of maxima. In our application below, the centroid method has been used.

An Automatic change in the design parameters of any of the five elements above creates an adaptive fuzzy controller. Fuzzy control systems with fixed parameters are non-adaptive.

Other non-fuzzy elements which are also part of the control system include the sensors, the analogue–digital converters, the digital–analogue converters and the normalisation circuits. There are usually two types of normalisation circuits: one maps the physical values of the control inputs onto a normalised universe of discourse and the other maps the normalised value of the control output variables back onto its physical domain.

4.4 Types of fuzzy controller [Jantzen, 2007]

4.4.1 Fuzzy P Controller

In discrete time, a proportional controller is defined by:

$$u(n) = K_p e(n) \tag{4.1}$$

The fuzzy proportional (FP) controller in the block diagram in Fig. 4.2 accordingly acts on the error e, and its control signal is U. Signals are represented by lower case symbols before gains and upper case symbols after gains. Thus the notation E represents the term error, and

E = GE $*$ e (the symbol $*$is multiplication), and u represents control, where GU $*$ u = U.

The FP controller has two tuning gains GE and GU, where the crisp proportional controller has just one, K_p. The control signal U(n), at the time instant n is generally a nonlinear function of the input e(n),

$$U(n) = f(GE * e(n)) * GU \tag{4.2}$$

The function *f* denotes the rule base mapping. It is generally nonlinear, as mentioned; but with a favourable choice of design with approximation is:

$$f(GE * e(n)) \approx GE * e(n) \tag{4.3}$$

Insertion into (4.2) yields the control signal:

$$U(n) = GE * e(n) * GU = GE * GU * e(n) \tag{4.4}$$

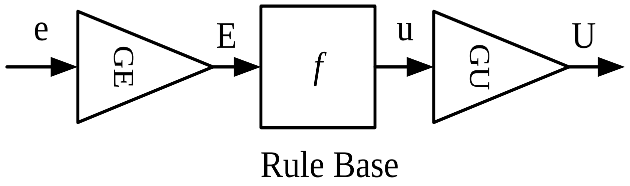


Fig. 4.2 Fuzzy proportional controller, FP

Comparing with (4.1) the product of the gain factors for the controller corresponds to the proportional gain,

$$GE * GU = Kp \tag{4.5}$$

The approximation is exact if, firstly, we choose the same universe for premise sets and conclusion sets. Given a target proportional gain K_p- from a tuned, crisp P controller– Equation (4.5) determines one fuzzy gain factor when the other is chosen. The equation has one degree of freedom, since the fuzzy P controller has one more gain factor to adjust than the crisp P controller. This can be used to exploit the full range of the premise universe.

4.4.2 Fuzzy PD Controller

Because of the plant dynamics, it will take some time before a change in the control signal is noticeable in the plant output, and the proportional controller will be equally late in correcting for an error. Derivative action helps to predict the future error, and the PD controller uses the derivative action to improve closed-loop stability. The discrete time PD controller is,

$$u(n) = K_p\left(e(n) + T_d\frac{e(n)-e(n-1)}{T_s}\right) \tag{4.6}$$

With the I-action set to zero ($1/T_i = 0$). T_d seconds ahead of the time instant n, where the estimate is obtained by linear extrapolation of the straight line connecting e(n−1) and e(n). With $T_d = 0$ the controller is purely proportional, but when T_d is gradually increased, it will dampen possible oscillations. If T_d is increased too much the step response of the closed-loop system becomes over damped, and it will start to oscillate again. Input to the fuzzy proportional-derivative (FPD) controller in Fig. 4.3 is e(n) and ė(n), where

$$\dot{e}(n) \approx \left[e(n) - e(n-1)\right].Ts \tag{4.7}$$

The backward difference is a simple discrete approximation to the differential quotient, and more accurate digital implementations are available.

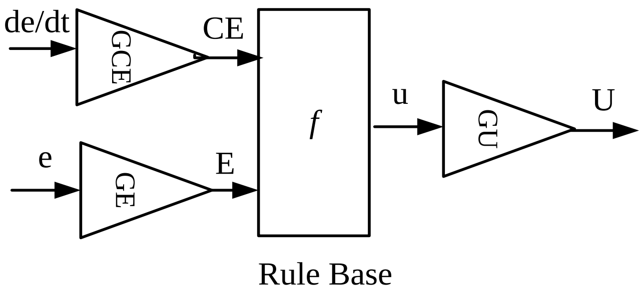


Fig. 4.3 Fuzzy PD controller.

The notation CE represents the change in error, and CE = GCE $*$ ė. Notice that (4.7) deviates from the straight difference e(n)−e(n−1) used in the early days of fuzzy control. The control signal U(n), at the time instant n, is a nonlinear function of error and change in error,

$$U(n) = f(GE*e(n),\ GCE*(n))*GU \tag{4.8}$$

Again the function *f* is the rule base mapping, only this time it is a surface depending on two variables. It is usually nonlinear, but with a favourable choice of design, an approximation is:

$$f\left(GE*e(n),\ GCE*(n)\right) \approx GE*e(n) + GCE*\dot{e}(n) \tag{4.9}$$

Insertion into (4.8) yields the control action for the linear controller,

$$\begin{aligned} U(n) &= \left(GE*e(n) + GCE*(n)\right)*GU \\ &= GE*GU*\left(e(n) + (GCE/GE)\right)\dot{e}(n) \end{aligned} \tag{4.10}$$

Comparing (4.6) and (4.10), the gains are related as follows:

$$GE * GU = Kp \tag{4.11}$$

$$GCE / GE = T_d \tag{4.12}$$

The approximation is exact when the fuzzy control surface is a plane acting like a summation. The conclusion universe must be defined as the sum of the premise universes, then the control surface will be the plane u(n) = E(n) + CE(n). By that choice, the controller is equivalent to a crisp PD controller, and we can exploit (4.11) and (4.12).

The fuzzy PD controller may be applied when proportional control is inadequate. The derivative term reduces overshoot, but it may be sensitive to noise as well as abrupt changes of the reference causing derivative kick in (4.7).

There are four consecutive cases marked by circles in the plots:

Case 1: E> 0, CE < 0 the error is large and positive, and the plant output is moving towards the reference. The error E = GE ∗ e is positive as long as the plant output is below the reference. Furthermore, the change in error is negative, as long as the plant output is increasing. The situation corresponds to the fourth quadrant of the phase plane. The phase trajectory spirals in a clockwise direction.

Case 2: E< 0, CE < 0 the plant output has overshoot the reference and is still moving away from the reference. The error is negative, since the plant output is above the reference. Furthermore, the change in error is negative, since the plant output is still increasing. The situation corresponds to the third quadrant of the phase plane.

Case 3: E< 0, CE > 0 the plant output is returning towards the reference. The error is negative, since the plant output is above the reference. Furthermore, the change in error is positive, since the plant output is now decreasing. The situation corresponds to the second quadrant of the phase plane.

Case 4: E> 0, CE > 0 the plant output is moving away from the reference during an

Undershoot. The error is positive, and the plant output is below the reference. Furthermore, the change in error is positive, and the plant output is decreasing. The situation corresponds to the first quadrant of the phase plane.

Each case corresponds to a quadrant in the phase plane, and the trajectory of the response can be affected or shaped to an extent by local rules in each quadrant.

4.4.3 Fuzzy PID Controller

If the closed-loop system exhibits a sustained error in steady state, integral action is necessary. The integral action will increase (decrease) the control signal if there is a positive (negative) error, even for small magnitudes of the error. Thus, a controller with integral action will always return to the reference in steady state.

A fuzzy PID controller acts on three inputs: error, integral error, and change in error. A rule base with three premise inputs can be a problem. With three premise inputs, and, for example, three linguistic terms for each input, the complete rule base consists of $3^3 = 27$ rules, making it cumbersome to maintain. Furthermore, it is difficult to settle on rules concerning the integral action, because the initial and final values of the integral depends on the load. The integral action in the crisp PID controller serves its purpose, however, and a simple design is to combine crisp integral action and a fuzzy PD rule base in the fuzzy PD+I (FPID) controller (see Fig. 4.4).

The integral error IE = GIE∫edt is proportional to the accumulation of all previous error measurements in discrete time, with

$$\int edt \approx \sum_{j=1}^{n} e(j)T_s \tag{4.13}$$

Rectangular integration is a simple approximation to the integral, and more accurate approximations exist. The control signal U(n) after the gain GU, at the time instant n, is a nonlinear function of error, change in error, and integral error,

$$U(n)=\left[f(GE*e(n),GCE*\dot{e}(n))+GIE\sum_{j=1}^{n}e(j)T_s \right]*GU \tag{4.14}$$

$$\begin{aligned} U(n) &\approx \left[GE*e(n)+GCE*\dot{e}(n))+GIE\sum_{j=1}^{n}e(j)T_s \right]*GU \\ &= GE*GU*\left[e(n)+\frac{GCE}{GE}*\dot{e}(n))+\frac{GIE}{GE}\sum_{j=1}^{n}e(j)T_s \right]*GU \end{aligned} \tag{4.15}$$

In the last line we have assumed that GE is non-zero. Comparing with (4.15), the gains are related as follows:

$$\begin{aligned} GE*GU &= K_p \\ \frac{GCE}{GE} &= T_d \\ \frac{GIE}{GE} &= \frac{1}{T_i} \end{aligned} \tag{4.16}$$

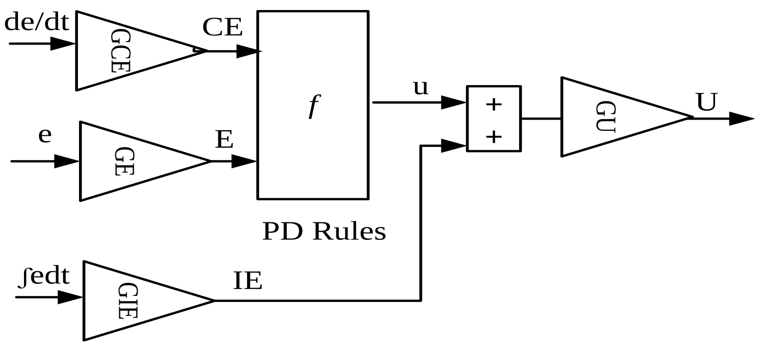


Fig. 4.4 Fuzzy PID controller.

The FPID controller provides all the benefits of PID control, but also the disadvantages regarding derivative kick. The integral error removes any steady state error, but can also cause integrator windup.

4.4.4 Fuzzy Incremental Controller

An incremental controller adds a change in control signal u to the current control signal

$$
\begin{aligned}
& u(n) = u(n-1) + \Delta u(n) T_s \Rightarrow \\
& \Delta u(n) = K_p \left(\frac{e(n) - e(n-1)}{T_s} + \frac{1}{T_i} e(n) \right)
\end{aligned}
\tag{4.17}
$$

The controller output is an increment to the current control signal. The fuzzy incremental (FInc) controller in Fig. 4.5 is of almost the same configuration as the FPD controller except for the added integrator. The conclusion in the rule base is now called change in output (cu), and the gain on the output is, accordingly, GCU. The control signal U(n) at time instant n is the sum of all previous increments,

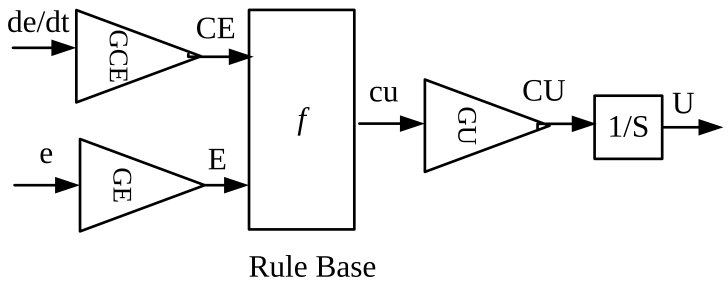


Fig. 4.5 Incremental fuzzy controller.

$$
\begin{aligned}
U(n) &= \sum_{j=1}^{n} \left(cu(j) * GCU * T_s \right) \\
&= \sum_{j=1}^{n} \left(f(GE * e(j), GCE * \dot{e}(j)) * GCU * T_s \right)
\end{aligned}
\tag{4.18}
$$

Notice again that this definition deviates from the historical fuzzy controllers, where the sampling period Ts was left out. The function f is again the control surface of a PD rule base. The mapping is usually nonlinear, but with the usual favourable choice of design, (4.9) is a linear approximation. Insertion into (4.18) yields the control action,

$$\begin{aligned}
U(n) &\approx \sum_{j=1}^{n}(GE*e(j)+GCE*\dot{e}(j))*GCU*T_s \\
&= GCU*\sum_{j=1}^{n}\left[GE*e(j)+GCE*\frac{e(j)-e(j-1)}{T_s}\right]*T_s \\
&= GCU*\left[GE*\sum_{j=1}^{n}e(j)*T_s+GCE*\sum_{j=1}^{n}(e(j)-e(j-1))\right] \qquad (4.19)\\
&= GCE*GCU*\left[\frac{CE}{GCE}\sum_{J=1}^{n}(e(j)*T_s)+e(n)\right]
\end{aligned}$$

By comparing (4.19) it is clear that the controller is a crisp PI controller (T_d = 0), and the gains are related as follows:

$$\begin{cases} \mathrm{GCE}*\mathrm{GCU} = \mathrm{K_p} \\ \mathrm{GE}/\mathrm{GCE} = 1/\mathrm{T_i} \end{cases} \qquad (4.20)$$

Notice that the proportional gain K_p now depends on GCE. The gain $1/T_i$ is determined by the ratio between the two fuzzy input gains, and is the inverse of the derivative gain T_d in FPD control; the gains GE and GCE change roles in FPD and FInc controllers. It is an advantage that the controller output is driven directly from an integrator, because (1) simply limiting the integrator prevents integrator windup, and (2) the integrator cancels noise to an extent that smooths the control signal.

The fuzzy P controller may be used as a starting point. To improve the settling time and reduce overshoot, fuzzy PD is the choice. If there is a steady state error, a FInc controller or a fuzzy PID is the choice.

4.5 Fuzzy logic-Based DC Bus Voltage Controller of APFC.

4.5.1 Description of the method

In this work, FLC has been used in order to regulate and maintain dc-link voltage of the boost converter that is depicted in Fig. 4.6 and 4.7 in the desired value. FLC contains two inputs variables, as shown in Fig. 4.8: the first is the error, e(k), which is calculated as the difference between the measured dc-link voltage and the desired reference voltage; the, second input consists of the change in the voltage error, Δ e(k), which helps the controller to be fast and efficient. As output (Fig.4.9), the reference current i_{ref} is produced.

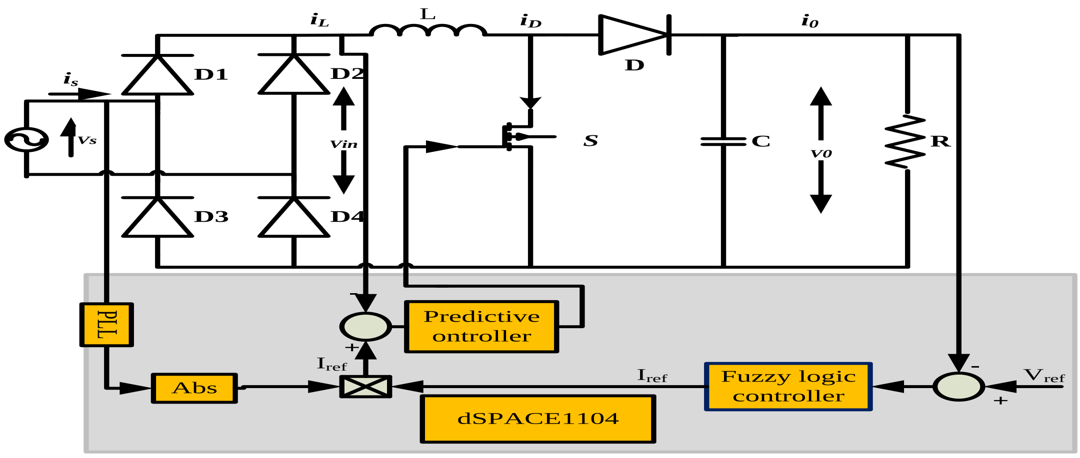


Fig. 4.6 PFC pre-regulator with FLC.

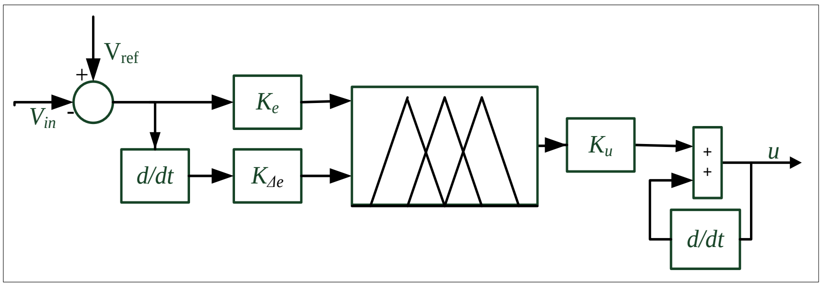


Fig. 4.7 Proposed Fuzzy logic controller.

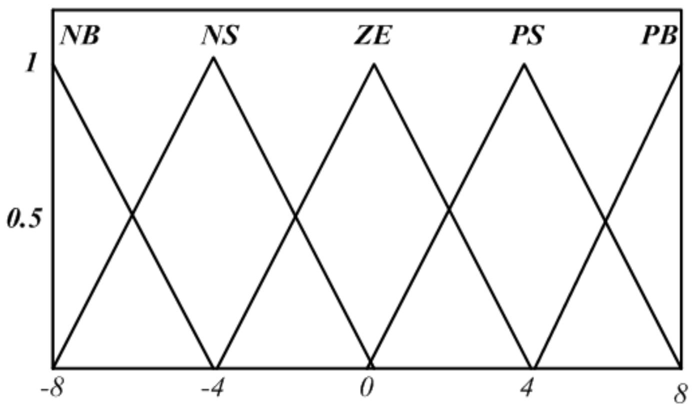


Fig. 4.8 Membership functions for inputs variable.

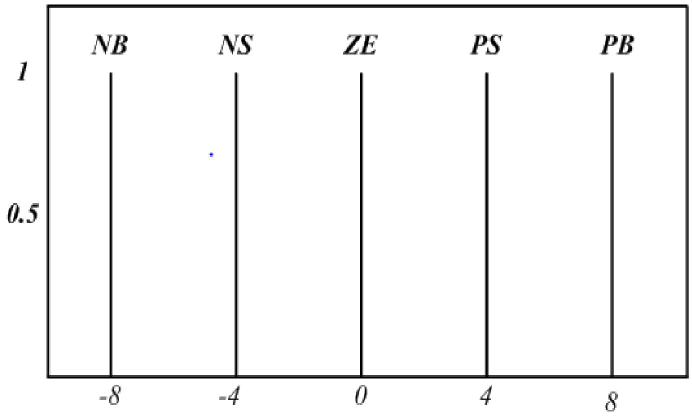


Fig. 4.9 Membership functions for output variable.

According to Fig. 4.7, the controller has normalized by use the scaling factors to get a optimal performance as fallow:

$$\begin{aligned} u_k &= u_{k-1} + K_u * u_k \\ e &= K_e * e \\ \Delta e &= K_{\Delta e} * \Delta e \end{aligned} \qquad (4.21)$$

A total of 25 rules are designed to get the best performance of the controller as illustrated a Table 4.1. Each fuzzy rule in Table 4.1 is in the fallowing form;

Rule (i): if e(k) is NB and Δe(k) is NB then i_{ref} is NB

Table 4.1 Fuzzy rules table

Error e(k)	**Change in error Δe(k)**				
	NB	**NS**	**ZE**	**PS**	**PB**
NB	NB	NB	NS	NS	ZE
NS	NB	NB	NS	ZE	PS
ZE	NB	NS	ZE	PS	PB
PS	NS	ZE	PS	PS	PB
PB	ZE	PS	PS	PB	PB

All steps of the proposed method are as Fig. 4.10.

Step 1: Identification of the parameters.

Step 2: Determinate of suitable voltage and introduce the FLC for regulate it.

Step 3: Determinate I_{ref} through FLC for efficient of the suitable mode and introduce predictive control.

Step 4: Verified that of all duty cycles are calculated by predictive control.

Step 5: Exploitation of the converter

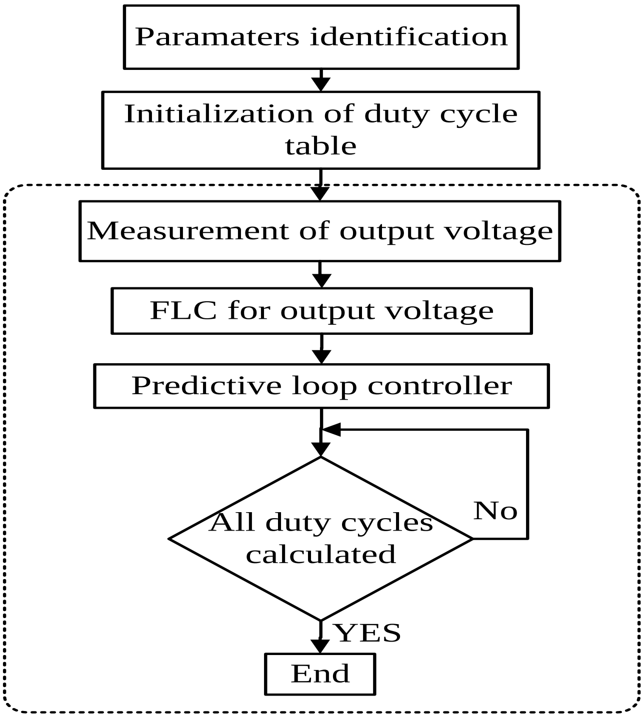


Fig. 4.10 Flowchart of applied method.

4.5.2 Simulation results

The simulation of the proposed model was conducted with MATLAB/ Simulink environment using a fixed step size of 40e^{-6}s; the main parameters of the model are given in Table 4.2. Simulation results of the system with constant load (200Ω) and constant DC link voltage (110 V) are shown in Fig. 4.11. Moreover, various simulation tests have been performed to evaluate the dynamic performance of the controller during the variation of the DC link voltage reference, as shown in Fig. 4.12, and the steps of load, as shown in Fig. 4.13 According to these figures, it can be observed that the proposed control strategy is robust and not influenced during parameters changes (load, output voltage reference), the current take a sinusoidal form in phase with the grid voltage, the steady state error and settling time are enhanced and the output voltage follows the reference value perfectly.

Table 4.2 Circuit parameters

Switching frequency	20KHz
Resistance load	200Ω
Output capacity	1100µF
Input inductance	20mH
DC-link voltage reference	110 V
Source voltage frequency	50Hz
Supply voltage (rms)	50V

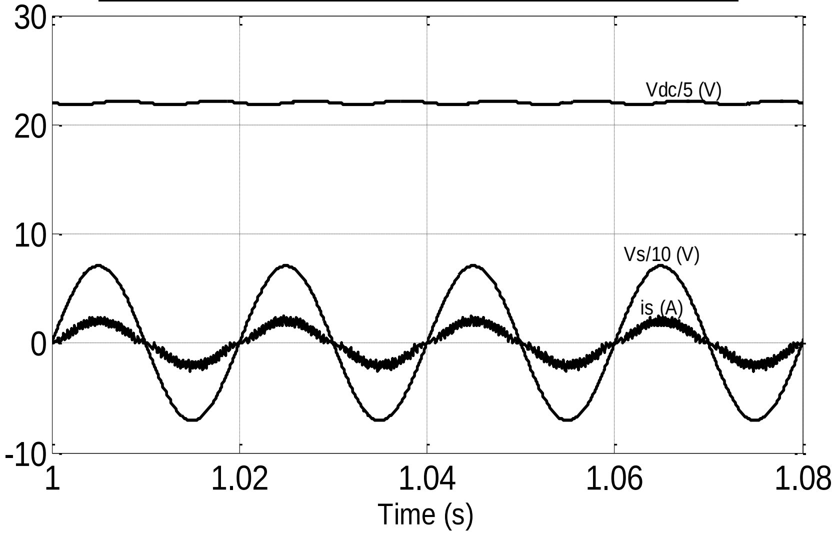


Fig. 4.11 Simulation results, input and output voltage, input current.

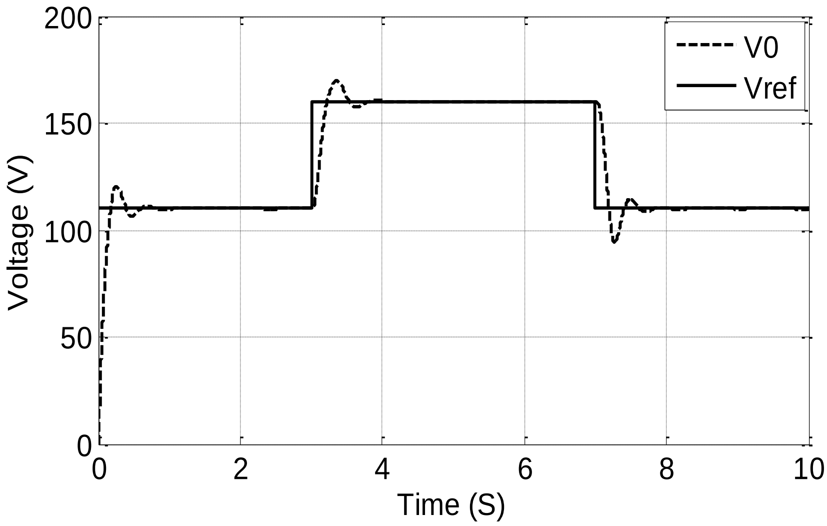


Fig. 4.12 Variation of output voltage from 110V to 160V and from 160V to 110V.

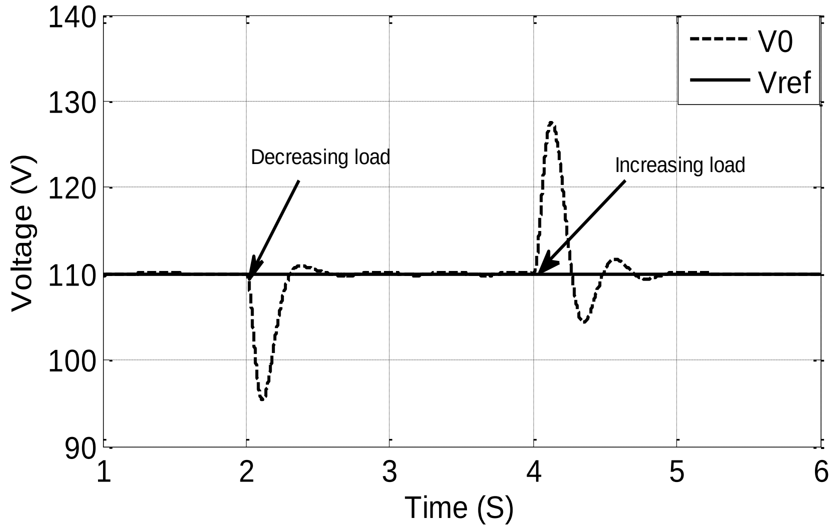


Fig. 4.13 Variation of load from 200Ω to 100Ω and from 100Ω to 200Ω.

4.5.3 Experimental results and discussion

To validate the proposed method, an experimental test bench has been developed, as shown in Fig. 2.11, in LAS laboratory, Setif1 University, Algeria . The prototype contains: an inverter SEMIKRON used as a rectifier, a transformer, current sensors, and voltage sensors. The control program has been simulated in the MATLAB/Simulink environment and implemented in real-time via dSPACE RTI1104.

Test 1: Here, the model is tested with APFC, and the results are presented in Figs. 4.14 and 4.15. The results highlight the effect of APFC with fixed load (200 Ω) and output voltage (110 V); it is observed that the output voltage follows the desired reference, and the source current is in phase with the input voltage. Hence, the THD is approximately 5% and PF is nearly 0.992, an improvement compared to the previous test.

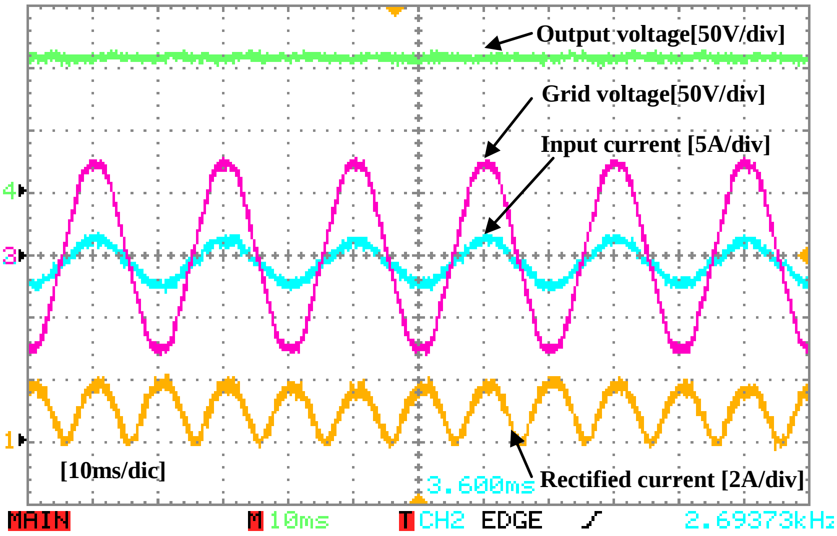


Fig. 4.14. Experimental results waveforms.

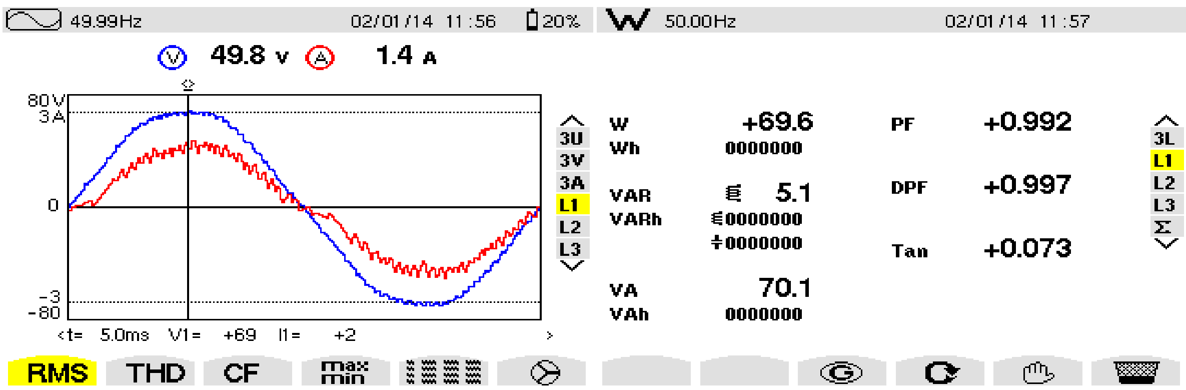


Fig. 4.15 Experimental measurement: PF, voltage and current of source

Test under load changes: in this test, a load variation was applied in order to test the influence of the load perturbation. Fig. 4.16 shows a slight remoteness of the output voltage due to sudden load changes, but the system performance was not decreased.

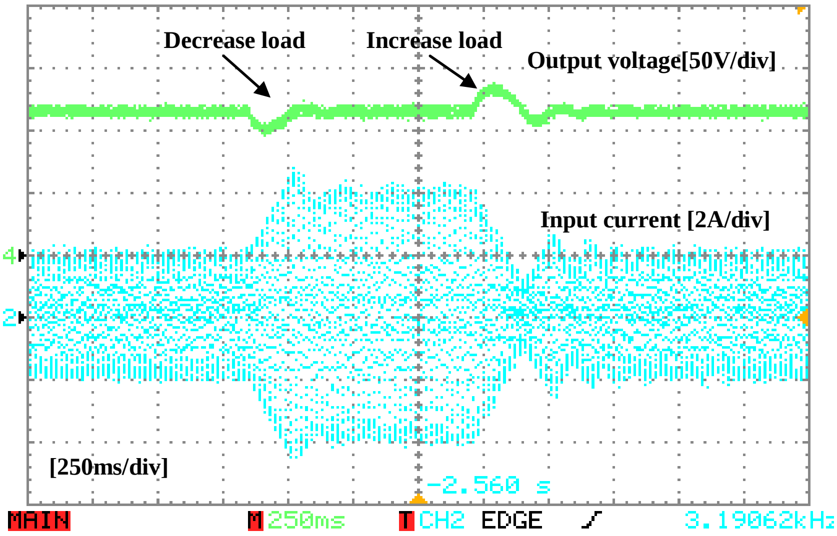
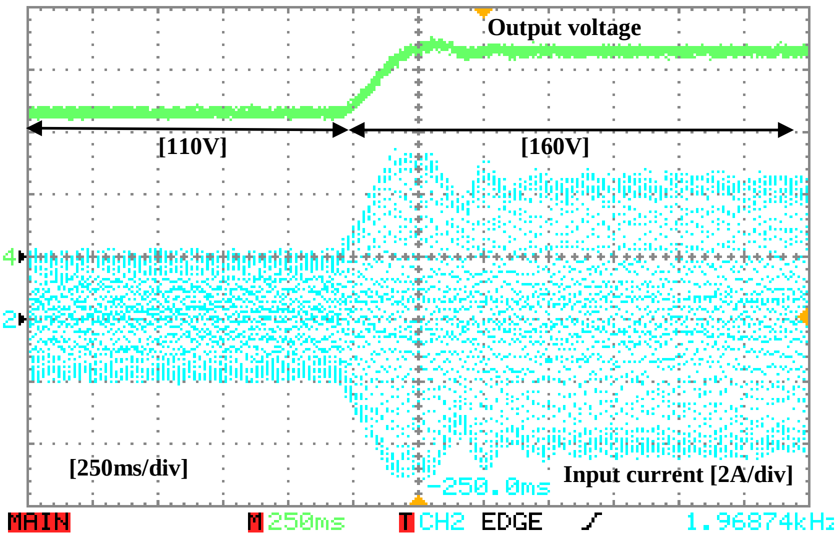
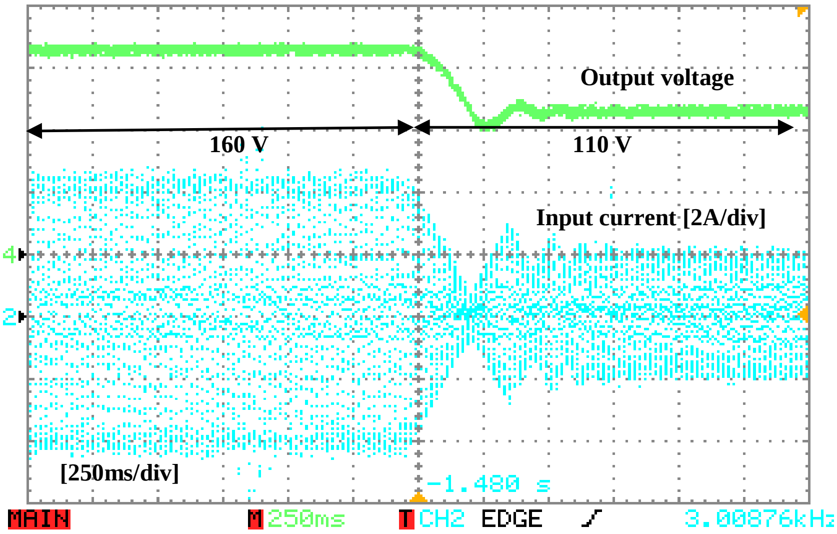


Fig. 4.16 Experimental results for load variation.

Test under voltage reference changes: the value of the output voltage reference has risen by roughly (160v). The results are presented in Fig. 4.17, and they show that the output voltage follows the reference with a lower settling time.



(a)



(b)

Fig. 4.17 Experimental results for voltage variation, **(a)** increasing from 110V to 160V and **(b)** decreasing from 160V to 110V

To show the merits of proposed approaches, a comparison among the proposed and previous methods has been done take into account steady and transient states. All the experimental results are summarized in Table 4.3. Based on three cases; the change of load, change of reference output voltage, and PF.

Table 4.3 Transient values corresponding to experimental results with comparison.

	Increasing load		Decreasing load		Increasing voltage		Decreasing voltage		PF
	ΔV_0(V)	Δt(s)	ΔV_0(V)	Δt(s)	ΔV_0(V)	Δt(s)	ΔV_0(V)	Δt(s)	
FS-MPC	8	0.05	10	0.08	30	0.1	30	0.07	0.996
[Bouafassa, 2014a]	15	0.28	10	0.25	40	0.5	40	0.5	0.992
[Bouafassa 2014b]	10	0.09	18	0.1	30	0.075	30	0.05	0.993
[Kessal, 2011]	12	0.45	16	0.64	32	0.6	32	0.6	0.995
[Kessal, 2012]	14	0.4	13	0.3	38	0.5	38	0.5	0.995
[Kessal, 2014]	25	0.18	25	0.15	40	0.1	40	0.1	0.998

After the Table 4.3, is can be seen that the proposed controllers exhibits better performances compared to the other methods in steady state performance and transient responses, when the overshoot and undershoot in output voltage is enhanced. Moreover, THD required the international standard such as IEEE 519, the PF is near to unity, and the response time is very acceptable.

1.6 Summary

In this chapter a briefly for fuzzy logic controller is presented with its application for improve power factor correction, the proposed algorithm is successful simulated and implemented in real time via dSPACE 1104. The obtained results are very satisfactory compared to previous methods, where the power factor is enhanced around 0.992 and the THD value required the international requirements such as IEEE 519.

General Conclusion

- **The chief work and research results of the thesis**

This thesis has been presented a survey of the theoretical background of unconventional control theory, in particular higher order sliding mode control, model predictive control and fuzzy logic controller, and it has been shown by presenting also new theoretical developments. The major contribution of the present thesis is the application of above control methodologies for improving the performance of the single phase active power factor correction (PFC) AC-DC boost converter.

In the PFC control technologies, we often encounter the external unknown and uncertain disturbances. The most primary is the load, which poses a severe challenge to the system control performances. The application of robustness control techniques has become the studying hot point.

It is well-known that many design methodologies are focused on nonlinear system control design. Nevertheless, some approaches are used to simply design a feedback controller but not necessarily provide robust properties. For this reason, if the conventional PI or PID controller is designed, the final achieved feedback properties may not be satisfactory for a wide range of operating condition and parameter uncertainties. Therefore, unconventional controls have to be made in order to attain enhanced system stability and controller robustness.

In order to verify the simulation results, this thesis adopts dSPACE 1104 to complete the physical experiment test. Through the results from the MATLAB/Simulink simulation, the system downloads software programs of control algorithm into the memorizer of dSPACE's subordinate computer in order to control the PFC to work, and ultimately it achieves on-line and real-time experiment of AC-DC boost. Experimental results show the proposed methods can be successfully applied to PFC. The method not only retains the robustness, but also effectively eliminates the problems in the traditional control. This lays the foundation for the widespread application in the future.

The model predictive control is considered a solution when a slow dynamic of the system have been considered. In chapter 2, we used the model predictive control to improve the performance of the PFC, The main contribution of this chapter is to introduce a novel control approach for PFC based on an optimal predictive current control law and its experimental validation around dSPACE 1104. Further, the objectives are to achieve unity power factor, low THD, minimal digital hardware, and robustness guaranteed for different output voltage and load fluctuation.

The basic notions of the first and higher order sliding mode control theory has been given in Chapter 3, while a brief introduction to the higher order sliding mode control theory and a description of the main features and advantages of higher order sliding modes have been discussed briefly. The application of the higher order sliding mode control combined with model predictive control for PFC is presented by its simulation results and experimental validation.

In chapter 4, we introduced the fuzzy logic controller combined with model predictive control, the simulation and experimental results are very satisfactory, and the proposed controller is robust against parameters variation.

- **Future works**

The study of unconventional control theory is one difficult and challenging topic in present control theories, there are still some important issues need to be explored and further studied:

- ✓ Economic study of the proposed control methods in order to decrease the cost.
- ✓ Try to decrease digital hardware.
- ✓ Integrate the proposed control methods for PFC in the Microgrid system

References

[Allgöwer, 1998] F. Allgöwer, and A. Zheng. Proceedings of international symposium on nonlinear model predictive control : assessment and future directions. Ascona, Switzerland, 1998.

[Allgower, 2000] F. Allgöwer, A. Zheng. Nonlinear model predictive control, progress in systems and control theory. Birkhauser Verlag, **Vol. 26**, 2000.

[Asada, 1986] H. Asada, and J. J.E. Slotine. Robot Analysis and Control. John WILEY & Sons Edition, 1986.

[Azazi, 2014] H.Z. Azazi, E.E. El-Kholy, S.A. Mahmoud, and S.S. Shokralla. Power factor correction using predictive current control for three-phase induction motor drive system. Electric Power Component and System, **Vol. 42(2)**:190–202, 2014.

[Bouafassa, 2014a] A. Bouafassa, L. Rahmani, A. Kessal, B. Babes. Unity power factor converter based on a Fuzzy controller and Predictive Input Current, **Vol. 53(6)**: 1817-1821, 2014.

[Bouafassa, 2014b] A. Bouafassa, L. Rahmani, S. Mekhilef. Design and real time implementation of single phase boost power factor correction converter. doi:10.1016/j.isatra.2014.10.004.

[Boukezzoula, 2000] R. Boukezzoula. Commande floue d'une classe de systèmes non linéaires: application au problème de suivi de trajectoire. PhD thesis, Université de Chambéry, 2000.

[Bibian, 2001] S. Bibian, and H. Jin. Digital control with improved performance for boost power factor correction circuits. In: Proceedings of the IEEE Applied Power Electronics Conference APEC, **Vol. 1**:137–143, 2001.

[Braikia, 2011] K. Braikia, M. Chettouh, B. Tondu, P. Acco, and M. Hamerlain. Improved control strategy of 2-sliding controls applied to a flexible robot arm. Advanced Robotics Journal, **Vol. 25 (11–12)**, 2011.

[Camacho, 1993] E. F . Camacho. Constrained Generalized Predictive Control. IEEE Transactions on Automatic Control, **Vol. 38(2)**: 327–332, 1993.

[Castillo, 2008] O. Castillo and P. Melin. Type-2 fuzzy logic: theory and applications. Springer Verlag, 2008.

[Camacho, 2004] E.F Camacho, and C. Bordons Alba. Model Predictive Control. Eds springer, 2004.

[Chafaa, 2007] K. Chafaa, L. Saidi, M. Ghanai, and K. Benmahammed. Indirect adaptive interval type-2 fuzzy control for nonlinear systems. International Journal of Modelling, Identification and Control, **Vol. 2(2)**: 106–119, 2007.

[Clarke, 1987] D.W. Clarke, C. Mohtadi, and P. S. Tuffs . Generalized predictive control-part I: the basic algorithm. Automatica , **Vol. 23(2)**: 137–148, 1987.

[Chaoui, 2001] F.Z. Chaoui, F. Giri, and M. M'saad. Adaptive control of input-constrained type-1 plants stabilization and tracking. Automatica, **Vol. 37(2)**: 197–203, 2001.

[Chen, 1982] C. Chen, and L. Shaw. On receding horizon feedback control. Automatica, **Vol. 18(3)**: 349–352, 1982.

[Chen, 1998] H. Chen, and F. Allgöwer. A quasi-infinite horizon non linear model predictive control scheme with guaranteed stability. Automatica, **Vol. 34**: 1205–1217, 1998.

[Chen, 1999b] W.H . Chen, D.J. Ballance, P.J. Gawthrop, J.J. Gribble, and J. O'Reilly. Nonlinear PID predictive controller. IEE Proceedings, Part D-Control Theory and Applications **, Vol. 146(6)**: 603–611, 1999.

[Chen, 2000] W.H . Chen, D.J. Balance, and J. O'Reilly. Model predictive control of nonlinear systems: computational burden and stability. IEE Proceedings-Control Theory and Applications, **Vol. 147(4)**, 2000.

[Chen, 2003] W.H . Chen. Optimal control of nonlinear systems: a predictive control approach. Automatica , **Vol. 39(1)**: 633–641, 2003.

[Chen, 2004] W.H. Chen. Predictive control of general nonlinear systems using approximation. IEE Proceedings-Control Theory and Applications, **Vol. 151(2)**:137–144, 2004.

[Chettouh, 2008] M. Chettouh, R. Toumi, and M. Hamerlain. High order sliding modes for a robot driven by pneumatic artificial rubber muscles. International Journal of Robotics and Automation, **Vol. 23(2)**:88–97, 2008.

[Cirstea, 2002] M.N. Cirstea, A. Dinu, J.G. Khor, and M. McCormick. Neural and fuzzy logic control of drives and power systems. British Library, 2002.

[Darlington, 2000] J. Darlington, C.C. Pantelides , B. Rustem, and B.A. Tanyi. Decreasing the sensitivity of open-loop optimal solutions in decision making under uncertainty. European Journal of Operational Research, **Vol. 121(2)**: 343–362, 2000.

[DeCarlo,1988] R.A. DeCarlo, S.H. Zak, and G.P. Matthews. Variable structure control of nonlinear multivariable systems: A tutorial. Proceeding of IEEE, **Vol. 76(3)**: 212-232, 1988.

[Demircioglu, 1989] H. Demircioglu. Continuous-time Self-tuning Algorithms. Doctoral These, Glasgow University, Department of Mechanical Engineering, 1989.

[Demircioglu, 1991] H. Demircioglu, and P. J. Gawthrop. Continuous-time Generalized Predictive Control. Automatica, **Vol. 27(1)**: 55–74, 1991.

[Demircioglu, 1992a] H. Demircioglu, and P. J. Gawthrop. Multivariable Continuous-time Generalized Predictive Control. Automatica , **Vol. 28(4)**: 697–713, 1992.

[Demircioglu, 1999] H. Demircioglu. Constrained continuous-time generalised predictive control. IEE Proceedings-Control Theory and Applications , **Vol. 146(5)**: 470–476, 1999.

[Demircioglu, 1992b] H . Demircioglu, and D. W . Clarke. CG PC with guaranteed stability properties. IEE Proceedings, Part D- Control Theory and Applications , **Vol. 139(4)**:371–380, 1992.

[Demircioglu, 2000] H. Demircioglu, and E. Karasu. Generalized predictive control. A practical application and comparison of discrete-and continuous -time versions. IEEE Control Systems Magazine, **Vol. 20(5)**: 36–47, 2000.

[Edwards,1998] C. Edwards and S.K. Spurgeon. Sliding mode control: Theory and applications. Tayler & francis, 1998.

[Emelyanov, 1986a] S.V. Emelyanov, S.K. Korovin, and L.V. Levantvosky. Higher order sliding modes in the binary control systems. Soviet Physics, Doklady, **Vol. 31(4)**: 291–293, 1986.

[Emelyanov, 1986b] S.V. Emelyanov, S.K. Korovin, and L.V. Levantvosky. Second order sliding modes in controlling uncertain systems. Soviet Journal of Computer and Science, **Vol. 24**: 63–68, 1986.

[Emelyanov, 1986c] S.V. Emelyanov, S.K. Korovin, and L.V. Levantvosky. A new class of second order sliding algorithms. Mathematical Modeling, **Vol. 2(3)**:89–100, 1986. In Russian.

[Eker, 2010] İ. Eker. Second-order sliding mode control with experimental application. ISA Transactions, **Vol. 49(3)**:394–405, 2010.

[Essounbouli, 2006] N. Essounbouli, A. Hamzaoui, and J. Zaytoon. Fuzzy sliding mode control for a class of nonlinear continuous systems. International Journal of Computer Applications in Technology, **Vol. 25(2)**:174–182, 2006.

[Feng, 2002] W. Feng, J. O'Reilly, and D.J. Ballance. MIMO nonlinear PID predictive controller. IEE Proceedings-Control Theory and Applications, **Vol. 149(3)**: 203–208, 2002.

[Fliess,1990a] M. Fliess. Generalized controller canonical form for linear and nonlinear dynamics. IEEE Transaction on Automatic Control, Vol. 35(9): 994-1001, 1990.

[Fliess,1990b] M. Fliess. What the Kalman state variable representation is food for. Proceeding of 29th Conference on Decision and Control, Honolulu, Hawaii, 1282-1287, 1990.

[Filippov, 1988] A.F. Filippov. Differential equations with discontinuous righthand sides. Kluwer Academic Publishers, 1988.

[Isidori, 1989] A. Isidori. Nonlinear control systems. Springer–Verlag Edition, 1989.

[Gasso, 2002] K. Gasso, G. Mourot, and J. Ragot. Structure identification of multiple models with output error local models. In 15th IFAC World Congress on Automatic Control, Barcelona, Spain, 2002.

[Gawthrop, 1998] P. J. Gawthrop, H . Demircioglu, and I.I. Siller-Alcalá. Multivariable continuous –time generalised predictive control: A states pace approach to linear and nonlinear systems. IEE Proceedings - Control Theory and Applications , **Vol. 145(3)**: 241–250, 1998.

[García, 1989] C. E. García, D. M . Prett, and M . Morari. Model predictive control : theory and practice – a survey. Automatica, **Vol. 25(3)**: 335–348, 1989.

[Gawthrop, 1987] P. J. Gawthrop. Continuous-time self-tuning control. Letchworth, U. K. : Reaserch studies Press , **Vol. 1**, 1987.

[Grüne, 2011] L. Grüne, and J. Pannek, . Nonlinear Model Predictive Control. Eds springer, 2011.

[Huangfu, 2010] Yi-Geng Huangfu. Research of Nonlinear System High Order Sliding Mode Control and its Applications for PMSM. Doctoral thesis University of Technology of Belfort-Montbéliard, 2010.

[Harashima, 1986] F. Harashima, H. Hashimoto, and K. Maruyama. Practical robust control of robot arm using variable structure system. Proceedings of IEEE International Conference on Robotics and Automation, **Vol. 3**:532–539, 1986.

[Hermann,1977] R. Hermann, and A.J. Krener, Nonlinear controllability and Observability, IEEE Transactions on Automation and Control, **Vol. 22(9)**: 728–740, 1977.

[Howze,1973] J.W. Howze. Necessary and sufficient condition for decoupling using output feedback. IEEE Transactions on Automatic Control, **Vol. 18**: 44-46, 1973.

[Hussain, 2011] A. Hussain, N. Essounbouli, A. Hamzaoui, F. Nollet, and J. Zaytoon. Type-2 fuzzy sliding mode control without reaching phase for nonlinear system. Engineering Applications of Artificial Intelligence, **Vol. 24(1)**: 23–38, 2011.

[Isidori,1995] A. Isidori. Nonlinear Control Systems. Springer-Verlag, London, third edition, 1995.

[Jantzen, 2007] J. Jantzen. Foundations of Fuzzy Control. John Wiley & Sons Ltd, 2007.

[Kosko, 1994] B. Kosko. Fuzzy systems as universal approximators. IEEE transactions on computers, **Vol. 43(11)**: 1329–1333, 1994.

[Karnik, 1998a] N. N. Karnik and J. M. Mendel. An introduction to type-2 fuzzy logic systems. Technical report, Univ. Southen California, 1998.

[Karnik, 1998b] N.N. Karnik and J.M. Mendel. Type-2 fuzzy logic systems : type-reduction. In IEEE International Conference on Systems, Man, and Cybernetics, **Vol. 2**: 2046–2051, 1998.

[Karnik, 1999] N.N. Karnik, Jerry M. Mendel, and Q. Liang. Type-2 Fuzzy Logic Systems. IEEE Transactions on Fuzzy Systems, **Vol. 7(6)**: 643-658, 1999.

[Kessal, 2014] A. Kessal, and L. Rahmani. Ga-optimized parameters of sliding-mode controller based on both output voltage and input current with an application in the PFC of AC/DC converters. IEEE Transactions on Power Electronics, **Vol. 29(6)**: 3159-3165, 2014.

[Kessal, 2011] A. Kessal, L. Rahmani, J.P. Gaubert, and M. Mostefai. Analysis and design of an isolated single-phase power factor corrector with a fast regulation. Electric Power Systems Research, **Vol. 81**:1825-1831. 2011.

[Kessal, 2012] A. Kessal, L. Rahmani, J.P. Gaubert, M. Mostefai. Experimental design of a fuzzy controller for improving power factor of boost rectifi er. Int J Electron, **Vol. 99**(12):1611–21, 2012.

[Kothare, 1996] M . Kothare, V . Balakrishnan, and M . Morari. Robust constrained model predictive control using linear matrix inequalities. Automatica, Vol. 32(10): 1361–1379, 1996.

[Kunusch, 2009] C. Kunusch, P.F. Puleston, M. A. Mayosky, and J. Riera. Sliding Mode Strategy for PEM Fuel Cells Stacks Breathing Control Using a Super-Twisting Algorithm. IEEE Transactions on Control Systems Technology, **Vol. 17(1)**: 167-174, 2009.

[Kwon, 1977] W . H . Kwon, and A . E. Pearson. "A modified quadratic cost problem and feedback stabilization of a linear system. IEEE Trans ., **Vol. AC-22(5)**: 838–842, 1977.

[Lall, 1994] S. Lall, and K. Glover. A game theoritic approach to moving horizon control. Oxford University Press, 1994.

[Lee, 1967] E. B. Lee, and L. Markus . Foundations of optimal control theory. Wiley, 1967.

[Levant, 1985] L.V. Levantovsky. Second order sliding algorithms: their realization. Dynamics of Heterogeneous Systems, 32–43, 1985.

[Levantovsky, 1985] L.V. Levantovsky. Second order sliding algorithms: their realization. Dynamics of Heterogeneous Systems, Institute for System Studies, Moscow, **Vol. 20(2)**: 32–43, 1985. In Russian.

[Levant, 1993] A. Levant. Sliding order and sliding accuracy in sliding mode control. International Journal of Control, **Vol. 58(6)**:1247–1263, 1993.

[Levant, 1998] A. Levant. Robust exact differentiation via sliding mode technique. Automatica, **Vol. 34(3)**: 397–384, 1998.

[Levant, 2000] A. Levant, A. Pridor, R. Gitizadeh, I. Yaesh, and J.Z. Ben-Asher. Aircraft pitch control via second order sliding technique. Journal of Guidance, Control, and Dynamics, **Vol. 23(4)**: 586–594, 2000.

[Levant, 2003] A. Levant. Higher-order sliding modes, differentiation and output-feedback control. International Journal of Control, **Vol. 76(9–10)**: 924–941, 2003.
[Levant, 2005] A. Levant. Homogeneity approach to high-order sliding mode design. Automatica, **Vol. 41(5)**: 823–830, 2005.
[Lopez, 2000] P. Lopez and A.S. Nouri. Théorie Elementaire et Pratique de la Commande par les Régimes Glissants. Springer–Verlag Edition, 2000.
[Mamdani, 1974] E.H. Mamdani. Application of fuzzy algorithms for control of simple dynamic plant. Proc. IEE, **Vol. 121(12)**: 1585–1588, 1974.
[Lu, 1995] P. Lu. Optimal predictive control for continuous nonlinear systems. International Journal of Control, **Vol. 62(3)**: 633–649, 1995.
[Lu, 1998] X.Y. Lu, and S.K. Spurgeon. Asymptotic stabilization of multiple input nonlinear systems via sliding modes. Dynamics and Control, **Vol. 8**: 231-254, 1998.
[Magni, 1997] L. Magni, and L. Sepulchre. Stability margins of nonlinear receding-horizon control via inverse optimality. Systems and Control Letters, **Vol. 32**: 241–245, 1997.
[Magni, 2001] L. Magni, H . Nijmeijer, and A. J. van der Schaft. A receding horizon approach to the nonlinear H^{∞} control problem. Automatica , **Vol. 37(5)**: 429–435, 2001.
[Magni, 2011] L. Magni, D. M. Raimondo, and F. Allgöwer. Nonlinear Model Predictive Control. **Vol. 384**, Eds springer, 2011.
[Mayne, 1990] D. Q. Mayne, and H . Michalska. Receding horizon control of nonlinear systems. IEEE Transactions on Automatic Control, **Vol. 35(7)**: 814–824, 1990.
[Mayne, 2000] D.Q Mayne, J. B. Rawlings, C.V. Rao, and P.O.M. Scokaert. Constrained model predictive control: Stability and optimality. Automatica, **Vol. 36**: 789–814, 2000.
[Mamdani, 1976] E.H. Mamdani. Advances in the linguistic synthesis of fuzzy controllers. International Journal of Man-Machine Studies, **Vol. 8(6)**: 669–678, 1976.
[Mansouri, 2009] B. Mansouri, N. Manamanni, K. Guelton, A. Kruszewski, and TM Guerra. Output feedback LMI tracking control conditions with H criterion for uncertain and disturbed TS models. Information Sciences, **Vol. 179(4)**: 446–457, 2009.
[Manceur, 2012] M. Manceur, N. Essounbouli, and A. Hamzaoui. Second-order sliding fuzzy interval type-2 control for an uncertain system with real application. IEEE Transaction Fuzzy System, **Vol. 20(2)**: 262–75, 2012.
[Mendel, 2013] Jerry M. Mendel, and X. Liu. Simplified Interval Type-2 Fuzzy Logic Systems. IEEE Transactions on Fuzzy Systems, **Vol. 21(6)**: 1056-1069, 2013.
[Mendel, 2014] Jerry M. Mendel. General Type-2 Fuzzy Logic Systems Made Simple: A Tutorial. IEEE Transactions on fuzzy systems, **Vol. 22(5)**: 1162-1182, 2014.
[Michalska, 1993] H . Michalska, and D. Q. Mayne. Robust receding horizon control of constrained nonlinear systems. IEEE Transactions on Automatic Control, **Vol. 38(11)**: 1623–1633, 1993.
[Muske, 1993] K. R. Muske, and J. B. Rawlings. Model predictive control with linear models. A.I.CH.E Journal, **Vol. 39(2)**: 262–287, 1993.
[Perruquetti, 2002] W. Perruquetti and J.P. Barbot. Sliding mode control in engineering. Marcel Dekker Edition, 2002.
[Peterka, 1984] V . Peterka. Predictor based self-tuning control. Automatica , **Vol. 20**: 39–50, 1984.

[Pierre, 1995] D.A . Pierre, and J.W. Pierre. Digital controller design-alternative emulation approaches. ISA Transactions, **Vol. 34(3)**: 219–228, 1995.
[Propoi, 1963] A .I. Propoi. Use of LP methods for synthesizing sampled-data automatic systems". Automn Remote Control , **Vol. 24(7)**: 837–844, 1963.
[Prett, 1979] D. M . Prett, and R . D. Gillette. Optimization and constrained multivariable control of a catalytic cracking unit. AIChE National meeting, Houston , Texas ; also Proceedings joint Automatic Control Conference, San Francisco, California , 1979.
[Rattan, 1989] K. S. Rattan. Compensating for computational delay in digital equivalent of continuous control systems. IEEE Transactions on Automatic Control, **Vol. 34**: 895–899, 1989.
[Rawlings, 1993] J.B. Rawlings, and K.R. Muske. The stability of constrained receding control. IEEE Transactions on Automatic Control, **Vol. 38(10)**: 1512–1516, 1993.
[Rawlings, 1994] J. B. Rawlings, E.S. Meadows, and K. Muske. Nonlinear model predictive control: a tutorial and survey. In Proceedings of IFAC ADCHEM, Japan. 1994.
[Rodriguez, 2012] J. Rodriguez, and P. Cortes. Predictive control of power converters and electrical drives, John Wiley, 2012.
[Ronco, 1999] E. Ronco, T. Arsan, and P. J. Gawthrop. Open-loop intermittent feedback control: Practical continuous-time G PC. IEE Proceedings-Control Theory and Applications, **Vol. 146(5)**: 426–434, 1999.
[Richalet, 1976] J. Richalet, A . Rault, J. L. Testud, and J. Papon. Algorithmic control of industrial processes. Proceedings of the Fourth IFAC symposium on identification and system parameter estimation, 1119–1167, 1976.
[Richalet, 1978] J. Richalet, A. Rault, J.L. Testud, and J. Papon. Model predictive heuristic control: Application to industrial processes. Automatica, **Vol. 14**: 413–428, 1978.
[Scokaert, 1999] P.O.M . Scokaert, D.Q. Mayne, and J.B. Rawlings. Sub optimal model predictive control. IEEE Transactions on Automatic Control, **Vol. 44(3)**: 648–654, 1999.
[Slotine, 1983] J.J. Slotine, and S.S. Sastry. Tracking control of non-linear system using sliding surfaces with application to robot manipulators. International Journal of Control, **Vol. 38(2)**: 465-492, 1983.
[Slotine, 1986] J.J.E. Slotine, and J.A. Coestsee. Adaptive sliding controller synthesis for nonlinear systemes. International Journal of Control, **Vol. 43(6)**: 1631–1651, 1986.
[Slotine, 1991] J.J.E. Slotine, and W. Li. Applied nonlinear control. Prentice–Hall Edition, 1991.
[Slotine, 1993] J.J Slotine, and J.K. Hedrick. Robust input output feedback linearization. International Journal of Control, **Vol. 57(5)**: 1133-1139, 1993.
[Sira-Ramirez, 1988] H. Sira-Ramirez. Differential geometric methods in variable-structure
[Sira-Ramirez, 1992] H. Sira-Ramirez. On the sliding mode control of nonlinear systems. Systems and Control Letters, **Vol. 19**: 303-312, 1992.
control. International Journal of Control, **Vol. 48(4)**:1359–1390, 1988.
[Sira-Ramirez, 1993] H. Sira-Ramirez. On the dynamic sliding mode control of nonlinear systems. International Journal of Control, **Vol. 57(5)**: 1039-1061, 1993.
[Soroush, 1997] M. Soroush, and H.M . Soroush. Input-output linearising nonlinear model predictive control. International Journal of Control, **Vol. 68(6)**: 1449–1473, 1997.

[Sivanandam, 2007] S. N. Sivanandam, S. Sumathi, and S. N. Deepa. Introduction to fuzzy logic using matlab. Springer-Verlag, 2007.

[Sugeno, 1988] M. Sugeno, and G. Kang. Structure identification of fuzzy model . Fuzzy Sets and Systems, **Vol. 28(1):** 15–33, 1988.

[Takagi, 1985] T. Takagi, and M. Sugeno. Fuzzy identification of systems and its application to mo delling and control. IEEE Transactions on Systems Man, and Cybernetics, **Vol. 15(1)**: 116–132, 1985.

[Tanaka, 1998] K. Tanaka, T. Ikeda, and H.O. Wang. Fuzzy regulators and fuzzy observers: relaxed stability conditions and LMI-based designs. IEEE Transactions on Fuzzy Systems, **Vol. 6(2)**: 250–265, 1998.

[Tanaka, 2001] K. Tanaka, and H.O. Wang. Fuzzy control systems design and analysis: a linear matrix inequality approach. Wiley-Interscience, 2001.

[Tuffs, 1985] P.S. Tuffs, and D. W. Clarke. Self-tuning control of offset: a unified approach. IEE Proceedings , Part D-Control Theory and Applications , **Vol.132(3)**: 100–110, 1985.

[Tseng, 2001] C.S. Tseng, B.S. Chen, and H.J. Uang. Fuzzy tracking control design for nonlinear dynamic systems via TS fuzzy model. IEEE Transactions on Fuzzy Systems, **Vol. 9(3)**: 381–392, 2001.

[Utkin, 1976] V.I. Utkin. Sliding mode and their applications in variable structure systems. MIR Edition, 1976.

[Utkin, 1977] V.I.Utkin. Variable structure systems with sliding modes. IEEE Transactions on Automatic Control, **Vol. 22(2)**: 212–222, 1977.

[Utkin, 1992] V.I. Utkin. Sliding Mode minimization and Control Problems. Springer–Verlag Edition, 1992.

[Utkin, 1992] V.I. Utkin. Sliding modes in control and optimization. Springer-Verlag, 1992.

[Utkin, 1999] V.I. Utkin, J. Guldner, J. Shi, and M. Shijun. Sliding mode control in electromechanical systems. Taylor & Francis Edition, 1999.

[Qin, 2003] S. J. Qin, and T. A . Badgwell. A survey of industrial model predictive control technology ". Control Engineering Practice, **Vol. 11**: 733–764, 2003.

[Wang, 1994] L.X. Wang. Adaptive fuzzy systems and control- Design and stability analysis. Prentice Hall, 1994.

[Wang, 1996] H.O. Wang, K. Tanaka, and M.F. Griffin. An approach to fuzzy control of nonlinear systems : Stability and design issues. IEEE Transactions on Fuzzy Systems, **Vol. 4(1)**: 14–23, 1996.

[Wang, 2006] Z. Wang, Z. Chen, Q. Sun, and Z. Yuan. "Robust Stability Analysis of CG PC. Proceedings of the 6th World Congress on Intelligent Control and Automation, China, 324 - 328, 2006.

[Wissel, 1997] D. von Wissel, R . Nikoukhah, S. L. Campbell, and F . Delebecque. Effects of computational delay in descriptor-based trajectory tracking control. International Journal of Control, **Vol. 67(2)**: 251–273, 1997.

[Zadeh, 1965] L.A. Zadeh. Fuzzy sets. Information and control, Vol. 8(3): 338–353, 1965.

[Zeng, 2000] K. Zeng, N.Y. Zhang, and W.L. Xu. A comparative study on sufficient conditions for takagi-sugeno fuzzy systems as universal approximators. IEEE Transactions on Fuzzy Systems, **Vol. 8(6)**: 773–780, 2000.

Abstract- Robustness request of nonlinear system control has already become one of the main research interests for lot of researchers. As known to all, traditional controls based on local linearization for the nonlinear system has been proposed, but these methods have a limitation and small scale motion in the presence of uncertainties and output disturbance, which often limits bandwidth of closed loop, so that system tracking performance and robustness will be decreased, for above reasons, this thesis present an attempt to study deeply the unconventional controls and their applications for nonlinear system. We have used three unconventional controls based on model predictive control, high order sliding mode control and fuzzy logic controller. These kinds of robust control strategies applied for improving the performance of power factor correction. All the proposed methods are verified through simulation on the MATLAB/Simulink platform; the results show good performances in both steady and transient states. To show the applicability of the proposed methods in industrial sectors an experiment is conducted through a test bench based on dSPACE 1104. The experimental results proved that the proposed controllers enhanced the performance of the system greatly under different parameters variations.

Keywords: unconventional control, Nonlinear system, Model predictive control, High order sliding mode control, Fuzzy logic controller, Power factor correction, MATLAB, dSPACE 1104.

Résumé- Le besoin de la robustesse de commande du système non linéaire est déjà devenu l'un des primordiaux intérêts de la recherche pour beaucoup des chercheurs. Comme connu de tous, les commandes traditionnels basés sur la linéarisation locale pour le système non linéaire a été proposé, mais ces méthodes ont une limitation et un petit mouvement à grande échelle dans la présence d'incertitudes et des perturbations, ce qui limite souvent la bande passante de la boucle fermée, de sorte que le suivi de la performance et de la robustesse du système seront réduites, pour ces raisons, cette thèse présente une tentative d'étudier profondément les commandes non conventionnelles et de leurs applications pour le système non linéaire. Nous avons utilisé trois commandes non conventionnelles basés sur la commande prédictive, commande par mode glissent d'ordre supérieur et commande par logique floue. Ces types de stratégies de commande robustes appliquées pour améliorer la performance du correcteur de facteur de puissance. Toutes les méthodes proposées sont validées par simulation sur la plate-forme MATLAB/Simulink; les résultats montrent de bonnes performances dans les deux états dynamique et transitoire. Pour montrer l'applicabilité des méthodes proposées dans les secteurs industriels une expérience est menée à travers un banc d'essai expérimental basé sur la dSPACE 1104. Les résultats expérimentaux ont prouvé que les contrôleurs proposés ont largement améliorés la performance du système sous différentes variations des paramètres.

Mots-clés: Commande non conventionnelle, système non linéaire, commande prédictive, commande par mode glissent d'ordre supérieur, commande par logique floue, correction du facteur de puissance, MATLAB, dSPACE 1104.

ملخص : الحاجة إلى متانة نظام التحكم للأنظمة غير الخطيّ أصبحت بالفعل واحدة من البحوث المهمة لكثير من الباحثين. كما هو معروف للجميع، هناك نظام التحكم التقليدي القائم على الخطية المحلية للنظام غير الخطي، ولكن هذه الأساليب لها قيود وحركة صغيرة الحجم في وجود حالة عدم اليقين والاضطراب، والتي غالبا ما تحد من عرض النطاق الترددي للحلقة المغلقة، بحيث يتم تخفيض رصد أداء ومتانة النظام، لهذه الأسباب، هده الأطروحة تعرض محاولة لاستكشاف الضوابط غير تقليدية وتطبيقاتها للنظام غير الخطي . لقد قمنا باستعمال ثلاثة عناصر تحكم غير تقليدية تقوم على السيطرة التنبؤية، المبدأ ألانزلاقي ذا الوضع العالي و المنطق الضبابي. هذه الاستراتيجيات استعملت لتحسين معالج معامل القدرة. يتم التحقق من صحة كل الأساليب المقترحة عن طريق المحاكاة بالاعتماد علي برنامج المحاكاة BALTAM/Smiknilu. نتائج المحاكاة كانت جيدة في الحالتين الدائمة و العابرة . للتحقق من عمل الأساليب المقترحة في المجال الصناعي قمنا باختبار تجريبي اعتمادا علي بطاقة التحكم1104 ECAPSd. أظهرت النتائج التجريبية أن النظم المقترحة قد حسنت بشكل كبير أداء النظام مهما تغيرت المعاملات.

كلمات مفتاحيه: تحكم غير تقليدي ، الأنظمة غير الخطيّ، السيطرة التنبؤية، المبدأ ألانزلاقي ذا الوضع العالي ، المنطق الضبابي، BALTAM/Smiknilu، معالج معامل القدرة، ECAPSd 1104